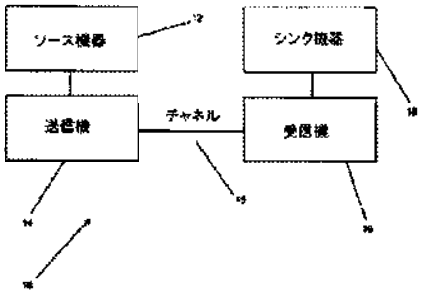
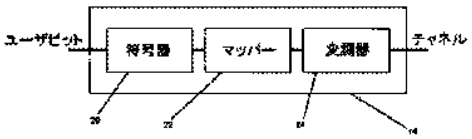
【００５２】
反対に、受信機が、最適化されたコンステレーション／符号レートの共調節機構に基づいて性能品質を実現することができる。かかるアプローチは、受信機品質測定値のある形式も含むことができる。これは、受信機のＳＮＲ又はビットエラーレートの推定値でもよい。例えば、ビットエラーレートがある許容閾値を超えた場合を例にとって説明する。この場合、バックチャネルを介して、受信機が送信機にリンクのスペクトル効率を下げるように要求することができる。これは送信機が、符号器モジュール及びマッパーモジュールにおいて、代替の容量最適化されたコンステレーション／符号レートの組にスワッピングして、続いて受信機に、デマッパーモジュール／復号器モジュールの補完的な組み合わせにスワップするように信号送信することによって行うことができる。
＜幾何学的形状のＱＡＭコンステレーション＞
【００５３】
ＱＡＭ（Ｑｕａｄｒａｔｕｒｅ ａｍｐｌｉｔｕｄｅ ｍｏｄｕｌａｔｉｏｎ：直交振幅変調）コンステレーションは、ＰＡＭコンステレーションをＱＡＭ同相成分及び直交成分に直交化することによって構築される。このようにして構築されたコンステレーションは、低複雑度デマッパーを有するため、多くの用途で有用となりうる。
【００５４】
図２１に、パルス振幅変調コンステレーションから構築した直交振幅変調コンステレーションの例を示す。図示の実施の形態は、１．５ビットの次元（９．０ｄＢのＳＮＲに相当）（図１３ｂ参照）ごとのユーザビットレートでＰＤ容量について最適化されたＰＡＭ－８コンステレーションを用いて構築された。このＰＡＭ－８コンステレーションにおけるラベルとポイントの組は、｛（０００，－１．７２），（００１，－０．６１），（０１０，１．７２），（０１１，－０．６２），（１００，０．６２），（１０１，０．０２），（１１０，０．８１），（１１１，－０．０２）｝である。図２１を検討すると、ＱＡＭコンステレーションの構築は、同相次元の８個のＰＡＭ－８ポイントの各々について、直交次元のＰＡＭ－８ポイントの完全な群を複製することによって、達成されることがわかる。ラベリングは、ＰＡＭ－８ラベルを、同相次元のＬＳＢ範囲と直交次元のＭＳＢ範囲に割り当てることによって達成される。得られた８ｘ８の外積は、極めて低複雑度のデマッパーを構築することができる高度に構造化されたＱＡＭ－６４を形成する。同相成分と直交成分の直交性のため、得られたＱＡＭ－６４コンステレーションの容量特性は、次元当たりのＰＡＭ－８コンステレーションの容量特性と同一となる。
＜Ｎ次元のコンステレーションの最適化＞
【００５５】
１Ｄ（例えば、ＰＡＭ）のコンステレーションを設計し、次に２－Ｄ（ＱＡＭ）に拡張するよりも、追加的な空間次元により表される追加の自由度を最適化するステップを直接的に利用することもできる。一般的に、Ｎ次元コンステレーションと、関連するラベリングを設計することができる。最適化ステップの複雑度は、次元の数とともに指数関数的に増加し、これによって生成される受信機デマッパーの複雑度も同様に増加する。この構築も、本発明の実施の形態を構成し、生成するためにより多くの「ランタイム」を必要とする。
＜フェージングチャネルのための容量最適化コンステレーション＞
【００５６】
本発明の実施の形態によると、上記に概略したものと同様の処理を用いて、フェージングチャネルのための容量最適化されたコンステレーションを設計することができる。この処理は、容量を算出する手法が、フェージングチャネルを考慮して変更されている点を除くと、基本的に同様である。フェージングチャネルは、以下の式を用いて表すことができる。
Ｙ＝ａ（ｔ），Ｘ＋Ｎ
ここで、Ｘは送信された信号、Ｎは加法性ホワイトガウスノイズ信号、ａ（ｔ）は、時間の関数であるフェージング分布である。

【００５７】
フェージングチャネルの場合、受信機の瞬間的なＳＮＲは、フェージング分布に従って変化する。フェージング分布がレイリーであり、システムの平均ＳＮＲはＡＷＧＮチャネルの場合と同様のまま、Ｅ［Ｘ^2］／Ｅ［Ｎ^2］という特性を有する。従って、フェージングチャネルの容量は、瞬間的なＳＮＲの分布を導くレイリーフェージング分布のａにわたって、所与の平均ＳＮＲで、ＡＷＧＮ容量の予想値を用いて算出することができる。
【００５８】
多くのフェージングチャネルはレイリー分布に従う。図２２ａから図２４ｂは、レイリーフェージングチャネルでＰＤ容量について最適化されたＰＡＭ－４，８，及び１６コンステレーションの軌跡図である。ユーザビットレート／次元に対する軌跡図と、ＳＮＲに対する軌跡図を示す。同様の処理を用いて、共通容量などのその他の容量測定値を用いて、及び／又は異なる変調方式を用いて最適化された容量最適化されたコンステレーションを取得することができる。

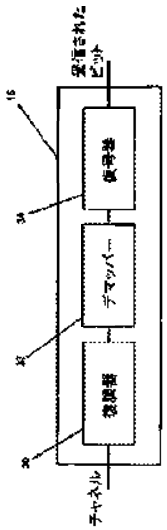
【図１】



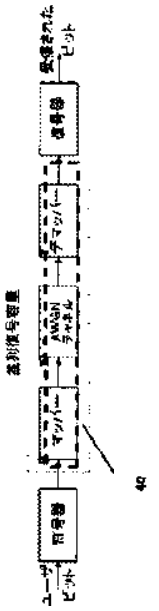
【図２】



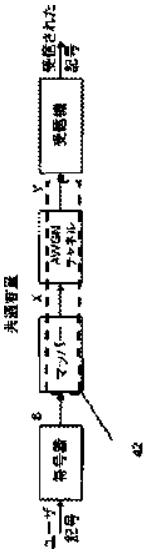
【図３】



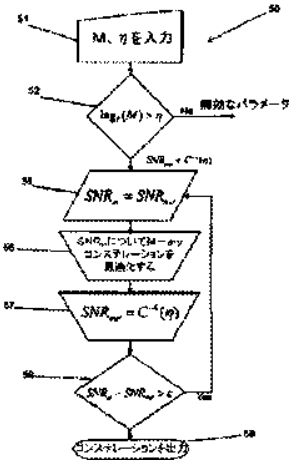
【図４ａ】



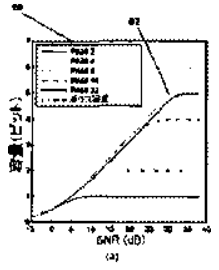
【図４ｂ】



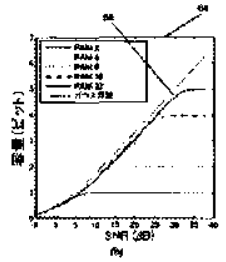
【図５】



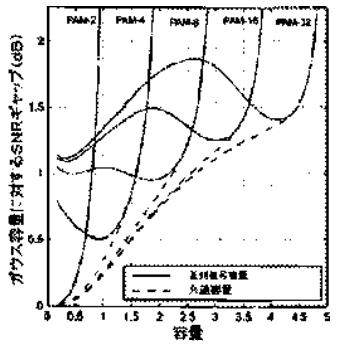
【図６ａ】



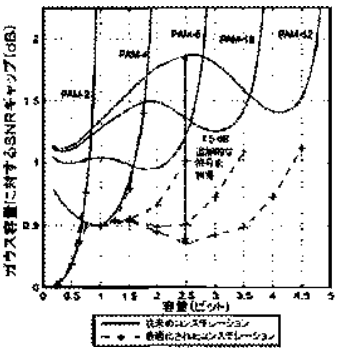
【図６ｂ】



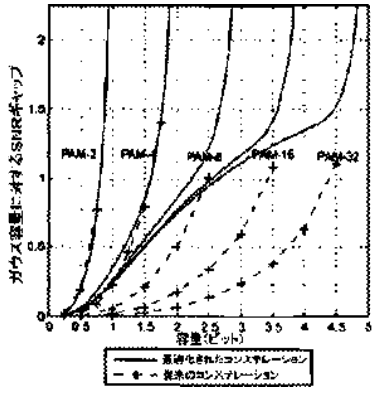
【図７】


PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
容量

【図８ａ】


PAM-2
PAM-4
PAM-8
PAM-16
PAM-32

【図８ｂ】


PAM-2
PAM-4
PAM-8
PAM-16
PAM-32

【図９】

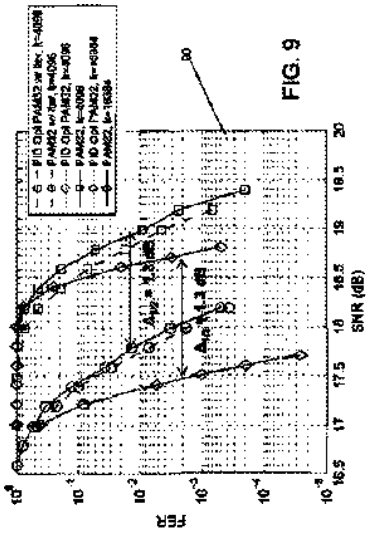
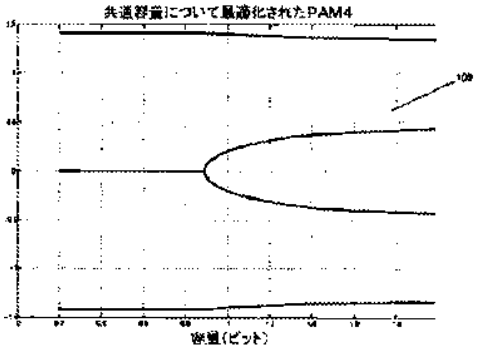
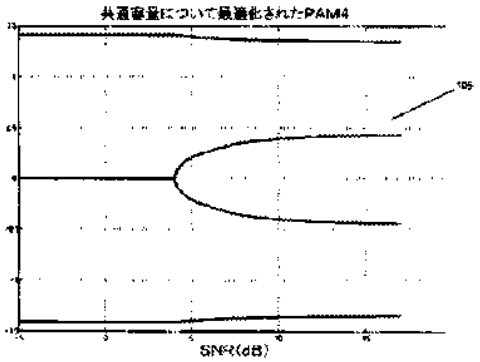


FIG. 9
SNR (dB)
FER

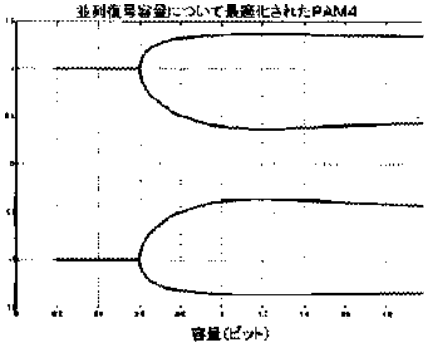
【図１０ａ】



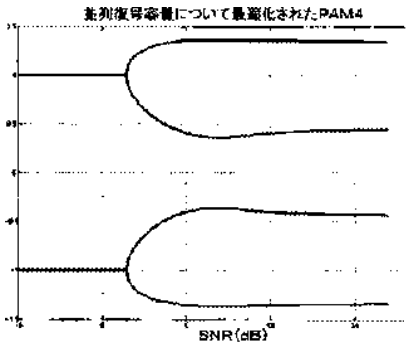
【図１０ｂ】



【図１０ｃ】



【図１０ｄ】



【図１１ａ】

異なるレートで共通容量について最適化されたPAM－4コンステレーション

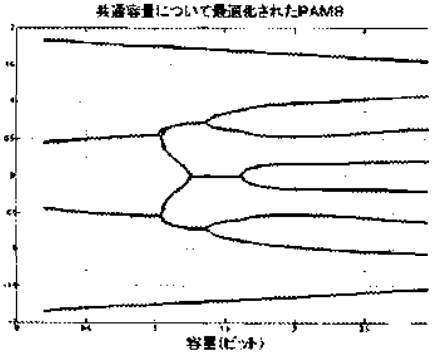
(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.03	2.71	5.00	7.15	9.24
x_3	-1.41	-1.41	-1.40	-1.37	-1.36
x_1	0.00	0.00	-0.20	-0.33	-0.39
x_2	0.00	0.00	0.20	0.33	0.39
x_3	1.41	1.41	1.40	1.37	1.36

【図１１ｂ】

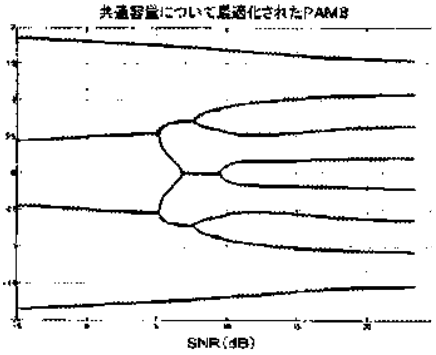
異なるレートで並列復号容量について最適化されたPAM－4コンステレーション

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.19	3.11	5.26	7.22	9.25
x_0	-1.00	-1.30	-1.36	-1.37	-1.36
x_1	-1.00	-0.56	-0.39	-0.33	-0.39
x_2	1.00	1.30	1.36	0.33	1.36
x_3	1.00	0.56	0.39	1.37	0.39

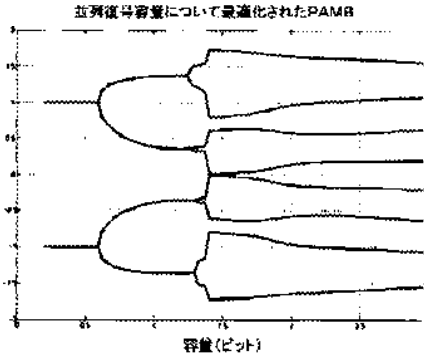
【図１２ａ】



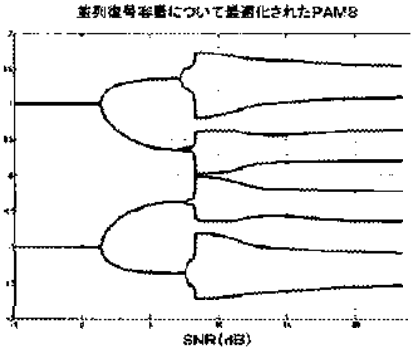
【図１２ｂ】



【図１２ｃ】



【図１２ｄ】



【図１３ａ】

異なるレートで共通容量について最適化されたPAM－8コンステレーション

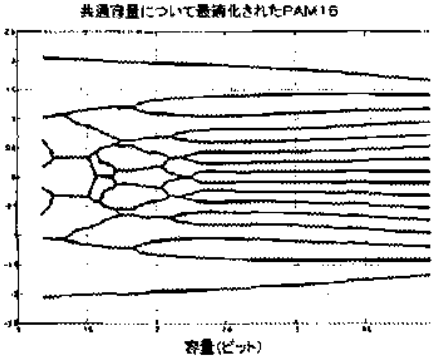
(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.00	4.82	8.66	12.26	15.93
x_0	-1.81	-1.76	-1.70	-1.66	-1.60
x_1	-0.50	-0.55	-0.64	-0.97	-1.03
x_2	-0.50	-0.55	-0.63	-0.53	-0.58
x_3	-0.50	-0.55	-0.60	-0.17	-0.19
x_4	0.50	0.55	0.00	0.17	0.19
x_5	0.50	0.55	0.63	0.53	0.58
x_6	0.50	0.55	0.84	0.97	1.03
x_7	1.81	1.76	1.70	1.66	1.60

【図１３ｂ】

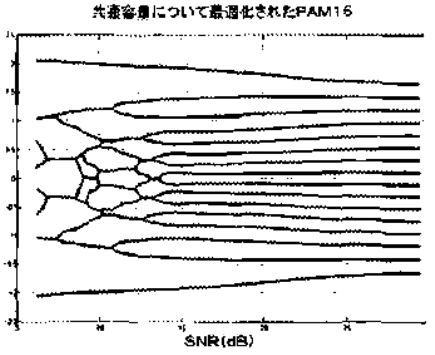
異なるレートで並列復号容量について最適化されたPAM−8コンステレーション

(bpc)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.19	5.27	9.00	12.42	15.93
x_0	-1.00	-1.36	-1.72	-1.64	-1.60
x_1	-1.00	-1.36	-0.81	-0.93	-1.03
x_2	-1.00	-0.39	1.72	1.64	-0.19
x_3	-1.00	-0.39	-0.62	-0.58	-0.58
x_4	1.00	1.36	0.62	0.58	1.60
x_5	1.00	1.36	0.02	0.35	1.03
x_6	1.00	0.39	0.81	0.97	0.19
x_7	1.00	0.39	-0.02	-0.15	0.58

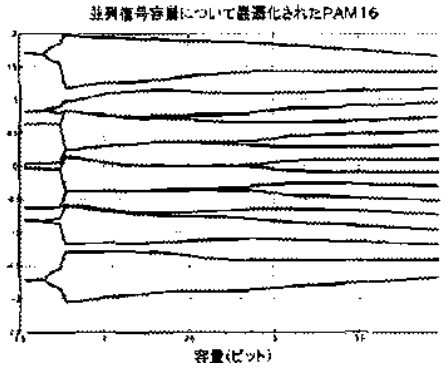
【図１４ａ】



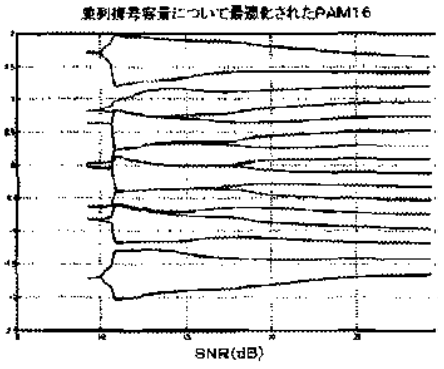
【図１４ｂ】



【図１４ｃ】



【図１４ｄ】



【図１５ａ】

異なるレートで共通容量について最適化されたPAM－16コンステレーション

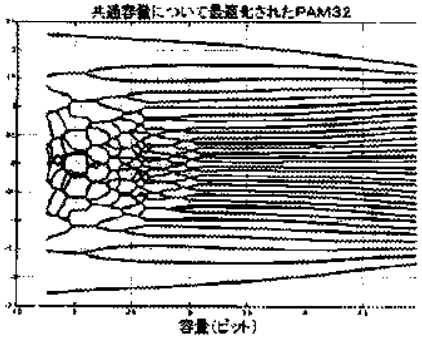
(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	9.52	11.94	13.25	16.00	22.12
x_0	-2.08	-1.96	-1.91	-1.85	-1.76
x_1	-1.16	-1.33	-1.49	-1.42	-1.42
x_2	-1.16	-1.80	-1.05	-1.15	-1.15
x_3	-0.90	-0.89	-0.82	-0.84	-0.90
x_4	-0.74	-0.69	-0.68	-0.62	-0.68
x_5	-0.84	-0.45	-0.45	-0.41	-0.47
x_6	-0.34	-0.17	-0.24	-0.26	-0.28
x_7	-0.34	-0.07	-0.09	-0.08	-0.09
x_8	0.34	0.07	0.09	0.08	0.09
x_9	0.84	0.47	0.24	0.26	0.28
x_{10}	0.84	0.45	0.45	0.41	0.47
x_{11}	0.84	0.89	0.69	0.62	0.68
x_{12}	0.90	0.89	0.82	0.84	0.90
x_{13}	1.16	1.30	1.05	1.10	1.15
x_{14}	1.16	1.33	1.40	1.42	1.42
x_{15}	2.08	1.96	1.91	1.85	1.76

【図１５ｂ】

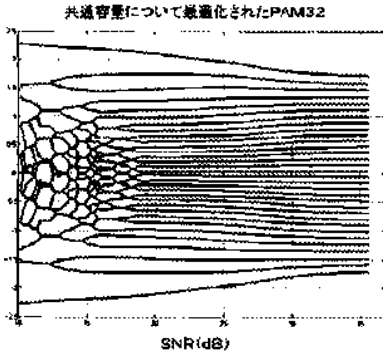
異なるレートで全則信号容量について最適化されたPAM－16コンステレーション

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	9.80	12.25	15.42	18.72	22.58
x_0	-1.72	-1.98	-1.89	-1.84	-1.75
x_1	-1.12	-1.29	-1.38	-1.42	-1.42
x_2	-0.81	-1.04	-1.09	-1.14	-1.15
x_3	-0.81	-1.17	-1.14	-1.16	-1.15
x_4	-1.72	-0.38	-0.33	-0.46	-0.47
x_5	-1.72	-0.65	-0.70	-0.65	-0.68
x_6	-0.62	-0.38	-0.34	-0.20	-0.28
x_7	-0.62	-0.49	-0.76	-0.83	-0.90
x_8	0.62	1.08	1.13	1.16	1.15
x_9	0.62	0.76	0.76	0.84	0.90
x_{10}	0.82	1.24	1.35	1.42	1.42
x_{11}	0.82	0.76	0.70	0.64	0.68
x_{12}	0.81	0.06	0.08	0.05	0.08
x_{13}	0.81	0.29	0.34	0.29	0.28
x_{14}	-0.82	0.06	0.08	-0.05	-0.09
x_{15}	-0.82	0.29	0.35	0.40	0.47

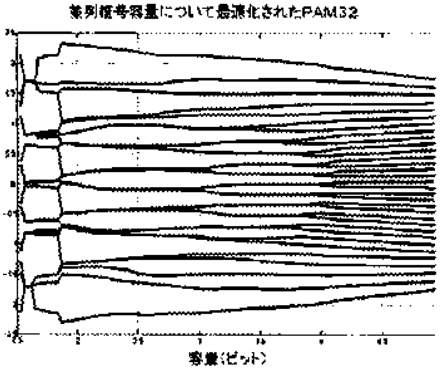
【図１６ａ】



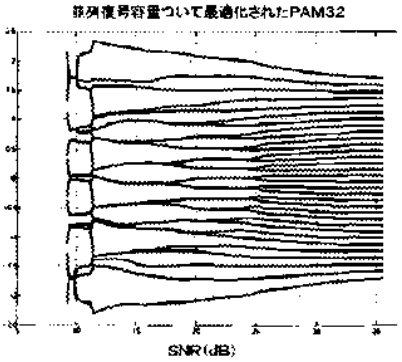
【図１６ｂ】



【図１６ｃ】

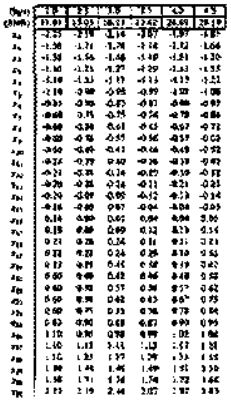


【図１６ｄ】



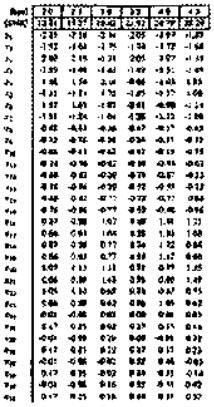
【図１７ａ】

異なるレートで共通容量について最適化されたPAM−32コンステレーション

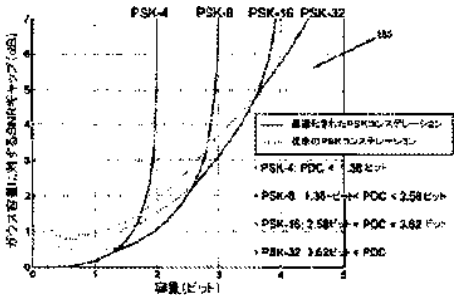


【図１７ｂ】

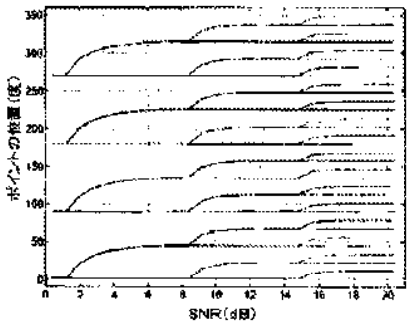
異なるレートで並列復号容量について最適化されたPAM−32コンステレーション



【図１８】



【図１９】



【図２０】

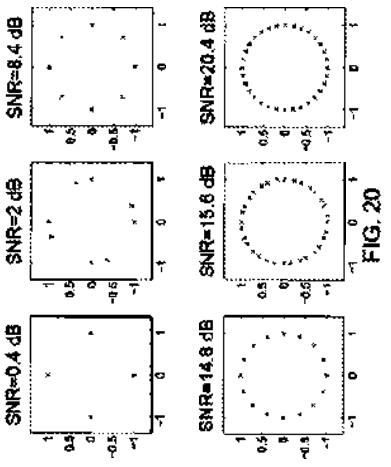
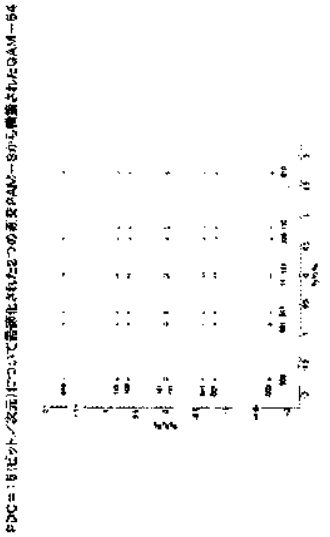
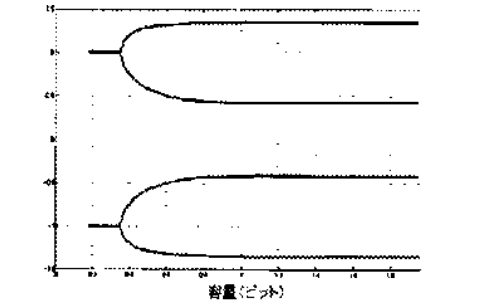


FIG. 20

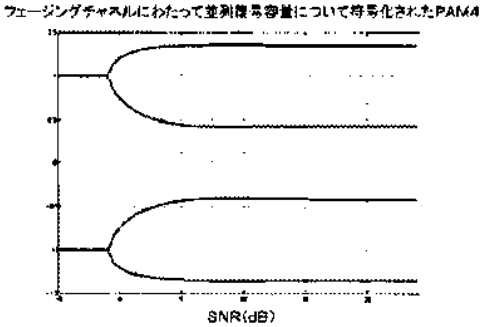
【図２１】



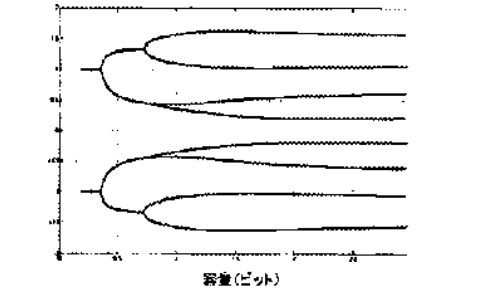
【図２２ａ】



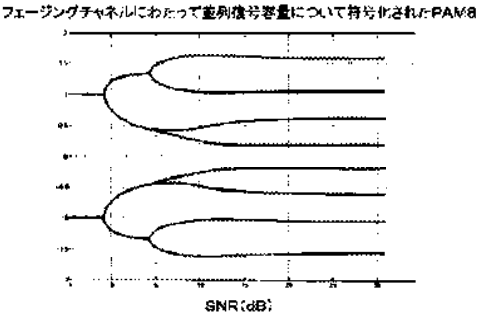
【図２２ｂ】



【図２３ａ】

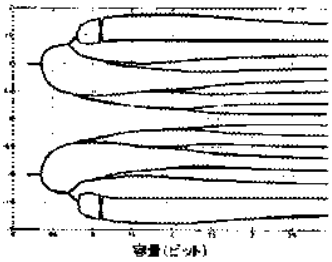


【図２３ｂ】



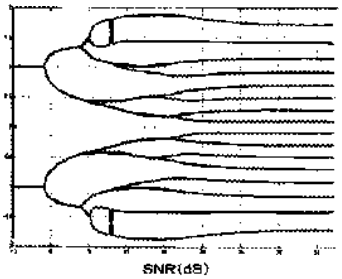
【図２４ａ】

フェージングチャネルにわたって並列復号容量について符号化されたPAM16



【図２４ｂ】

フェージングチャネルにわたって並列復号容量について符号化されたPAM16



フロントページの続き

(72)発明者 パーソウム マジュド エフ.
アメリカ合衆国 カリフォルニア州 アーバイン ドゥランゾ アイル 374
(72)発明者 ジョーンズ クリストファー リチャード
アメリカ合衆国 カリフォルニア州 パシフィック パリサデス カレ パトリシア 1563

審査官 彦田 克文

(56)参考文献 特開2005-269258 (JP, A)
国際公開第2007/074524 (WO, A1)
特開平09-130438 (JP, A)
特開2003-229835 (JP, A)

(58)調査した分野(Int.Cl., DB名)
H04L 27/34

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 2 153 561 B1

(12) **EUROPEAN PATENT SPECIFICATION**

(45) Date of publication and mention of the grant of the patent:
09.01.2019 Bulletin 2019/02

(21) Application number: **08795885.6**

(22) Date of filing: **05.06.2008**

(51) Int Cl.:
H04J 11/00 *(2006.01)* ***H04L 27/34*** *(2006.01)*
H04L 1/00 *(2006.01)*

(86) International application number:
PCT/US2008/065994

(87) International publication number:
WO 2008/151308 (11.12.2008 Gazette 2008/50)

(54) **METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS**

VERFAHREN UND VORRICHTUNG ZUR SIGNALISIERUNG MIT KAPAZITÄTSOPTIMIERTEN KONSTELLATIONEN

PROCÉDÉ ET APPAREIL DE SIGNALISATION DE CONSTELLATIONS OPTIMISANT LA CAPACITÉ

(84) Designated Contracting States:
AT BE BG CH CY CZ DE DK EE ES FI FR GB GR HR HU IE IS IT LI LT LU LV MC MT NL NO PL PT RO SE SI SK TR

(30) Priority: **05.06.2007 US 933319 P**

(43) Date of publication of application:
17.02.2010 Bulletin 2010/07

(73) Proprietor: **Constellation Designs, Inc.**
Pacific Palisades, CA 90272 (US)

(72) Inventors:
- **Barsoum, Maged F.**
San Jose, CA 95120 (US)
- **Jones, Christopher Richard**
Pacific Palisades, CA 90272 (US)

(74) Representative: **Derry, Paul Stefan**
Venner Shipley LLP
200 Aldersgate
London EC1A 4HD (GB)

(56) References cited:
US-A1- 2004 022 179
US-A1- 2006 155 843
US-A1- 2007 116 161
US-B1- 6 603 801

- **RICCARDO DE GAUDENZI ET AL: "Performance analysis of turbo-coded APSK modulations over Nonlinear Satellite Channels", IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS, IEEE SERVICE CENTER, PISCATAWAY, NJ, US, vol. 5, no. 9, 6 September 2006 (2006-09-06), pages 2396-2407, XP002457811, ISSN: 1536-1276, DOI: 10.1109/TWC.2006.1687763**
- **DIRK SOMMER ET AL: "Signal Shaping by Non-Uniform QAM for AWGN Channels and Applications Using Turbo Coding", PROC. ITG CONF. SOURCE AND CHANNEL CODING, 31 January 2000 (2000-01-31), pages 81-86, XP055110228,**
- **RICCARDO DE GAUDENZI ET AL: "Turbo-coded APSK modulations design for satellite broadband communications", INTERNATIONAL JOURNAL OF SATELLITE COMMUNICATIONS AND NETWORKING, JOHN WILEY & SONS, CHICHESTER, GB, vol. 24, no. 4, 1 July 2006 (2006-07-01), pages 261-281, XP008134566, ISSN: 1542-0973, DOI: 10.1002/SAT.841 [retrieved on 2006-05-19]**
- **FEI ZESONG ET AL: "Shaping gain by non-uniform qam constellation i with binary turbo coded modulation", PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS, 2003. PIMRC 2003. 14TH IEEE PROCEEDINGS ON SEPT. 7-10, 2003, IEEE, PISCATAWAY, NJ, USA, vol. 2, 7 September 2003 (2003-09-07), pages 1863-1867, XP010678797, ISBN: 978-0-7803-7822-3**

Note: Within nine months of the publication of the mention of the grant of the European patent in the European Patent Bulletin, any person may give notice to the European Patent Office of opposition to that patent, in accordance with the Implementing Regulations. Notice of opposition shall not be deemed to have been filed until the opposition fee has been paid. (Art. 99(1) European Patent Convention).

Printed by Jouve, 75001 PARIS (FR)

(Cont. next page)

EP 2 153 561 B1

- MAGED F BARSOUM ET AL: "Constellation Design via Capacity Maximization", INFORMATION THEORY, 2007. ISIT 2007. IEEE INTERNATIONAL SYMPOSIUM ON, IEEE, PISCATAWAY, NJ, USA, 24 June 2007 (2007-06-24), pages 1821-1825, XP031440978, ISBN: 978-1-4244-1397-3

Description

BACKGROUND

[0001] The present invention generally relates to bandwidth and/or power efficient digital transmission systems and more specifically to the use of unequally spaced constellations having increased capacity.

[0002] The term "constellation" is used to describe the possible symbols that can be transmitted by a typical digital communication system. A receiver attempts to detect the symbols that were transmitted by mapping a received signal to the constellation. The minimum distance (dmin) between constellation points is indicative of the capacity of a constellation at high signal-to-noise ratios (SNRs). Therefore, constellations used in many communication systems are designed to maximize d_{min}. Increasing the dimensionality of a constellation allows larger minimum distance for constant constellation energy per dimension. Therefore, a number of multi-dimensional constellations with good minimum distance properties have been designed.

[0003] Communication systems have a theoretical maximum capacity, which is known as the Shannon limit. Many communication systems attempt to use codes to increase the capacity of a communication channel. Significant coding gains have been achieved using coding techniques such as turbo codes and LDPC codes. The coding gains achievable using any coding technique are limited by the constellation of the communication system. The Shannon limit can be thought of as being based upon a theoretical constellation known as a Gaussian distribution, which is an infinite constellation where symbols at the center of the constellation are transmitted more frequently than symbols at the edge of the constellation. Practical constellations are finite and transmit symbols with equal likelihoods, and therefore have capacities that are less than the Gaussian capacity. The capacity of a constellation is thought to represent a limit on the gains that can be achieved using coding when using that constellation..

[0004] Prior attempts have been made to develop unequally spaced constellations. For example, a system has been proposed that uses unequally spaced constellations that are optimized to minimize the error rate of an uncoded system. Another proposed system uses a constellation with equiprobable but unequally spaced symbols in an attempts to mimic a Gaussian distribution.

[0005] Other approaches increases the dimensionality of a constellation or select a new symbol to be transmitted taking into consideration previously transmitted symbols. However, these constellation were still designed based on a minimum distance criteria..

[0006] The use of non-uniform QAM constellation for capacity improvement has been disclosed in the following document: D. Sommer and G. P. Fettweis, "Signal shaping by non-uniform QAM for AWGN channels and applications using turbo coding", Proc. 2000 ITG Conf. Source and Channel Coding, pp. 81-86. The optimizing of APSK constellations has been disclosed in the following document: R. De Gaudenzi, A. Guillen i Fabregas and A. Martinez, "Performance analysis of turbo-coded APSK modulations over nonlinear satellite channels," in IEEE Transactions on Wireless Communications, vol. 5, no. 9, pp. 2396-2407, September 2006.

SUMMARY OF THE INVENTION

[0007] The invention is defined and limited by the scope of appended claims 1-17. In the following description, any embodiment(s) referred to and not falling within the scope of the appended claims, is (are) merely example(s) useful to the understanding of the invention.

[0008] Systems and methods are described for constructing a modulation such that the constrained capacity between a transmitter and a receiver approaches the Gaussian channel capacity limit first described by Shannon [ref Shannon 1948]. Traditional communications systems employ modulations that leave a significant gap to Shannon Gaussian capacity. The modulations of the present invention reduce, and in some cases, nearly eliminate this gap. The invention does not require specially designed coding mechanisms that tend to transmit some points of a modulation more frequently than others but rather provides a method for locating points (in a one or multiple dimensional space) in order to maximize capacity between the input and output of a bit or symbol mapper and demapper respectively. Practical application of the method allows systems to transmit data at a given rate for less power or to transmit data at a higher rate for the same amount of power.

[0009] One embodiment of the invention includes a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter, includes a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate, a mapper configured to map encoded bits to symbols in a symbol constellation, a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper. In addition, the receiver includes a demodulator configured to demodulate the received signal via the communication channel, a demapper configured to estimate likelihoods from the demodulated signal, a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper. Furthermore, the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

[0010] A further embodiment of the invention includes encoding the bits of user information using a coding scheme, mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced

signal-to-noise ratio compared to a signal constellation that maximizes d_{min}, modulating the symbols in accordance with a modulation scheme, transmitting the modulated signal via the communication channel, receiving a modulated signal, demodulating the modulated signal in accordance with the modulation scheme, demapping the demodulated signal using the geometrically shaped signal constellation to produce likelihoods, and decoding the likelihoods to obtain an estimate of the decoded bits.

[0011] Another embodiment of the invention includes selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0012] A still further embodiment of the invention includes selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0013] Still another embodiment of the invention includes selecting an appropriate constellation size and a desired SNR, and optimizing the location of the points of the constellation to maximize a capacity measure of the constellation.

[0014] A yet further embodiment of the invention includes obtaining a geometrically shaped PAM constellation with a constellation size that is the square root of said given constellation size, where the geometrically shaped PAM constellation has a capacity greater than that of a PAM constellation that maximizes d_{min}, creating an orthogonalized PAM constellation using the geometrically shaped PAM constellation, and combining the geometrically shaped PAM constellation and the orthogonalized PAM constellation to produce a geometrically shaped QAM constellation.

[0015] Another further embodiment of the invention includes transmitting information over a channel using a geometrically shaped symbol constellation, and modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate.

BRIEF DESCRIPTION OF DRAWINGS

[0016]

FIG. 1 is a conceptual illustration of a communication system in accordance with an embodiment of the invention.

FIG. 2 is a conceptual illustration of a transmitter in accordance with an embodiment of the invention.

FIG. 3 is a conceptual illustration of a receiver in accordance with an embodiment of the invention.

FIG. 4a is a conceptual illustration of the joint capacity of a channel.

FIG. 4b is a conceptual illustration of the parallel decoding capacity of a channel.

FIG. 5 is a flow chart showing a process for obtaining a constellation optimized for capacity for use in a communication system having a fixed code rate and modulation scheme in accordance with an embodiment of the invention.

FIG. 6a is a chart showing a comparison of Gaussian capacity and PD capacity for traditional PAM-2,4,8,16,32.

FIG. 6b is a chart showing a comparison between Gaussian capacity and joint capacity for traditional PAM-2,4,8,16,32.

FIG. 7 is a chart showing the SNR gap to Gaussian capacity for the PD capacity and joint capacity of traditional PAM-2,4,8,16,32 constellations.

FIG. 8a is a chart comparing the SNR gap to Gaussian capacity of the PD capacity for traditional and optimized PAM-2,4,8,16,32 constellations.

FIG. 8b is a chart comparing the SNR gap to Gaussian capacity of the joint capacity for traditional and optimized PAM-2,4,8,16,32 constellations.

FIG. 9 is a chart showing Frame Error Rate performance of traditional and PD capacity optimized PAM-32 constellations in simulations involving several different length LDPC codes.

FIGS. 10a - 10d are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 11a and 11b are design tables of PD capacity and joint capacity optimized PAM-4 constellations in accordance with embodiments of the invention.

FIGS. 12a - 12d are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 13a and 13b are design tables of PD capacity and joint capacity optimized PAM-8 constellations in

accordance with embodiments of the invention.

FIGS. 14a - 14d are locus plots showing the location of constellation points of a PAM- 16 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 15a and 15b are design tables of PD capacity and joint capacity optimized PAM-16 constellations in accordance with embodiments of the invention.

FIGS. 16a - 16d are locus plots showing the location of constellation points of a PAM-32 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 17a and 17b are design tables of PD capacity and joint capacity optimized PAM-32 constellations in accordance with embodiments of the invention.

FIG. 18 is a chart showing the SNR gap to Gaussian capacity for traditional and capacity optimized PSK constellations.

FIG. 19 is a chart showing the location of constellation points of PD capacity optimized PSK-32 constellations.

FIG. 20 is a series of PSK-32 constellations optimized for PD capacity at different SNRs.

FIG. 21 illustrates a QAM-64 constructed from orthogonal Cartesian product of two PD optimized PAM-8 constellations in accordance with an embodiment of the invention.

FIGS. 22a and 22b are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

FIGS. 23a and 23b are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

FIGS. 24a and 24b are locus plots showing the location of constellation points of a PAM-16 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

DETAILED DESCRIPTION OF THE INVENTION

[0017] Turning now to the drawings, communication systems in accordance with embodiments of the invention are described that use signal constellations, which have unequally spaced (i.e. 'geometrically' shaped) points. In several embodiments, the locations of geometrically shaped points are designed to provide a given capacity measure at a reduced signal-to-noise ratio (SNR) compared to the SNR required by.a constellation that maximizes d_{min}. In many embodiments, the constellations are selected to provide increased capacity at a predetermined range of channel signal-to-noise ratios (SNR). Capacity measures that can be used in the selection of the location of constellation points include, but are not limited to, parallel decode (PD) capacity and joint capacity.

[0018] In many embodiments, the communication systems utilize capacity approaching codes including, but not limited to, LDPC and Turbo codes. As is discussed further below, direct optimization of the constellation points of a communication system utilizing a capacity approaching channel code, can yield different constellations depending on the SNR for which they are optimized. Therefore, the same constellation is unlikely to achieve the same coding gains applied across all code rates; that is, the same constellation will not enable the best possible performance across all rates. In many instances, a constellation at one code rate can achieve gains that cannot be achieved at another code rate. Processes for selecting capacity optimized constellations to achieve increased coding gains based upon a specific coding rate in accordance with embodiments of the invention are described below. In a number of embodiments, the communication systems can adapt location of points in a constellation in response to channel conditions, changes in code rate and/or to change the target user data rate.

COMMUNICATION SYSTEMS

[0019] A communication system in accordance with an embodiment of the invention is shown in FIG. 1. The communication system 10 includes a source 12 that provides user bits to a transmitter 14. The transmitter transmits symbols over a channel to a receiver 16 using a predetermined modulation scheme. The receiver uses knowledge of the modulation scheme, to decode the signal received from the transmitter. The decoded bits are provided to a sink device that is connected to the receiver.

[0020] A transmitter in accordance with an embodiment of the invention is shown in FIG. 2. The transmitter 14 includes a coder 20 that receives user bits from a source and encodes the bits in accordance with a predetermined coding scheme. In a number of embodiments, a capacity approaching code such as a turbo code or a LDPC code is used. In other embodiments, other coding schemes can be used to providing a coding gain within the communication system. A mapper 22 is connected to the coder. The mapper maps the bits output by the coder to a symbol within a geometrically distributed signal constellation stored within the mapper. The mapper provides the symbols to a modulator 24, which modulates the symbols for transmission via the channel.

[0021] A receiver in accordance with an embodiment

of the invention is illustrated in FIG. 3. The receiver 16 includes a demodulator 30 that demodulates a signal received via the channel to obtain symbol or bit likelihoods. The demapper uses knowledge of the geometrically shaped symbol constellation used by the transmitter to determine these likelihoods. The demapper 32 provides the likelihoods to a decoder 34 that decodes the encoded bit stream to provide a sequence of received bits to a sink.

GEOMETRICALLY SHAPED CONSTELLATIONS

[0022] Transmitters and receivers in accordance with embodiments of the invention utilize geometrically shaped symbol constellations. In several embodiments, a geometrically shaped symbol constellation is used that optimizes the capacity of the constellation. Various geometrically shaped symbol constellations that can be used in accordance with embodiments of the invention, techniques for deriving geometrically shaped symbol constellations are described below.

SELECTION OF A GEOMETRICALLY SHAPED CONSTELLATIONS

[0023] Selection of a geometrically shaped constellation for use in a communication system in accordance with an embodiment of the invention can depend upon a variety of factors including whether the code rate is fixed. In many embodiments, a geometrically shaped constellation is used to replace a conventional constellation (i.e. a constellation maximized for d_{min}) in the mapper of transmitters and the demapper of receivers within a communication system. Upgrading a communication system involves selection of a constellation and in many instances the upgrade can be achieved via a simple firmware upgrade. In other embodiments, a geometrically shaped constellation is selected in conjunction with a code rate to meet specific performance requirements, which can for example include such factors as a specified bit rate, a maximum transmit power. Processes for selecting a geometric constellation when upgrading existing communication systems and when designing new communication systems are discussed further below.

UPGRADING EXISTING COMMUNICATION SYSTEMS

[0024] A geometrically shaped constellation that provides a capacity, which is greater than the capacity of a constellation maximized for d_{min}, can be used in place of a conventional constellation in a communication system in accordance with embodiments of the invention. In many instances, the substitution of the geometrically shaped constellation can be achieved by a firmware or software upgrade of the transmitters and receivers within the communication system. Not all geometrically shaped constellations have greater capacity than that of a constellation maximized for d_{min}. One approach to selecting a geometrically shaped constellation having a greater capacity than that of a constellation maximized for d_{min} is to optimize the shape of the constellation with respect to a measure of the capacity of the constellation for a given SNR. Capacity measures that can be used in the optimization process can include, but are not limited to, joint capacity or parallel decoding capacity.

JOINT CAPACITY AND PARALLEL DECODING CAPACITY

[0025] A constellation can be parameterized by the total number of constellation points, M, and the number of real dimensions, N_{dim}. In systems where there are no belief propagation iterations between the decoder and the constellation demapper, the constellation demapper can be thought of as part of the channel. A diagram conceptually illustrating the portions of a communication system that can be considered part of the channel for the purpose of determining PD capacity is shown in FIG. 4a. The portions of the communication system that are considered part of the channel are indicated by the ghost line 40. The capacity of the channel defined as such is the parallel decoding (PD) capacity, given by:

$$C_{PD} = \sum_{i=0}^{l-1} I(X_i; Y)$$

where X_i is the ith bit of the *l*-bits transmitted symbol, and *Y* is the received symbol, and *I(A;B)* denotes the mutual information between random variables *A* and *B*.

[0026] Expressed another way, the PD capacity of a channel can be viewed in terms of the mutual information between the output bits of the encoder (such as an LDPC encoder) at the transmitter and the likelihoods computed by the demapper at the receiver. The PD capacity is influenced by both the placement of points within the constellation and by the labeling assignments.

[0027] With belief propagation iterations between the demapper and the decoder, the demapper can no longer be viewed as part of the channel, and the joint capacity of the constellation becomes the tightest known bound on the system performance. A diagram conceptually illustrating the portions of a communication system that are considered part of the channel for the purpose of determining the joint capacity of a constellation is shown in FIG. 4b. The portions of the communication system that are considered part of the channel are indicated by the ghost line 42. The joint capacity of the channel is given by:

$$C_{JOINT} = I(X; Y)$$

[0028] Joint capacity is a description of the achievable

capacity between the input of the mapper on the transmit side of the link and the output of the channel (including for example AWGN and Fading channels). Practical systems must often 'demap' channel observations prior to decoding. In general, the step causes some loss of capacity. In fact it can be proven that $C_G \geq C_{JOINT} \geq C_{PD}$. That is, C_{JOINT} upper bounds the capacity achievable by C_{PD}. The methods of the present invention are motivated by considering the fact that practical limits to a given communication system capacity are limited by C_{JOINT} and C_{PD}. In several embodiments of the invention, geometrically shaped constellations are selected that maximize these measures.

SELECTING A CONSTELLATION HAVING AN OPTIMAL CAPACITY

[0029] Geometrically shaped constellations in accordance with embodiments of the invention can be designed to optimize capacity measures including, but not limited to PD capacity or joint capacity. A process for selecting the points, and the labeling, of a geometrically shaped constellation for use in a communication system having a fixed code rate in accordance with an embodiment of the invention is shown in FIG. 5. The process 50 commences with the selection (52) of an appropriate constellation size M and a desired capacity per dimension η. In the illustrated embodiment, the process involves a check (52) to ensure that the constellation size can support the desired capacity. In the event that the constellation size could support the desired capacity, then the process iteratively optimizes the M-ary constellation for the specified capacity. Optimizing a constellation for a specified capacity often involves an iterative process, because the optimal constellation depends upon the SNR at which the communication system operates. The SNR for the optimal constellation to give a required capacity is not known *a priori*. Throughout the description of the present invention SNR is defined as the ratio of the average constellation energy per dimension to the average noise energy per dimension. In most cases the capacity can be set to equal the target user bit rate per symbol per dimension. In some cases adding some implementation margin on top of the target user bit rate could result in a practical system that can provide the required user rate at a lower rate. The margin is code dependent. The following procedure could be used to determine the target capacity that includes some margin on top of the user rate. First, the code (e.g. LDPC or Turbo) can be simulated in conjunction with a conventional equally spaced constellation. Second, from the simulation results the actual SNR of operation at the required error rate can be found. Third, the capacity of the conventional constellation at that SNR can be computed. Finally, a geometrically shaped constellation can be optimized for that capacity.

[0030] In the illustrated embodiment, the iterative optimization loop involves selecting an initial estimate of the SNR at which the system is likely to operate (i.e. SNR_{in}). In several embodiments the initial estimate is the SNR required using a conventional constellation. In other embodiments, other techniques can be used for selecting the initial SNR. An *M*-ary constellation is then obtained by optimizing (56) the constellation to maximize a selected capacity measure at the initial SNR_{in} estimate. Various techniques for obtaining an optimized constellation for a given SNR estimate are discussed below.

[0031] The SNR at which the optimized M-ary constellation provides the desired capacity per dimension η (SNR_{out}) is determined (57). A determination (58) is made as to whether the SNR_{out} and SNR_{in} have converged. In the illustrated embodiment convergence is indicated by SNR_{out} equaling SNR_{in}. In a number of embodiments, convergence can be determined based upon the difference between SNR_{out} and SNR_{in} being less than a predetermined threshold. When SNR_{out} and SNR_{in} have not converged, the process performs another iteration selecting SNR_{out} as the new SNR_{in} (55). When SNR_{out} and SNR_{in} have converged, the capacity measure of the constellation has been optimized. As is explained in more detail below, capacity optimized constellation at low SNRs are geometrically shaped constellations that can achieve significantly higher performance gains (measured as reduction in minimum required SNR) than constellations that maximize d_{min}.

[0032] The process illustrated in FIG. 5 can maximize PD capacity or joint capacity of an *M*-ary constellation for a given SNR. Although the process illustrated in FIG. 5 shows selecting an *M*-ary constellation optimized for capacity, a similar process could be used that terminates upon generation of an *M*-ary constellation where the SNR gap to Gaussian capacity at a given capacity is a predetermined margin lower than the SNR gap of a conventional constellation, for example 0.5 db. Alternatively, other processes that identify M-ary constellations having capacity greater than the capacity of a conventional constellation can be used in accordance with embodiments of the invention. A geometrically shaped constellation in accordance with embodiments of the invention can achieve greater capacity than the capacity of a constellation that maximizes d_{min} without having the optimal capacity for the SNR range within which the communication system operates.

[0033] We note that constellations designed to maximize joint capacity may also be particularly well suited to codes with symbols over GF(q), or with multi-stage decoding.. Conversely constellations optimized for PD capacity could be better suited to the more common case of codes with symbols over GF(2)

OPTIMIZING THE CAPACITY OF AN M-ARY CONSTELLATION AT A GIVEN SNR

[0034] Processes for obtaining a capacity optimized constellation often involve determining the optimum location for the points of an *M*-ary constellation at a given

SNR. An optimization process, such as the optimization process 56 shown in FIG. 5, typically involves unconstrained or constrained non-linear optimization. Possible objective functions to be maximized are the Joint or PD capacity functions. These functions may be targeted to channels including but not limited to Additive White Gaussian Noise (AWGN) or Rayleigh fading channels. The optimization process gives the location of each constellation point identified by its symbol labeling. In the case where the objective is joint capacity, point bit labelings are irrelevant meaning that changing the bit labelings doesn't change the joint capacity as long as the set of point locations remains unchanged..

[0035] The optimization process typically finds the constellation that gives the largest PD capacity or joint capacity at a given SNR. The optimization process itself often involves an iterative numerical process that among other things considers several constellations and selects the constellation that gives the highest capacity at a given SNR. In other embodiments, the constellation that requires the least SNR to give a required PD capacity or joint capacity can also be found. This requires running the optimization process iteratively as shown in FIG. 5.

[0036] Optimization constraints on the constellation point locations may include, but are not limited to, lower and upper bounds on point location, peak to average power of the resulting constellation, and zero mean in the resulting constellation. It can be easily shown that a globally optimal constellation will have zero mean (no DC component). Explicit inclusion of a zero mean constraint helps the optimization routine to converge more rapidly. Except for cases where exhaustive search of all combinations of point locations and labelings is possible it will not necessarily always be the case that solutions are provably globally optimal. In cases where exhaustive search is possible, the solution provided by the non-linear optimizer is in fact globally optimal.

[0037] The processes described above provide examples of the manner in which a geometrically shaped constellation having an increased capacity relative to a conventional capacity can be obtained for use in a communication system having a fixed code rate and modulation scheme. The actual gains achievable using constellations that are optimized for capacity compared to conventional constellations that maximize d_{min} are considered below.

GAINS ACHIEVED BY OPTIMIZED GEOMETRICALLY SPACED CONSTELLATIONS

[0038] The ultimate theoretical capacity achievable by any communication method is thought to be the Gaussian capacity, C_G which is defined as:

$$C_G = \frac{1}{2} log_2 (1 + SNR)$$

[0039] Where signal-to-noise (SNR) is the ratio of expected signal power to expected noise power. The gap that remains between the capacity of a constellation and C_G can be considered a measure of the quality of a given constellation design.

[0040] The gap in capacity between a conventional modulation scheme in combination with a theoretically optimal coder can be observed with reference to FIGS. 6a and 6b. FIG. 6a includes a chart 60 showing a comparison between Gaussian capacity and the PD capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}, Gaps 62 exist between the plot of Gaussian capacity and the PD capacity of the various PAM constellations. FIG. 6b includes a chart 64 showing a comparison between Gaussian capacity and the joint capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}, Gaps 66 exist between the plot of Gaussian capacity and the joint capacity of the various PAM constellations. These gaps in capacity represent the extent to which conventional PAM constellations fall short of obtaining the ultimate theoretical capacity i.e. the Gaussian capacity..

[0041] In order to gain a better view of the differences between the curves shown in FIGS. 6a and 6b at points close to the Gaussian capacity, the SNR gap to Gaussian capacity for different values of capacity for each constellation are plotted in FIG. 7. It is interesting to note from the chart 70 in FIG. 7 that (unlike the joint capacity) at the same SNR, the PD capacity does not necessarily increase with the number of constellation points. As is discussed further below, this is not the case with PAM constellations optimized for PD capacity.

[0042] FIGS. 8a and 8b summarize performance of constellations for PAM-4, 8, 16, and 32 optimized for PD capacity and joint capacity (it should be noted that BPSK is the optimal PAM-2 constellation at all code rates). The constellations are optimized for PD capacity and joint capacity for different target user bits per dimension (i.e. code rates). The optimized constellations are different depending on the target user bits per dimension, and also depending on whether they have been designed to maximize the PD capacity or the joint capacity. All the PD optimized PAM constellations are labeled using a gray labeling which is not always the binary reflective gray labeling. It should be noted that not all gray labels achieve the maximum possible PD capacity even given the freedom to place the constellation points anywhere on the real line. FIG. 8a shows the SNR gap for each constellation optimized for PD capacity. FIG. 8b shows the SNR gap to Gaussian capacity for each constellation optimized for joint capacity. Again, it should be emphasized that each '+' on the plot represents a different constellation.

[0043] Referring to FIG. 8a, the coding gain achieved using a constellation optimized for PD capacity can be appreciated by comparing the SNR gap at a user bit rate per dimension of 2.5 bits for PAM-32. A user bit rate per dimension of 2.5 bits for a system transmitting 5 bits per

symbol constitutes a code rate of 1/2. At that code rate the constellation optimized for PD capacity provides an additional coding gain of approximately 1.5 dB when compared to the conventional PAM-32 constellation.

[0044] The SNR gains that can be achieved using constellations that are optimized for PD capacity can be verified through simulation. The results of a simulation conducted using a rate 1/2 LDPC code in conjunction with a conventional PAM-32 constellation and in conjunction with a PAM-32 constellation optimized for PD capacity are illustrated in FIG. 9. A chart 90 includes plots of Frame Error Rate performance of the different constellations with respect to SNR and using different length codes (i.e. k=4,096 and k=16,384). Irrespective of the code that is used, the constellation optimized for PD capacity achieves a gain of approximately 1.3 dB, which closely approaches the gain predicted from FIG. 8a.

CAPACITY OPTIMIZED PAM CONSTELLATIONS

[0045] Using the processes outlined above, locus plots of PAM constellations optimized for capacity can be generated that show the location of points within PAM constellations versus SNR. Locus plots of PAM-4, 8, 16, and 32 constellations optimized for PD capacity and joint capacity and corresponding design tables at various typical user bit rates per dimension are illustrated in FIGS. 10a - 17b. The locus plots and design tables show PAM-4,8,16,32 constellation point locations and labelings from low to high SNR corresponding to a range of low to high spectral efficiency.

[0046] In FIG.10a, a locus plot 100 shows the location of the points of PAM-4 constellations optimized for Joint capacity plotted against achieved capacity. A similar locus plot 105 showing the location of the points of Joint capacity optimized PAM-4 constellations plotted against SNR is included in FIG. 10b. In FIG. 10c. the location of points for PAM-4 optimized for PD capacity is plotted against achievable capacity and in FIG. 10d the location of points for PAM-4 for PD capacity is plotted against SNR. At low SNRs, the PD capacity optimized PAM-4 constellations have only 2 unique points, while the Joint optimized constellations have 3. As SNR is increased, each optimization eventually provides 4 unique points. This phenomenon is explicitly described in FIG. 11a and FIG. 11b where vertical slices of FIGS. 10ab and 10cd are captured in tables describing some PAM-4 constellations designs of interest. The SNR slices selected represent designs that achieve capacities = {0.5, 0.75, 1.0, 1.25, 1.5} bits per symbol (bps). Given that PAM-4 can provide at most $\log_2(4) = 2$ bps, these design points represent systems with information code rates R = {1/4, 3/8, 1/2, 5/8, 3/4} respectively.

[0047] FIGS. 12ab and 12cd present locus plots of PD capacity and joint capacity optimized PAM-8 constellation points versus achievable capacity and SNR. FIGS. 13a and 13b provide slices from these plots at SNRs corresponding to achievable capacities η = {0.5, 1.0, 1.5, 2.0, 2.5} bps. Each of these slices correspond to systems with code rate R = η bps / $\log_2(8)$, resulting in R ={1/6, 1/3, 1/2, 2/3, 5/6}. As an example of the relative performance of the constellations in these tables, consider FIG. 13b which shows a PD capacity optimized PAM-8 constellation optimized for SNR = **9.00 dB,** or 1.5 bps. We next examine the plot provided in FIG. **8a** and see that the gap of the optimized constellation to the ultimate, Gaussian, capacity (CG) is approximately 0.5 dB. At the same spectral efficiency, the gap of the traditional PAM-8 constellation is approximately 1.0 dB. The advantage of the optimized constellation is 0.5dB for the same rate (in this case R=1/2). This gain can be obtained by only changing the mapper and demapper in the communication system and leaving all other blocks the same.

[0048] Similar information is presented in FIGS. 14abcd, and 15ab which provide loci plots and *design* tables for PAM-16 PD capacity and joint capacity optimized constellations. Likewise FIGS. 16abcd, 17ab provide loci plots and design tables for PAM-32 PD capacity and joint capacity optimized constellations.

CAPACITY OPTIMIZED PSK CONSTELLATIONS

[0049] Traditional phase shift keyed (PSK) constellations are already quite optimal. This can be seen in the chart 180 comparing the SNR gaps of tradition PSK with capacity optimized PSK constellations shown in FIG. 18 where the gap between PD capacity and Gaussian capacity is plotted for traditional PSK-4,8,16,32 and for PD capacaity optimized PSK-4,8,16,32.

[0050] The locus plot of PD optimized PSK-32 points across SNR is shown in FIG. 19, which actually characterizes all PSKs with spectral efficiency $\eta \leq 5$. This can be seen in FIG. 20. Note that at low SNR (0.4 dB) the optimal PSK-32 design is the same as traditional PSK-4, at SNR = 8.4 dB optimal PSK-32 is the same as traditional PSK-8, at SNR = 14.8 dB optimal PSK-32 is the same as traditional PSK-16, and finally at SNRs greater than 20.4 dB optimized PSK-32 is the same as traditional PSK-32. There are SNRs between these discrete points (for instance SNR = 2 and 15. dB) for which optimized PSK-32 provides superior PD capacity when compared to traditional PSK constellations.

[0051] We note now that the locus of points for PD optimized PSK-32 in FIG. 19 in conjunction with the gap to Gaussian capacity curve for optimized PSK-32 in FIG. 18 implies a potential design methodology. Specifically, the designer could achieve performance equivalent or better than that enabled by traditional PSK-4,8,16 by using only the optimized PSK-32 in conjunction with a single tuning parameter that controlled where the constellation points should be selected from on the locus of FIG. 19. Such an approach would couple a highly rate adaptive channel code that could vary its rate, for instance, rate 4/5 to achieve and overall (code plus optimized PSK-32 modulation) spectral efficiency of 4 bits per symbol, down to 1/5 to achieve an overall spectral efficiency of 1 bit per

symbol. Such an adaptive modulation and coding system could essentially perform on the optimal continuum represented by the rightmost contour of FIG. 18.

ADAPTIVE RATE DESIGN

[0052] In the previous example spectrally adaptive use of PSK-32 was described. Techniques similar to this can be applied for other capacity optimized constellations across the link between a transmitter and receiver. For instance, in the case where a system implements quality of service it is possible to instruct a transmitter to increase or decrease spectral efficiency on demand. In the context of the current invention a capacity optimized constellation designed precisely for the target spectral efficiency can be loaded into the transmit mapper in conjunction with a code rate selection that meets the end user rate goal. When such a modulation / code rate change occurred a message could propagated to the receiver so that the receiver, in anticipation of the change, could select a demapper / decoder configuration in order to match the new transmit-side configuration.

[0053] Conversely, the receiver could implement a quality of performance based optimized constellation / code rate pair control mechanism. Such an approach would include some form of receiver quality measure. This could be the receiver's estimate of SNR or bit error rate. Take for example the case where bit error rate was above some acceptable threshold. In this case, via a backchannel, the receiver could request that the transmitter lower the spectral efficiency of the link by swapping to an alternate capacity optimized constellation / code rate pair in the coder and mapper modules and then signaling the receiver to swap in the complementary pairing in the demapper / decoder modules.

GEOMETRICALLY SHAPED QAM CONSTELLATIONS

[0054] Quadrature amplitude modulation (QAM) constellations can be constructed by orthogonalizing PAM constellations into QAM inphase and quadrature components. Constellations constructed in this way can be attractive in many applications because they have low-complexity demappers.

[0055] In FIG. 21 we provide an example of a Quadrature Amplitude Modulation constellation constructed from a Pulse Amplitude Modulation constellation. The illustrated embodiment was constructed using a PAM-8 constellation optimized for PD capacity at user bit rate per dimension of 1.5 bits (corresponds to an SNR of 9.0 dB) (see FIG. 13b). The label-point pairs in this PAM-8 constellation are {(000, -1.72), (001, -0.81), (010, 1.72), (011,-0.62), (100, 0.62), (101, 0.02), (110, 0.81), (111, -0.02)}. Examination of FIG. 21 shows that the QAM constellation construction is achieved by replicating a complete set of PAM-8 points in the quadrature dimension for each of the 8 PAM-8 points in the in-phase dimension. Labeling is achieved by assigning the PAM-8 labels to the LSB range on the in-phase dimension and to the MSB range on the quadrature dimension. The resulting 8x8 outer product forms a highly structured QAM-64 for which very low-complexity de-mappers can be constructed. Due to the orthogonality of the in-phase and quadrature components the capacity characteristics of the resulting QAM-64 constellation are identical to that of the PAM-8 constellation on a per-dimension basis.

N-DIMENSIONAL CONSTELLATION OPTIMIZATION

[0056] Rather than designing constellations in 1-D (PAM for instance) and then extending to 2-D (QAM), it is possible to take direct advantage in the optimization step of the additional degree of freedom presented by an extra spatial dimension. In general it is possible to design N-dimensional constellations and associated labelings. The complexity of the optimization step grows exponentially in the number of dimensions as does the complexity of the resulting receiver de-mapper. Such constructions constitute embodiments of the invention and simply require more 'run-time' to produce.

CAPACITY. OPTIMIZED CONSTELALTIONS FOR FADING CHANNELS

[0057] Similar processes to those outlined above can be used to design capacity optimized constellations for fading channels in accordance with embodiments of the invention. The processes are essentially the same with the exception that the manner in which capacity is calculated is modified to account for the fading channel. A fading channel can be described using the following equation:

$$Y = a(t).X+N$$

where X is the transmitted signal, N is an additive white Gaussian noise signal and $a(t)$ is the fading distribution, which is a function of time.

[0058] In the case of a fading channel, the instantaneous SNR at the receiver changes according to a fading distribution. The fading distribution is Rayleigh and has the property that the average SNR of the system remains the same as in the case of the AWGN channel, $E[X^2]/E[N^2]$. Therefore, the capacity of the fading channel can be computed by taking the expectation of AWGN capacity, at a given average SNR, over the Rayleigh fading distribution of a that drives the distribution of the instantaneous SNR.

[0059] Many fading channels follow a Rayleigh distribution. FIGS. 22a - 24b are locus plots of PAM-4, 8, and 16 constellations that have been optimized for PD capacity on a Rayleigh fading channel. Locus plots versus user bit rate per dimension and versus SNR are provided. Similar processes can be used to obtain capacity opti-

mized constellations that are optimized using other capacity measures, such as joint capacity, and/or using different modulation schemes.

Claims

1. A digital communication system (10), comprising:

   a transmitter (14) configured to transmit signals via a communication channel (15);
   wherein the transmitter (14), comprises:

   a coder (20) configured to receive user bits and output encoded bits at an expanded output encoded bit rate;
   a mapper (22) configured to map encoded bits to symbols in a modulated symbol constellation; and
   a modulator (24) configured to generate a modulated signal for transmission via the communication channel using symbols generated by the mapper.

   wherein the transmitter is **characterized in that** the symbol constellation is an unequally spaced symbol constellation obtained by optimizing locations and labeling of constellation points to maximize parallel decode capacity at a given signal-to-noise ratio so that the symbol constellation provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

2. The communication system (10) of claim 1, further comprising:

   a receiver (16) configured to receive signals via a communication channel (15),
   wherein the receiver comprises:

   a demodulator (30) configured to demodulate a signal received via the communication channel;
   a demapper (32) configured to estimate likelihoods from the demodulated signal using the symbol constellation; and
   a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32).

3. A digital communication system (10), comprising:

   a receiver (16) configured to receive signals via a communication channel (15); the receiver comprising:

   a demodulator (30) configured to demodulate a signal received via the communication channel;
   a demapper (32) configured to estimate likelihoods from the demodulated signal using a symbol constellation;
   a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32); and

   wherein the receiver is **characterized in that** the symbol constellation is an unequally spaced symbol constellation obtained by optimizing locations and labeling of constellation points to maximize parallel decode capacity at a given signal-to-noise ratio so that the symbol constellation provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

4. The communication system (10) of any preceding claim, wherein the unequally spaced symbol constellation is optimized subject to optimization constraints including any one of:

   lower and upper bounds on point location;
   peak to average power of the resulting constellation; and
   mean in the resulting constellation.

5. The communication system (10) of any preceding claim, wherein the channel is an Additive White Gaussian Noise, AWGN, channel.

6. The communication system (10) of any preceding claim, wherein the channel is a fading channel.

7. The communication system (10) of any preceding claim , wherein the constellations points of the unequally spaced symbol constellation are variable depending on the user rate.

8. The communication system (10) of any preceding claim, wherein the constellation points were obtained by optimizing the locations and the labeling of the constellation points to maximize parallel decode capacity measure at a given signal-to-noise ratio in such a way that the obtained constellation points have a greater capacity than the capacity of a constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

9. The communication system (10) of any preceding claim wherein the constellation points provide an SNR gap to Gaussian capacity at a given capacity

that is a predetermined margin lower than the signal-to-noise ratio gap to Gaussian capacity of a signal constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

10. The communication system (10) of claim 9, wherein the predetermined margin is 0.5 dB.

11. The communication system (10) of any preceding claim, wherein the symbol constellation comprises a ring of constellation points that are unevenly spaced around the ring.

12. The communication system (10) of claims 11, wherein the symbol constellation is a non-uniform PSK constellation.

13. The communication system (10) of any preceding claim, configured to utilize an LDPC code.

14. The communication system (10) of any of claims 2 or 3 wherein there are no belief propagation iterations between the decoder and the constellation demapper.

15. The communication system (10) of any preceding claim, wherein the constellation is selected in conjunction with a code rate to meet a maximum transmit power requirement.

16. A method performed in digital communication system (10), the method comprising:

a transmitter (14) transmitting signals via a communication channel (15) by;
a coder (20) receiving user bits and outputing encoded bits at an expanded output encoded bit rate;
a mapper (22) mapping encoded bits to symbols in a modulated symbol constellation; and
a modulator (24) generating a modulated signal for transmission via the communication channel using symbols generated by the mapper,
wherein the method is **characterized in that** the symbol constellation is an unequally spaced symbol constellation obtained by optimizing locations and labeling of constellation points to maximize parallel decode capacity at a given signal-to-noise ratio so that the symbol constellation provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

17. A method performed in a digital communication system (10), the method comprising:

a receiver (16) receiving signals via a communication channel (15) by:

a demodulator (30) demodulating a signal received via the communication channel;
a demapper (32) estimating likelihoods from the demodulated signal using a symbol constellation; and
a decoder (34) estimating decoded bits from the likelihoods generated by the demapper (32),

wherein the method is **characterized in that** the symbol constellation is an unequally spaced symbol constellation obtained by optimizing locations and labeling of constellation points to maximize parallel decode capacity at a given signal-to-noise ratio so that the symbol constellation provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

Patentansprüche

1. Digitales Kommunikationssystem (10), umfassend:

einen Sender (14), der dazu konfiguriert ist, Signale über einen Kommunikationskanal (15) zu übertragen; wobei der Sender (14) Folgendes umfasst:

einen Codierer (20), der dazu konfiguriert ist, Benutzerbits zu empfangen und codierte Bits mit einer erweiterten Ausgangscodierbitrate auszugeben;
einen Mapper (22), der dazu konfiguriert ist, die codierten Bits Symbolen in einer modulierten Symbolkonstellation zuzuordnen; und
einen Modulator (24), der dazu konfiguriert ist, ein moduliertes Signal zur Übertragung über den Kommunikationskanal unter Verwendung der durch den Mapper generierten Symbole zu generieren,

wobei der Sender **dadurch gekennzeichnet ist, dass** die Symbolkonstellation eine ungleichmäßig beabstandete Symbolkonstellation ist, die durch Optimieren von Stellen und Markierungen von Konstellationspunkten erlangt wird, um eine parallele Decodierkapazität bei einem gegebenen Signal-Rausch-Verhältnis zu maximieren, sodass die Symbolkonstellation eine gegebene Kapazität bei einem reduzierten Signal-Rausch-Verhältnis im Vergleich zu einer Si-

gnalkonstellation, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert, bereitstellt.

2. Kommunikationssystem (10) nach Anspruch 1, ferner umfassend:
einen Empfänger (16), der dazu konfiguriert ist, Signale über einen Kommunikationskanal (15) zu empfangen, wobei der Empfänger Folgendes umfasst:

einen Demodulator (30), der dazu konfiguriert ist, ein über den Kommunikationskanal empfangenes Signal zu demodulieren;
einen Demapper (32), der dazu konfiguriert ist, Wahrscheinlichkeiten aus dem demodulierten Signal unter Verwendung der Symbolkonstellation zu schätzen; und
einen Decodierer (34), der dazu konfiguriert ist, die decodierten Bits aus den durch den Demapper (32) generierten Wahrscheinlichkeiten zu schätzen.

3. Digitales Kommunikationssystem (10), umfassend:

einen Empfänger (16), der dazu konfiguriert ist, Signale über einen Kommunikationskanal (15) zu empfangen, wobei der Empfänger Folgendes umfasst:

einen Demodulator (30), der dazu konfiguriert ist, ein über den Kommunikationskanal empfangenes Signal zu demodulieren;
einen Demapper (32), der dazu konfiguriert ist, Wahrscheinlichkeiten aus dem demodulierten Signal unter Verwendung einer Symbolkonstellation zu schätzen;
einen Decodierer (34), der dazu konfiguriert ist, die decodierten Bits aus den durch den Demapper (32) generierten Wahrscheinlichkeiten zu schätzen; und

wobei der Empfänger **dadurch gekennzeichnet ist, dass** die Symbolkonstellation eine ungleichmäßig beabstandete Symbolkonstellation ist, die durch Optimieren von Stellen und Markierungen von Konstellationspunkten erlangt wird, um eine parallele Decodierkapazität bei einem gegebenen Signal-Rausch-Verhältnis zu maximieren, sodass die Symbolkonstellation eine gegebene Kapazität bei einem reduzierten Signal-Rausch-Verhältnis im Vergleich zu einer Signalkonstellation, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert, bereitstellt.

4. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die ungleichmäßig beabstandete Symbolkonstellation optimiert ist, vorbehaltlich Optimierungseinschränkungen, die eines der folgenden beinhalten:

untere und obere Begrenzung an der Punktstelle;
Spitzenleistung zu Durchschnittsleistung der resultierenden Konstellation; und
Mittelwert in der resultierenden Konstellation.

5. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei der Kanal ein Kanal des additiven weißen Gaußschen Rauschens (AWGN) ist.

6. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei der Kanal ein Fading-Kanal ist.

7. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die Konstellationspunkte der ungleichmäßig beabstandeten Symbolkonstellation abhängig von der Benutzerrate variabel sind.

8. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die Konstellationspunkte durch Optimieren der Stellen und der Markierung von Konstellationspunkten erlangt wurden, um ein paralleles Decodierkapazitätsmaß bei einem gegebenen Signal-Rausch-Verhältnis in einer Weise zu maximieren, dass die erlangten Konstellationspunkte eine größere Kapazität aufweisen als die Kapazität einer Konstellation, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert.

9. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die Konstellationspunkte eine SNR-Lücke zur Gaußschen Kapazität bei einer gegebenen Kapazität bereitstellen, die eine vorbestimmte Toleranz geringer als die Signal-Rausch-Verhältnislücke zur Gaußschen Kapazität einer Signalkonstellation ist, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert.

10. Kommunikationssystem (10) nach Anspruch 9, wobei die vorbestimmte Toleranz 0,5 dB ist.

11. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die Symbolkonstellation einen Ring von Konstellationspunkten umfasst, die ungleichmäßig um den Ring beabstandet sind.

12. Kommunikationssystem (10) nach Anspruch 11, wobei die Symbolkonstellation eine nicht einheitliche PSK-Konstellation ist.

13. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, dazu konfiguriert, einen LDPC-Code zu nutzen.

14. Kommunikationssystem (10) nach einem der Ansprüche 2 oder 3, wobei es keine Glaubenspropagationswiederholungen zwischen dem Decodierer und dem Konstellationsdemapper gibt.

15. Kommunikationssystem (10) nach einem der vorstehenden Ansprüche, wobei die Konstellation in Verbindung mit einer Coderate ausgewählt ist, um eine maximale Übertragungsleistungsanforderung zu erfüllen.

16. Verfahren, durchgeführt in einem digitalen Kommunikationssystem (10), wobei das Verfahren Folgendes umfasst:

einen Sender (14), der Signale über einen Kommunikationskanal (15) überträgt, durch:

einen Codierer (20), der Benutzerbits empfängt und codierte Bits mit einer erweiterten Ausgangscodierbitrate ausgibt;
einen Mapper (22), der die codierten Bits Symbolen in einer modulierten Symbolkonstellation zuordnet; und
einen Modulator (24), der ein moduliertes Signal zur Übertragung über den Kommunikationskanal unter Verwendung der durch den Mapper generierten Symbole generiert,

wobei das Verfahren **dadurch gekennzeichnet ist, dass** die Symbolkonstellation eine ungleichmäßig beabstandete Symbolkonstellation ist, die durch Optimieren von Stellen und Markierungen von Konstellationspunkten erlangt wird, um eine parallele Decodierkapazität bei einem gegebenen Signal-Rausch-Verhältnis zu maximieren, sodass die Symbolkonstellation eine gegebene Kapazität bei einem reduzierten Signal-Rausch-Verhältnis im Vergleich zu einer Signalkonstellation, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert, bereitstellt.

17. Verfahren, durchgeführt in einem digitalen Kommunikationssystem (10), wobei das Verfahren Folgendes umfasst:

einen Empfänger (16), der Signale über einen Kommunikationskanal (15) empfängt, durch:

einen Demodulator (30), der ein über den Kommunikationskanal empfangenes Signal demoduliert;
einen Demapper (32), der Wahrscheinlichkeiten aus dem demodulierten Signal unter Verwendung einer Symbolkonstellation schätzt; und
einen Decodierer (34), der die decodierten Bits aus den durch den Demapper (32) generierten Wahrscheinlichkeiten schätzt;

wobei das Verfahren **dadurch gekennzeichnet ist, dass** die Symbolkonstellation eine ungleichmäßig beabstandete Symbolkonstellation ist, die durch Optimieren von Stellen und Markierungen von Konstellationspunkten erlangt wird, um eine parallele Decodierkapazität bei einem gegebenen Signal-Rausch-Verhältnis zu maximieren, sodass die Symbolkonstellation eine gegebene Kapazität bei einem reduzierten Signal-Rausch-Verhältnis im Vergleich zu einer Signalkonstellation, die den Mindestabstand, d_{min}, zwischen den Konstellationspunkten der Symbolkonstellation maximiert, bereitstellt.

Revendications

1. Système de communication numérique (10), comprenant :

un transmetteur (14) configuré pour transmettre des signaux par l'intermédiaire d'un canal de communication (15) ;
dans lequel le transmetteur (14) comprend :

un codeur (20) configuré pour recevoir des bits utilisateur et des bits codés de sortie à une vitesse de bits codés de sortie accrue ;
un mappeur (22) configuré pour faire correspondre les bits codés à des symboles dans une constellation à symboles modulée ; et
un modulateur (24) configuré pour générer un signal modulé pour la transmission par l'intermédiaire du canal de communication à l'aide de symboles générés par le mappeur.

le transmetteur étant **caractérisé en ce que** la constellation à symboles est une constellation à symboles inégalement espacée obtenue par l'optimisation des emplacements et l'étiquetage des points de constellation pour maximiser la capacité de décodage parallèle à un rapport signal sur bruit donné de sorte que la constellation à symboles fournit une capacité donnée à un rapport signal sur bruit réduit en comparaison avec une constellation de signaux qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

2. Système de communication (10) selon la revendication 1, comprenant en outre :

un récepteur (16) configuré pour recevoir des signaux par l'intermédiaire d'un canal de communication (15),
dans lequel le récepteur comprend :

un démodulateur (30) configuré pour démoduler un signal reçu par l'intermédiaire du canal de communication ;
un démappeur (32) configuré pour estimer les probabilités à partir du signal démodulé à l'aide de la constellation à symboles ; et
un décodeur (34) qui est configuré pour estimer les bits décodés à partir des probabilités générées par le démappeur (32) .

3. Système de communication numérique (10), comprenant :

un récepteur (16) configuré pour recevoir des signaux par l'intermédiaire d'un canal de communication (15) ; le récepteur comprenant :

un démodulateur (30) configuré pour démoduler un signal reçu par l'intermédiaire du canal de communication ;
un démappeur (32) configuré pour estimer les probabilités à partir du signal démodulé à l'aide d'une constellation à symboles ;
un décodeur (34) qui est configuré pour estimer des bits décodés à partir des probabilités générées par le démappeur (32) ; et

dans lequel le récepteur est **caractérisé en ce que** la constellation à symboles est une constellation à symboles inégalement espacée obtenue par l'optimisation des emplacements et l'étiquetage des points de constellation pour maximiser la capacité de décodage parallèle à un rapport signal sur bruit donné de sorte que la constellation à symboles fournit une capacité donnée à un rapport signal sur bruit réduit en comparaison avec une constellation de signal qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

4. Système de communication (10) selon une quelconque revendication précédente, dans lequel la constellation à symboles inégalement espacée est optimisée en étant sujette aux contraintes d'optimisation comprenant l'un quelconque parmi :

des limites supérieures et inférieures sur l'emplacement des points ;
un pic pour la puissance moyenne de la constellation résultante ; et
une moyenne dans la constellation résultante.

5. Système de communication (10) selon une quelconque revendication précédente, dans lequel le canal est un canal à bruit blanc gaussien additif (AWGN).

6. Système de communication (10) selon une quelconque revendication précédente, dans lequel le canal est un canal à évanouissement.

7. Système de communication (10) selon une quelconque revendication précédente, dans lequel les points de constellation de la constellation à symboles inégalement espacée sont variables en fonction de la vitesse utilisateur.

8. Système de communication (10) selon une quelconque revendication précédente, dans lequel les points de constellation ont été obtenus par l'optimisation des emplacements et l'étiquetage des points de constellation pour maximiser la mesure de capacité de décodage parallèle à un rapport signal sur bruit donné de sorte que les points de constellation obtenus présentent une meilleure capacité que la capacité d'une constellation qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

9. Système de communication (10) selon une quelconque revendication précédente, dans lequel les points de constellation fournissent un écart SNR à la capacité gaussienne à une capacité donnée qui est une marge prédéterminée inférieure à l'écart de rapport signal sur bruit à la capacité gaussienne d'une constellation de signal qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

10. Système de communication (10) selon la revendication 9, dans lequel la marge prédéterminée est de 0,5 dB.

11. Système de communication (10) selon une quelconque revendication précédente, dans lequel la constellation à symboles comprend une bague de points de constellation qui sont espacés de manière inégale autour de la bague.

12. Système de communication (10) selon la revendication 11, dans lequel la constellation à symboles est une constellation PSK non uniforme.

13. Système de communication (10) selon une quelconque revendication précédente, configuré pour utiliser un code LDPC.

14. Système de communication (10) selon l'une quel-

conque des revendications 2 ou 3, dans lequel il n'existe pas d'itération de propagation de croyance entre le décodeur et le démappeur de constellation.

15. Système de communication (10) selon une quelconque revendication précédente, dans lequel la constellation est choisie en conjonction avec une vitesse de code pour satisfaire à une exigence d'énergie de transmission maximale.

16. Procédé effectué dans le système de communication numérique (10), le procédé comprenant :

un transmetteur (14) transmettant les signaux par l'intermédiaire d'un canal de communication (15) par ;
un codeur (20) recevant les bits utilisateur et les bits codés de sortie à une vitesse de bits codés de sortie accrue ;
un mappeur (22) faisant correspondre les bits codés à des symboles dans une constellation à symboles modulée ; et
un modulateur (24) générant un signal modulé pour la transmission par l'intermédiaire du canal de communication à l'aide de symboles générés par le mappeur,
le procédé étant **caractérisé en ce que** la constellation à symboles est une constellation à symboles inégalement espacée obtenue par l'optimisation des emplacements et l'étiquetage des points de constellation pour maximiser la capacité de décodage parallèle à un rapport signal sur bruit donné de sorte que la constellation à symboles fournit une capacité donnée à un rapport signal sur bruit réduit en comparaison avec une constellation de signal qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

17. Procédé effectué dans un système de communication numérique (10), le procédé comprenant :

un récepteur (16) recevant des signaux par l'intermédiaire d'un canal de communication (15) par :

un démodulateur (30) démodulant un signal reçu par l'intermédiaire du canal de communication ;
un démappeur (32) estimant les probabilités à partir du signal démodulé à l'aide d'une constellation à symboles ; et
un décodeur (34) estimant des bits décodés à partir des probabilités générées par le démappeur (32),

le procédé étant **caractérisé en ce que** la constellation à symboles est une constellation à symboles inégalement espacée obtenue par l'optimisation des emplacements et l'étiquetage des points de constellation pour maximiser la capacité de décodage parallèle à une rapport signal sur bruit donné de sorte que la constellation à symboles fournit une capacité donnée à un rapport signal sur bruit réduit en comparaison avec une constellation de signal qui maximise la distance minimale, d_{min}, entre les points de constellation de la constellation à symboles.

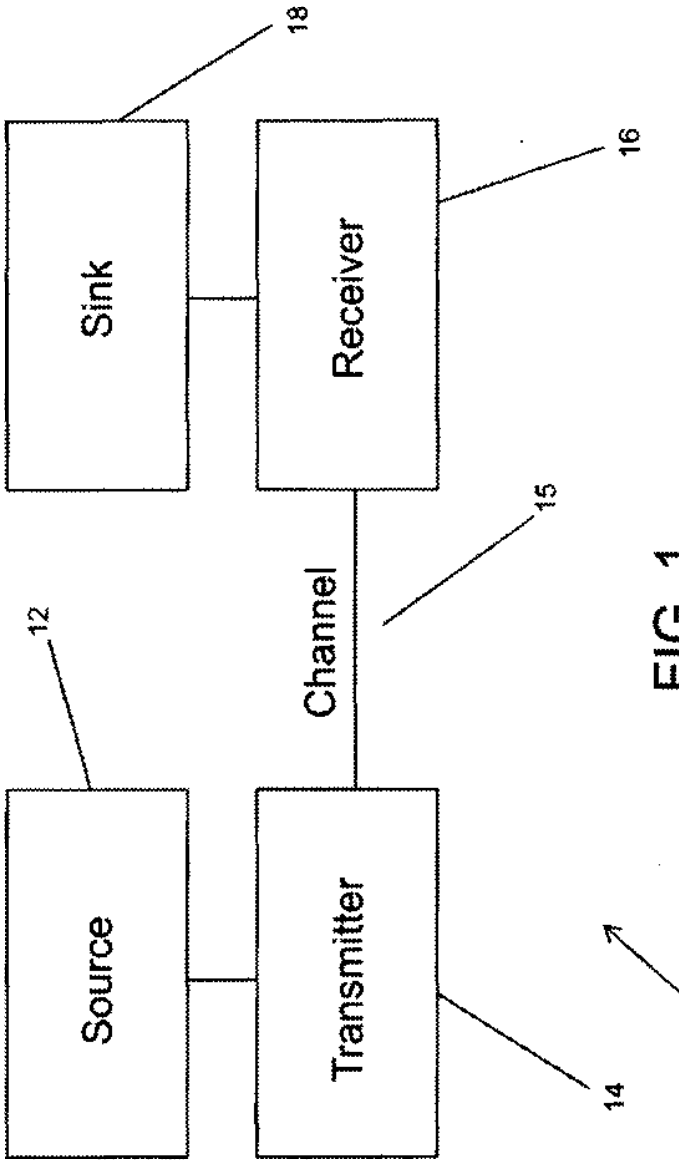


FIG. 1

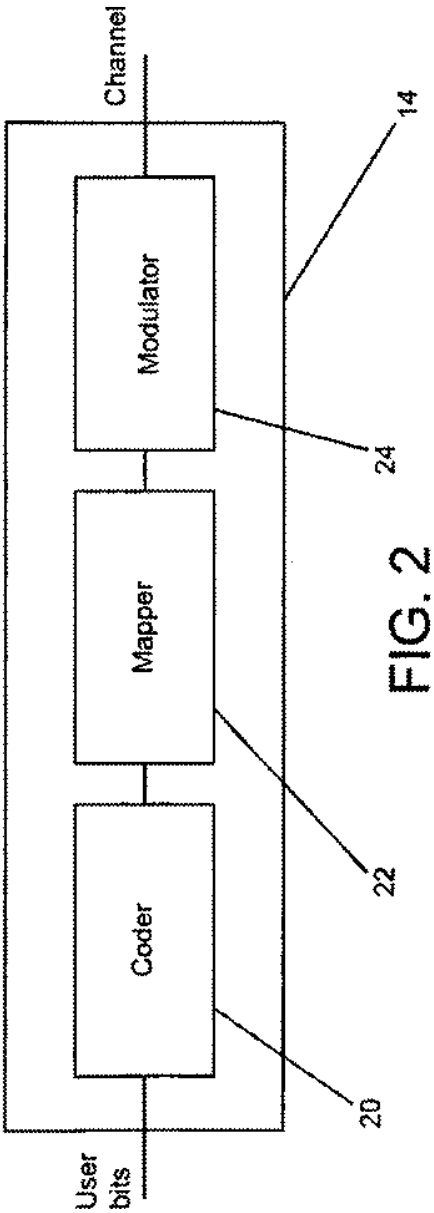

Channel
Modulator
Mapper
Coder
User bits
14
24
22
20

FIG. 2

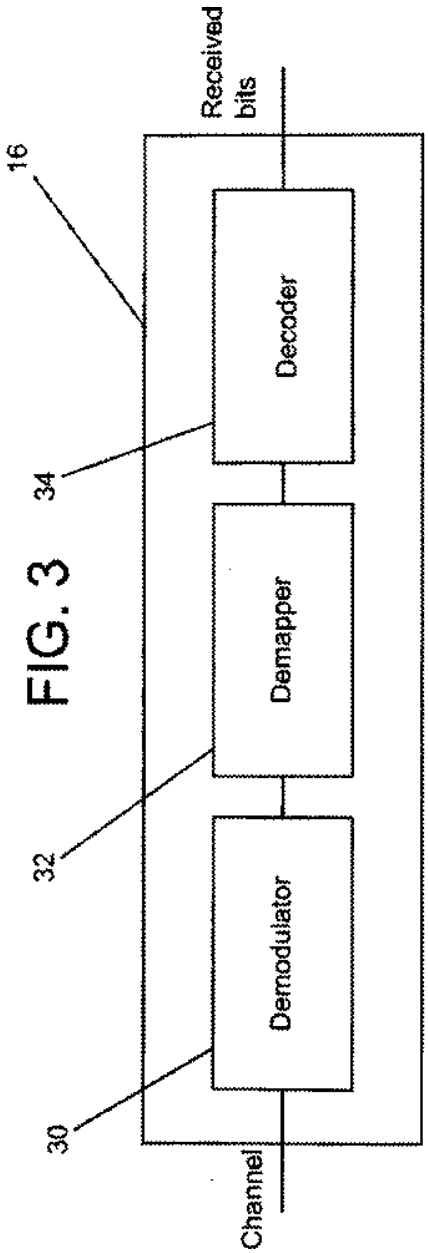

Received bits
Decoder
Demapper
Demodulator
Channel
16
34
32
30

FIG. 3

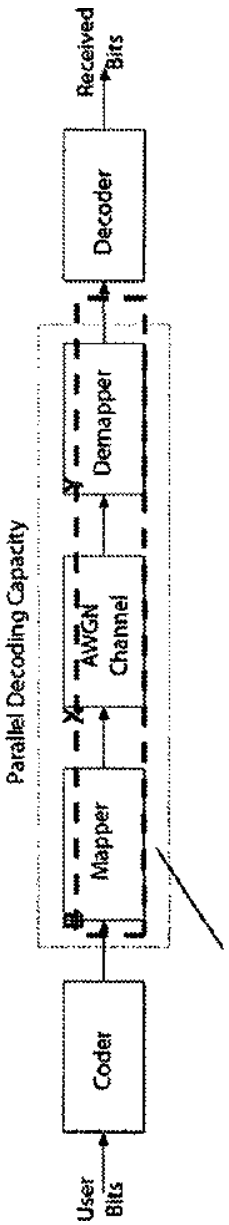

User Bits
Coder
B
Mapper
X
AWGN Channel
Y
Demapper
Decoder
Received Bits
Parallel Decoding Capacity
40

FIG. 4a

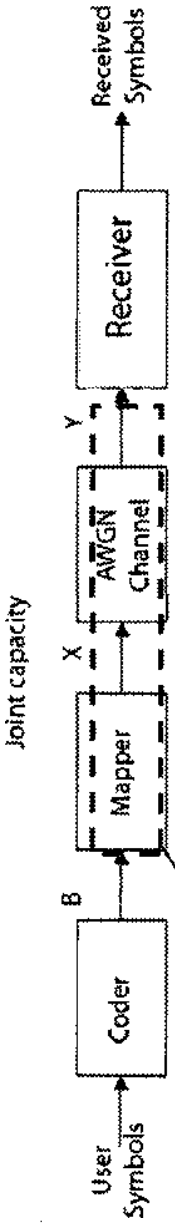

User Symbols
Coder
B
Mapper
X
AWGN Channel
Y
Receiver
Received Symbols
Joint capacity
42

FIG. 4b

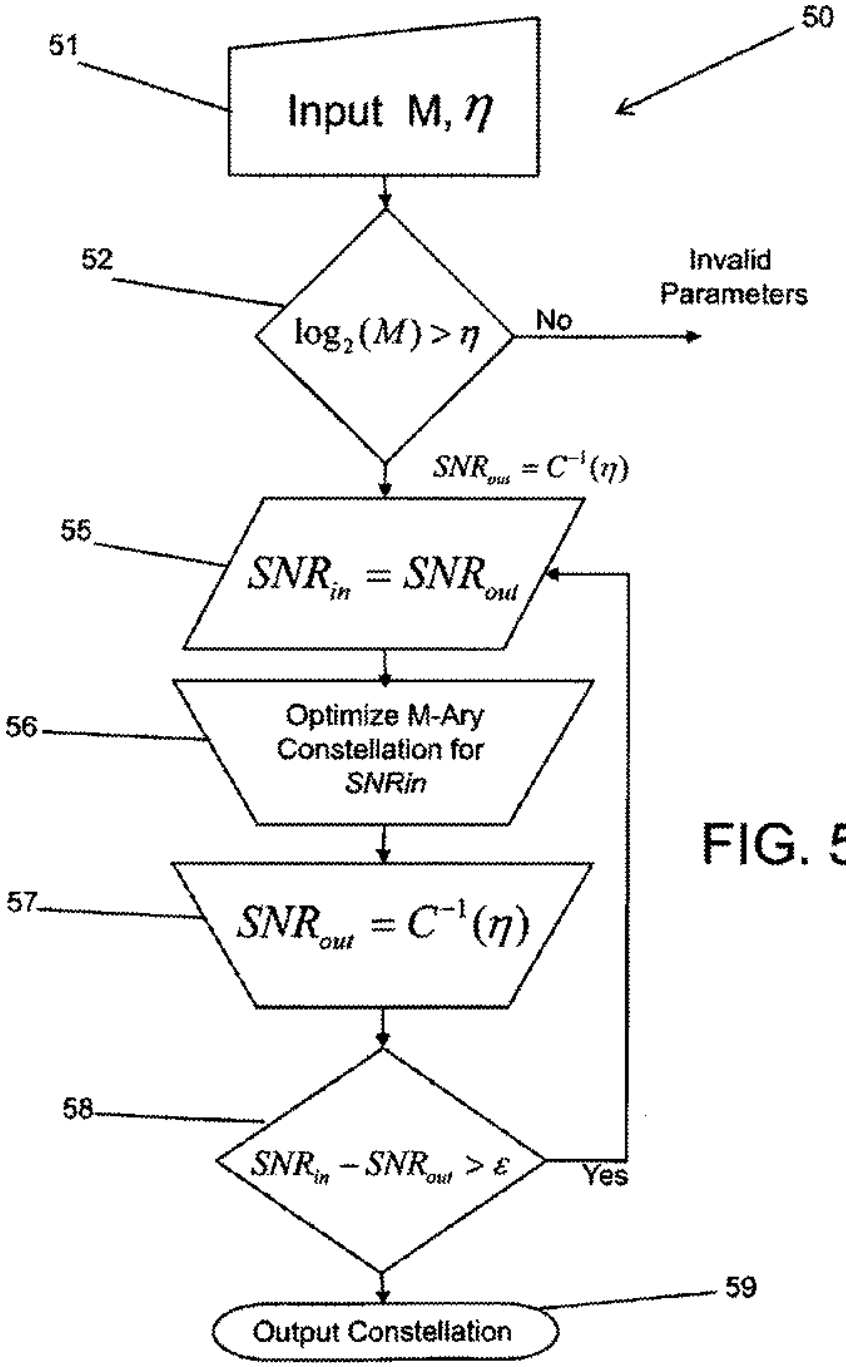

50
51
Input M, η
52
log2(M) > η
No
Invalid Parameters
SNRout = C-1(η)
55
SNRin = SNRout
56
Optimize M-Ary Constellation for SNRin
57
SNRout = C-1(η)
58
SNRin − SNRout > ε
Yes
59
Output Constellation

FIG. 5

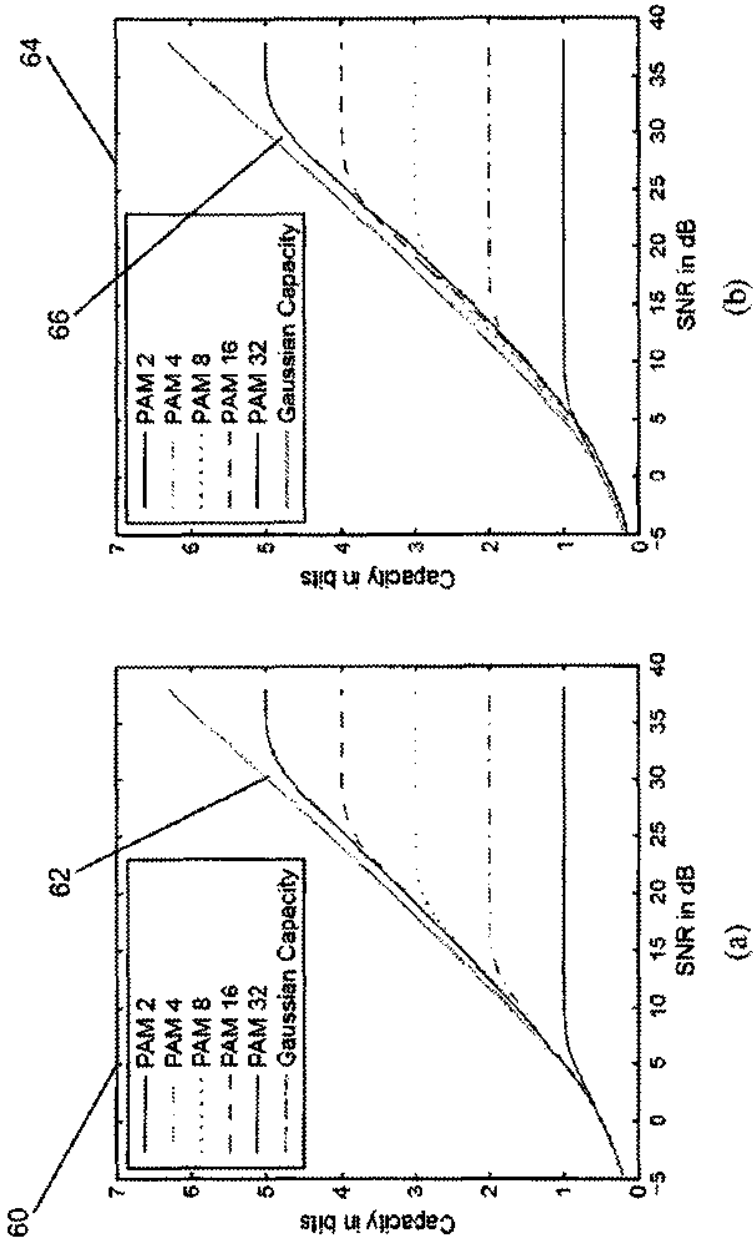

64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(b)

FIG. 6b

60
62
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(a)

FIG. 6a

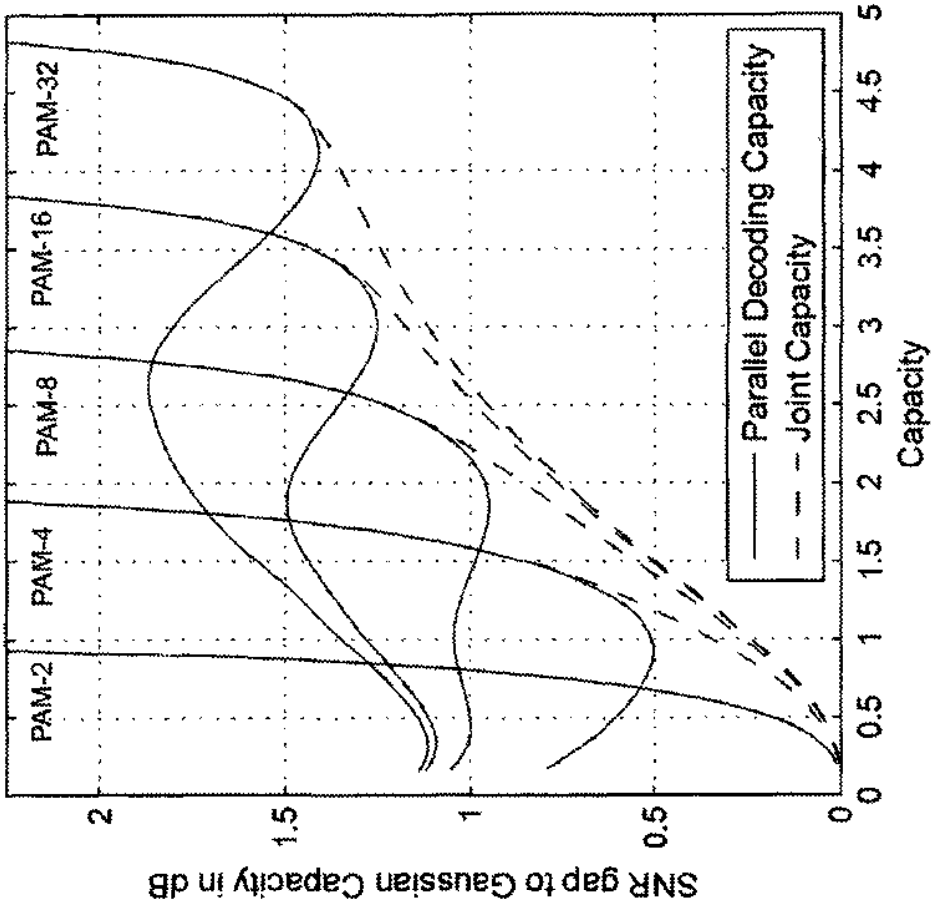

PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
Parallel Decoding Capacity
Joint Capacity
SNR gap to Gaussian Capacity in dB
Capacity
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 7

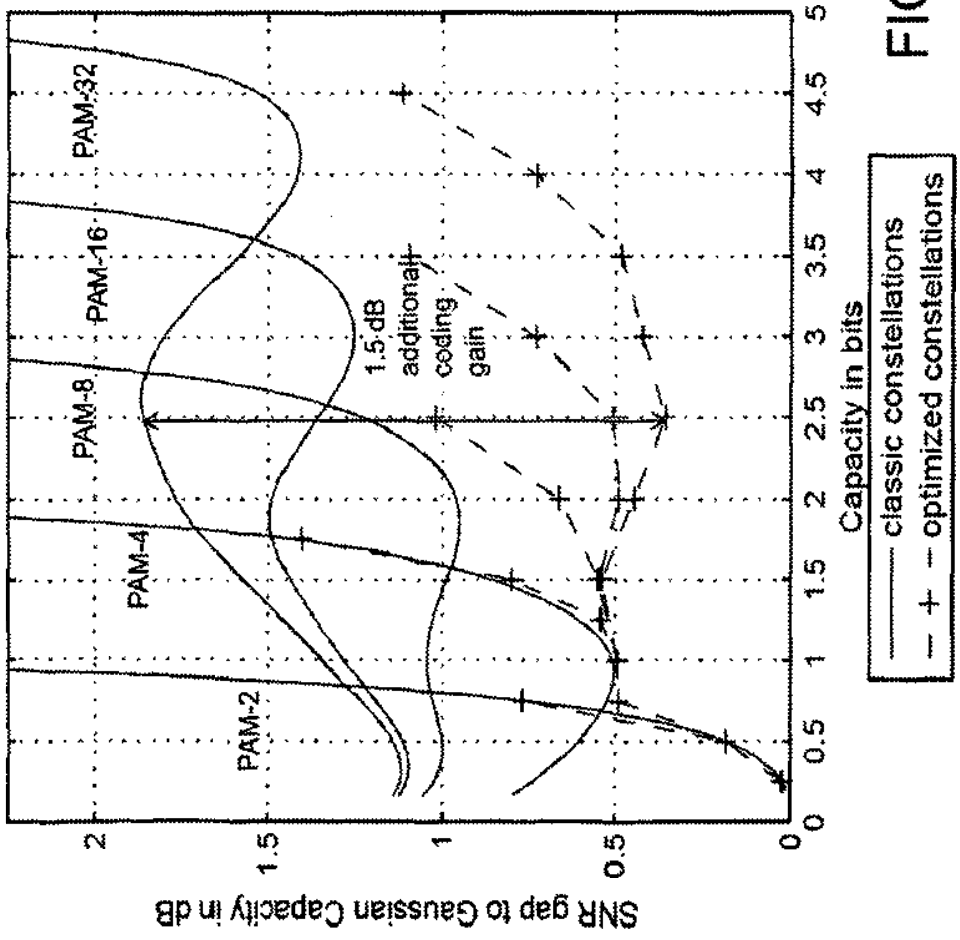

SNR gap to Gaussian Capacity in dB
Capacity in bits
classic constellations
optimized constellations
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
1.5 dB additional coding gain
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8a

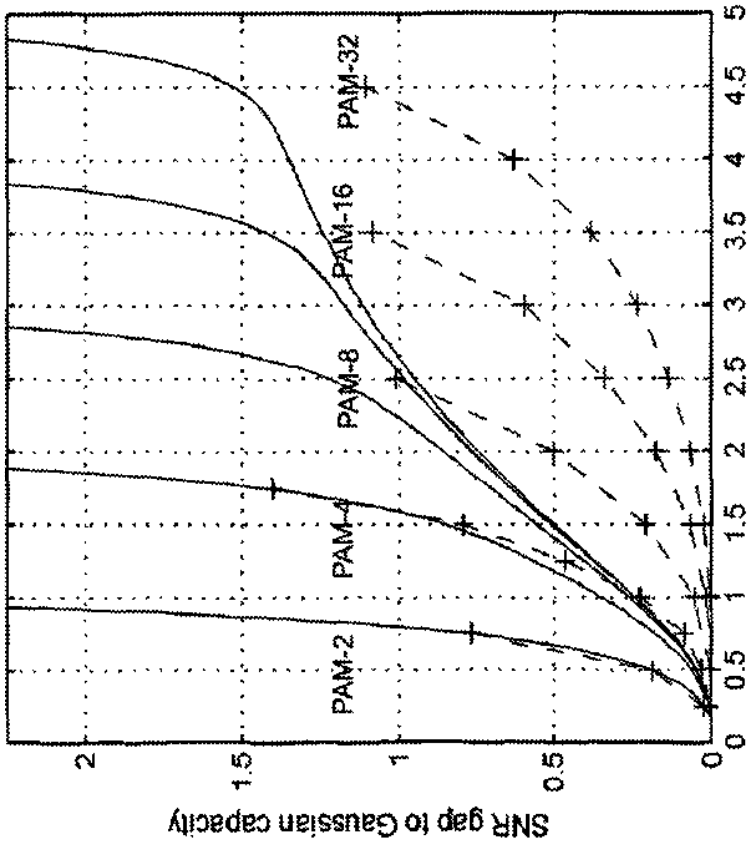

SNR gap to Gaussian capacity
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
classic constellations
optimized constellations
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8b

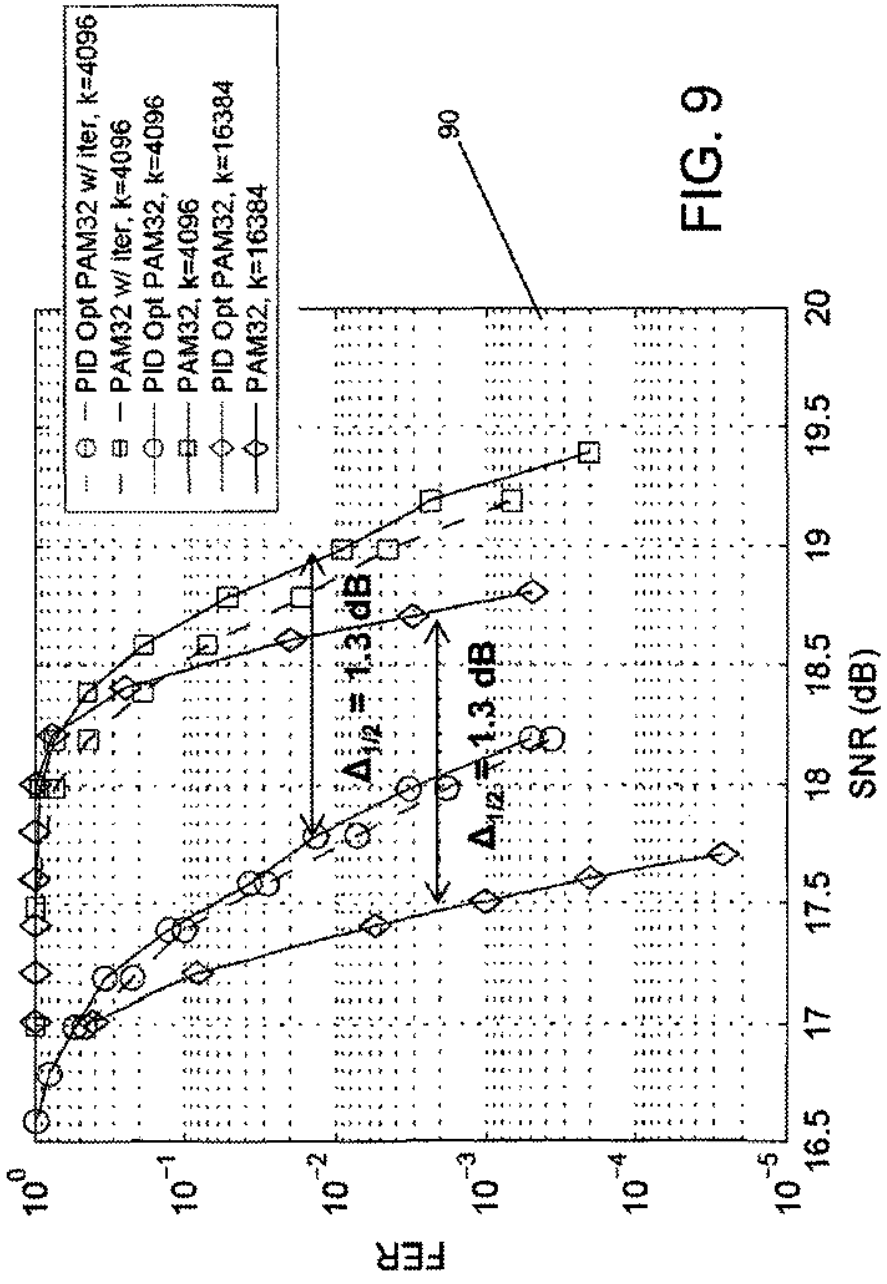

PID Opt PAM32 w/ iter, k=4096
PAM32 w/ iter, k=4096
PID Opt PAM32, k=4096
PAM32, k=4096
PID Opt PAM32, k=16384
PAM32, k=16384
90
$\Delta_{1/2}$ = 1.3 dB
$\Delta_{1/2}$ = 1.3 dB
FER
SNR (dB)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20

FIG. 9

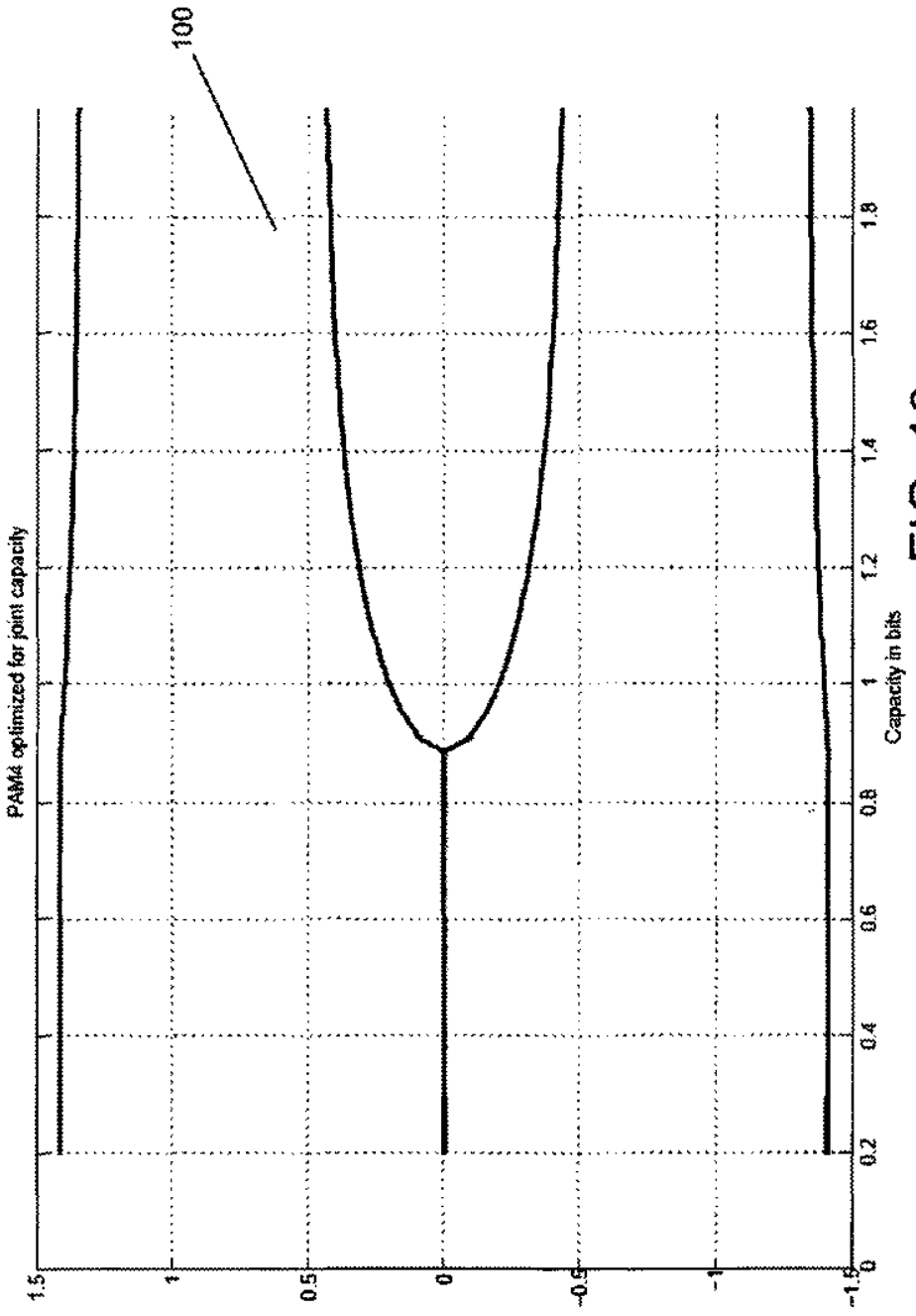

100
PAM4 optimized for joint capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10a

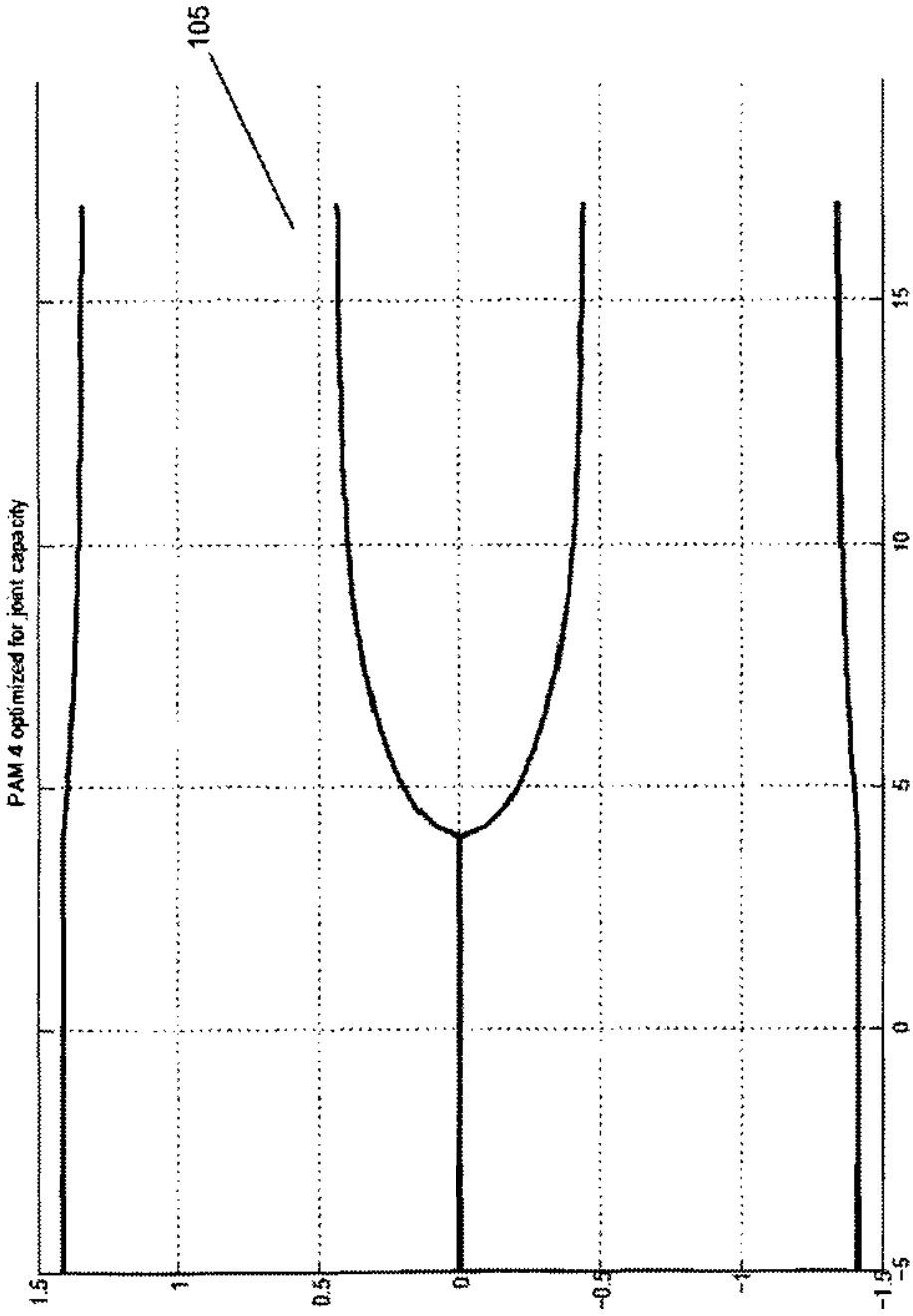

105
PAM 4 optimized for joint capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10b

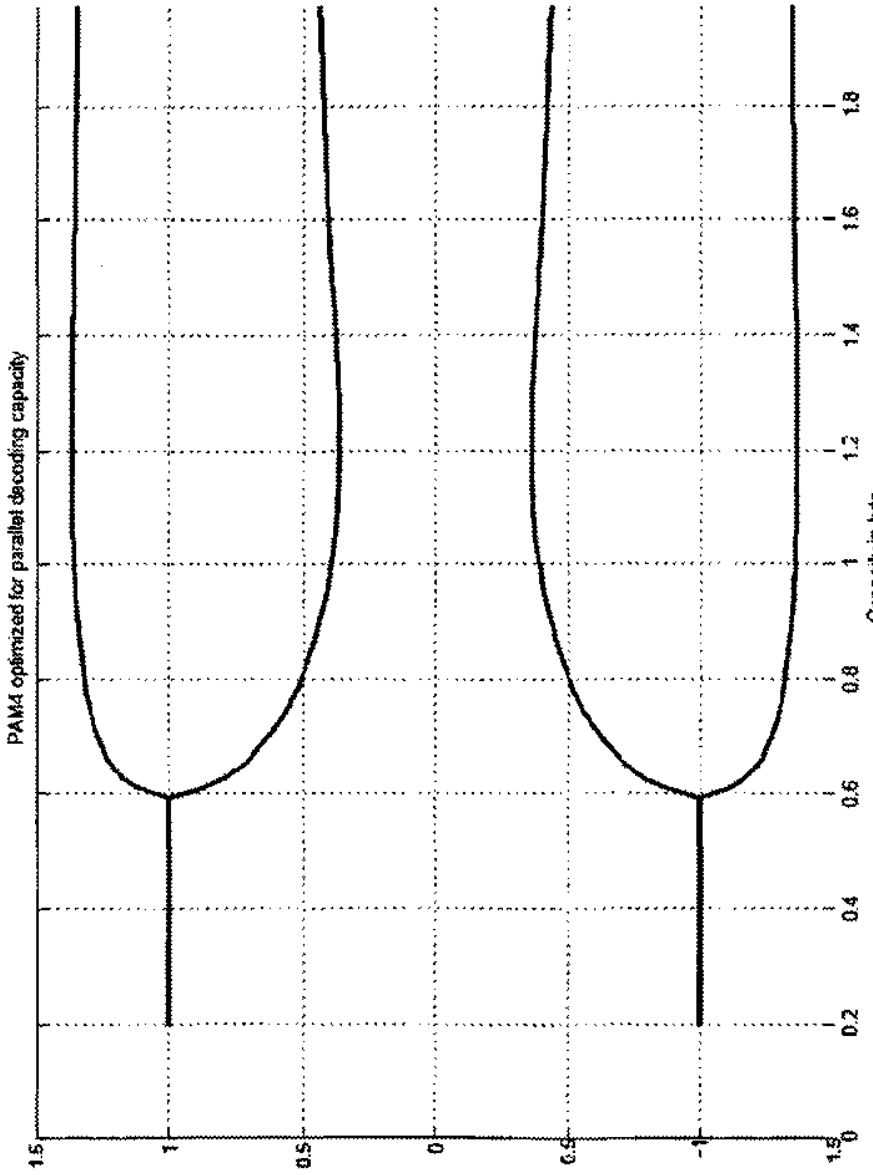

PAM4 optimized for parallel decoding capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10c

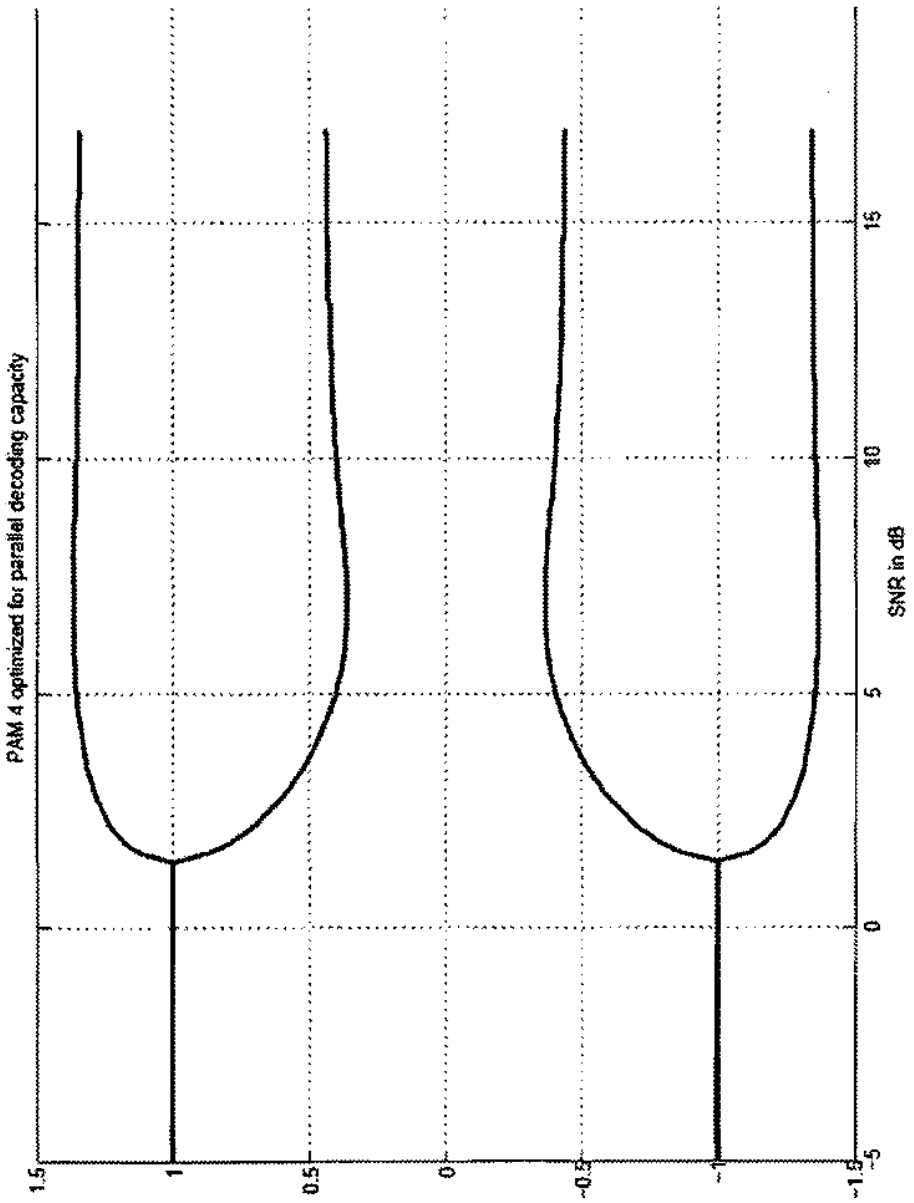

PAM 4 optimized for parallel decoding capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10d

PAM-4 constellations optimized for joint capacity at different rates

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.03	2.71	5.00	7.15	9.24
x_0	-1.41	-1.41	-1.40	-1.37	-1.36
x_1	0.00	0.00	-0.20	-0.33	-0.39
x_2	0.00	0.00	0.20	0.33	0.39
x_3	1.41	1.41	1.40	1.37	1.36

FIG. 11a

PAM-4 constellations optimized for parallel decoding capacity at different

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.19	3.11	5.26	7.22	9.25
x_0	-1.00	-1.30	-1.36	-1.37	-1.36
x_1	-1.00	-0.56	-0.39	-0.33	-0.39
x_2	1.00	1.30	1.36	0.33	1.36
x_3	1.00	0.56	0.39	1.37	0.39

FIG. 11b

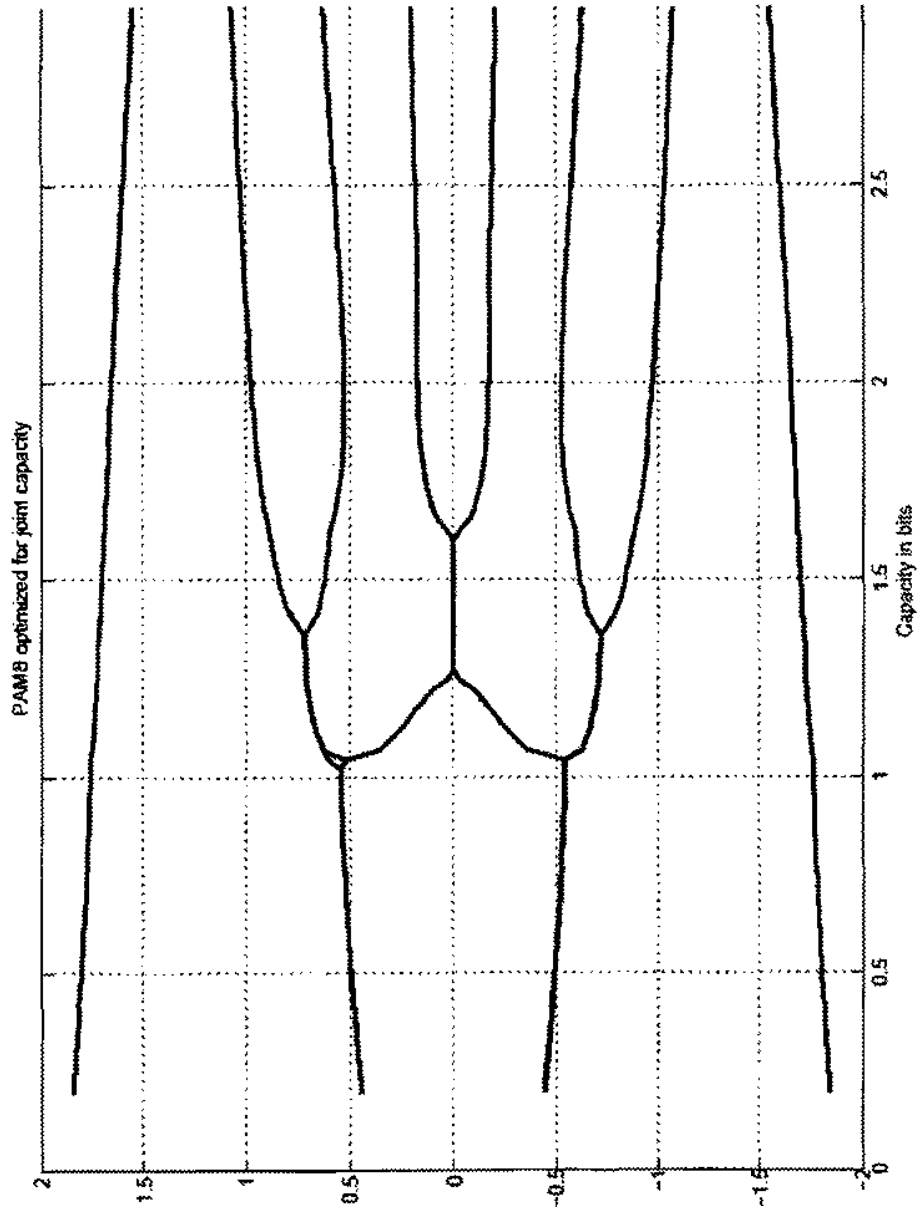

PAM8 optimized for joint capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12a

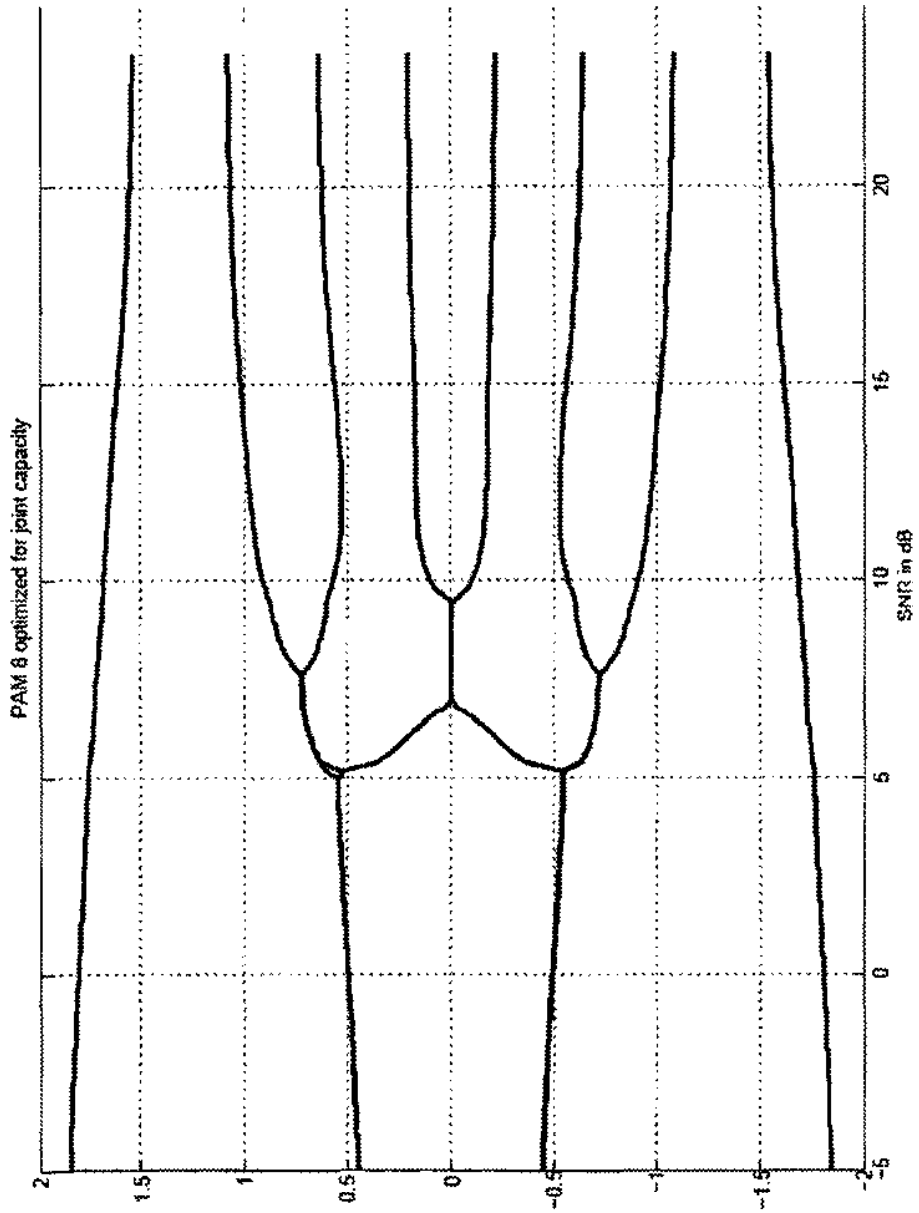

PAM 8 optimized for joint capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12b

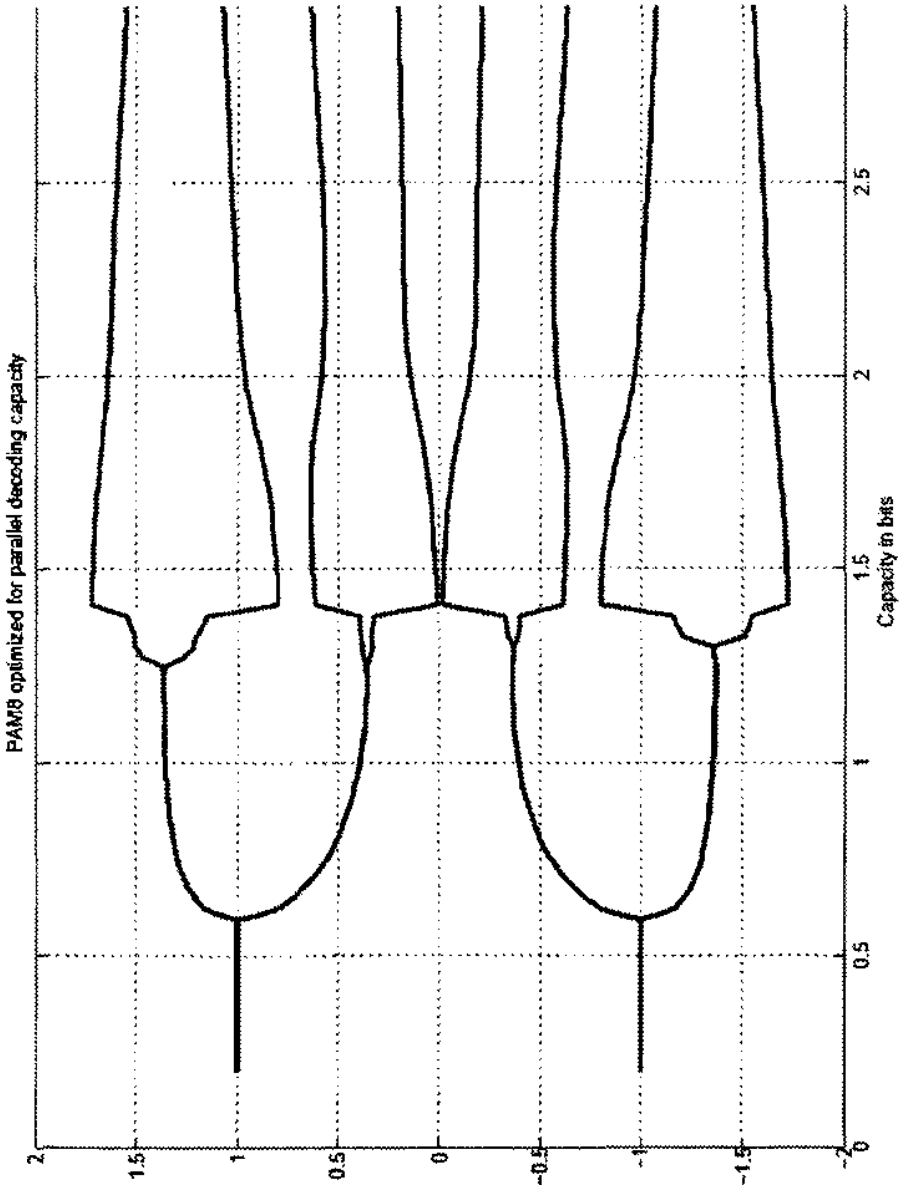

PAM8 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12c

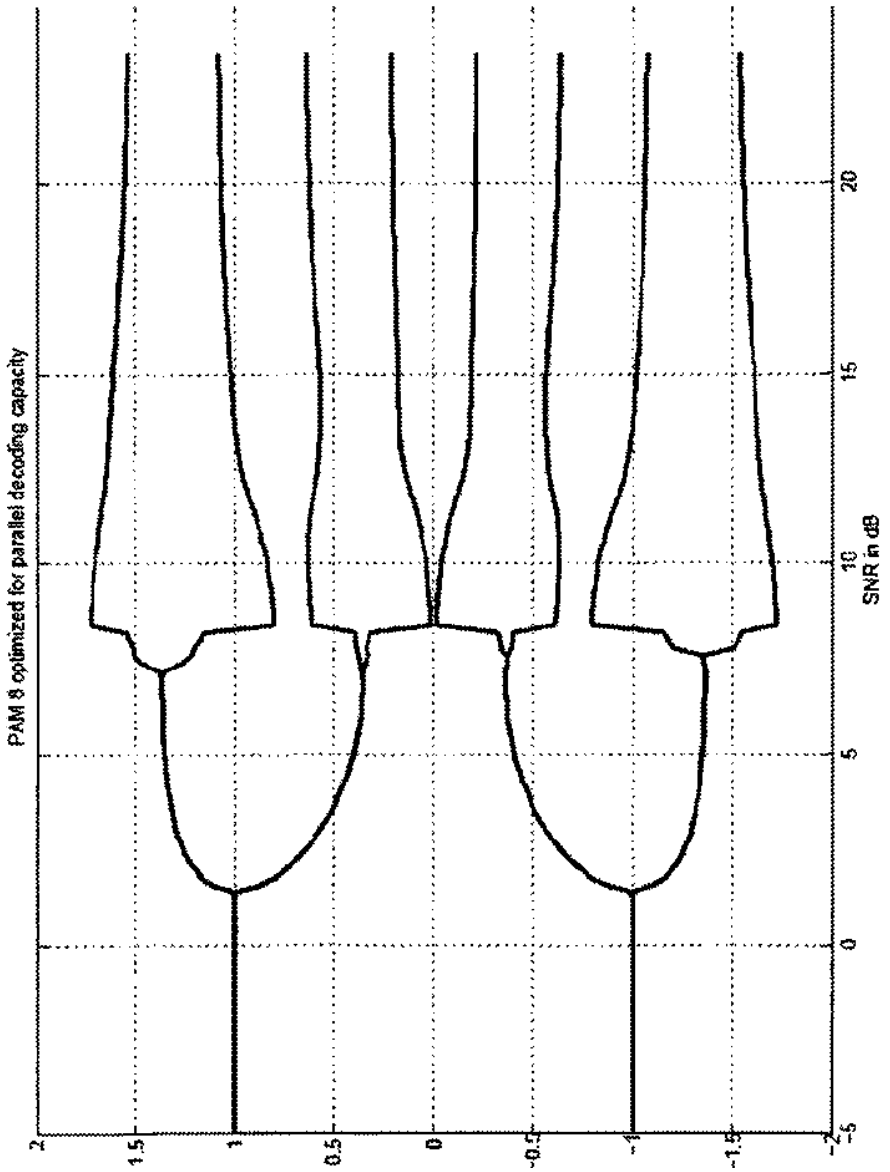

PAM 8 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12d

PAM-8 constellations optimized for joint capacity at different rates

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.00	4.82	8.66	12.26	15.93
x_0	-1.81	-1.76	-1.70	-1.66	-1.60
x_1	-0.50	-0.55	-0.84	-0.97	-1.03
x_2	-0.50	-0.55	-0.63	-0.53	-0.58
x_3	-0.50	-0.55	-0.00	-0.17	-0.19
x_4	0.50	0.55	0.00	0.17	0.19
x_5	0.50	0.55	0.63	0.53	0.58
x_6	0.50	0.55	0.84	0.97	1.03
x_7	1.81	1.76	1.70	1.66	1.60

FIG. 13a

PAM-8 constellations optimized for parallel decoding capacity at different

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.19	5,27	9.00	12.42	15.93
x_0	-1.00	-1.36	-1.72	-1.64	-1.60
x_1	-1.00	-1.36	-0.81	-0.97	-1.03
x_2	-1.00	-0.39	1.72	1.64	-0.19
x_3	-1.00	-0.39	-0.62	-0.58	-0.58
x_4	1.00	1.36	0.62	0.58	1.60
x_5	1.00	1.36	0.02	0.15	1.03
x_6	1.00	0.39	0.81	0.97	0.19
x_7	1.00	0.39	-0.02	-0.15	0.58

FIG. 13b

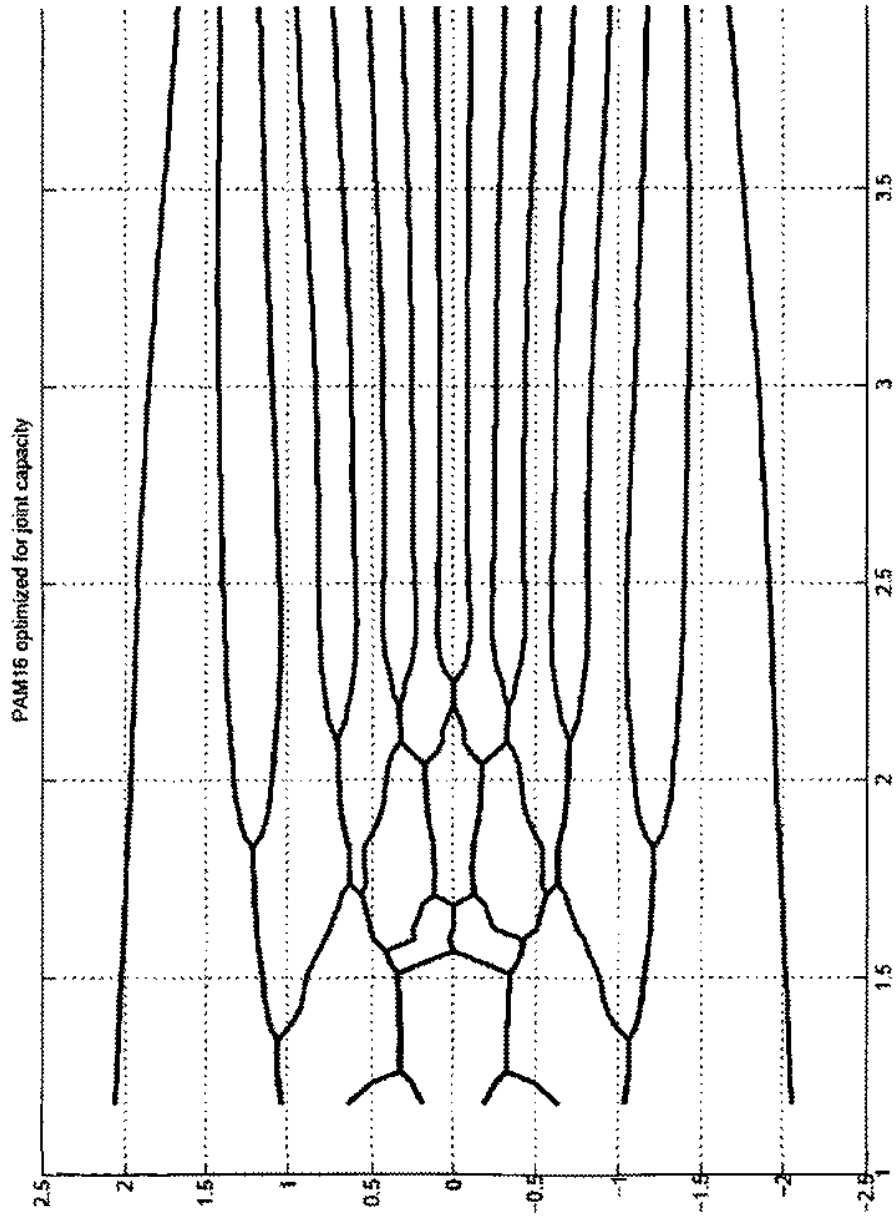

PAM16 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

FIG. 14a

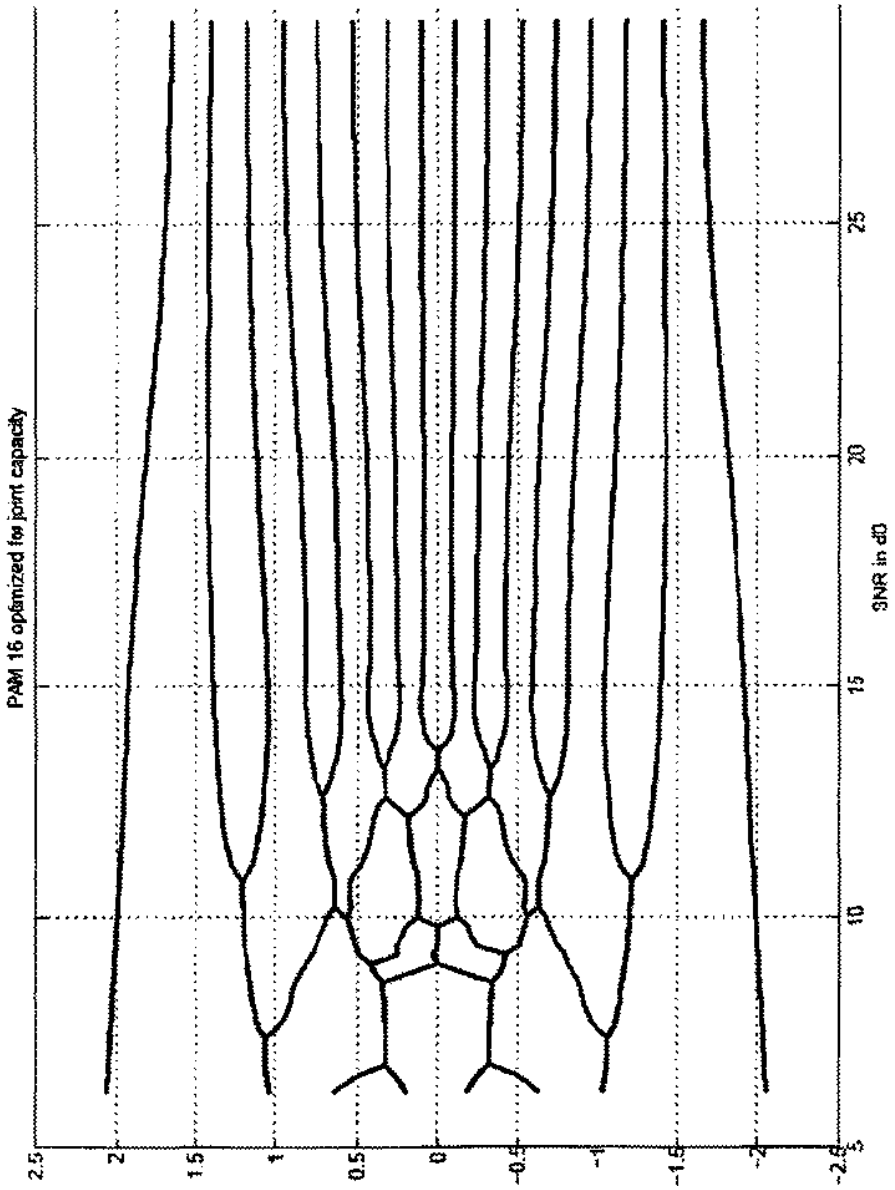

PAM 16 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14b

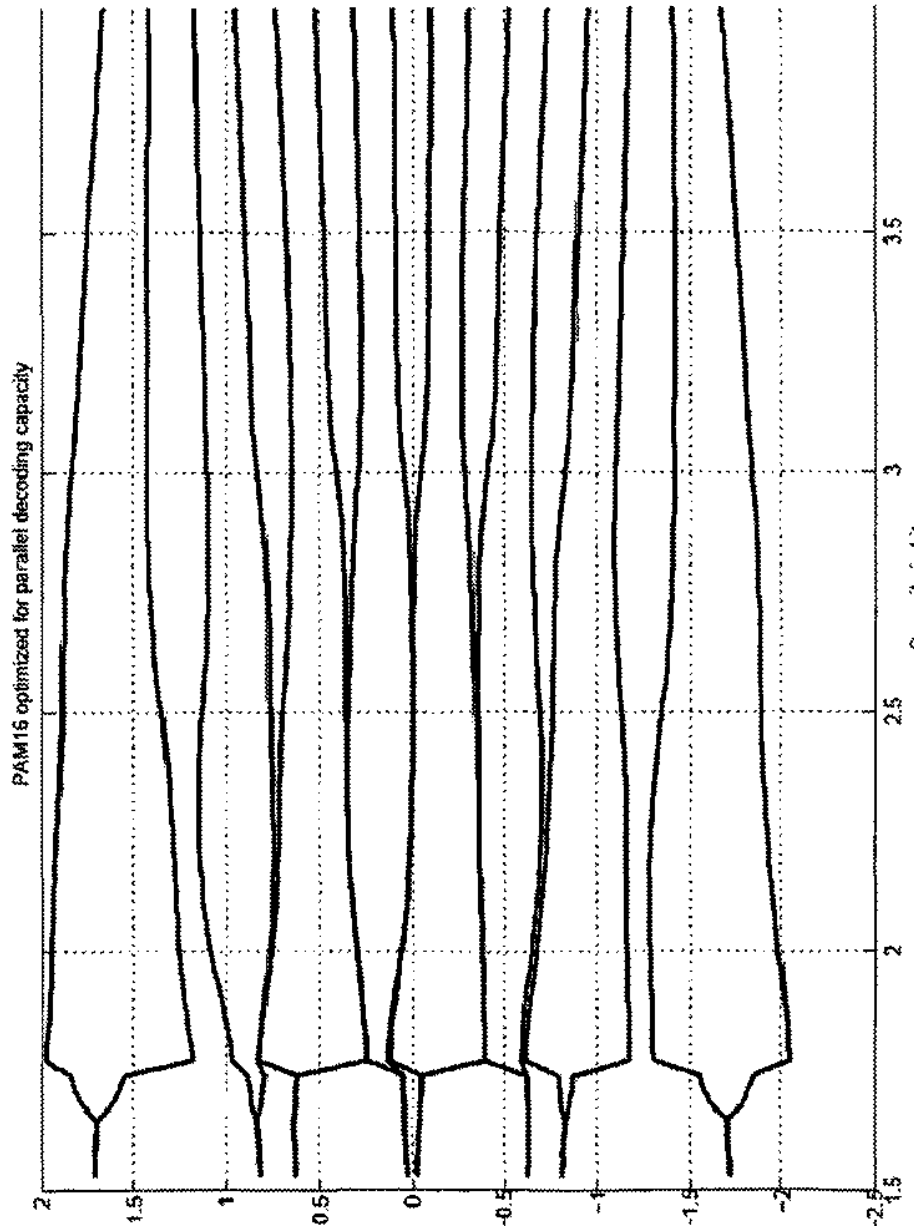

PAM16 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5

FIG. 14c

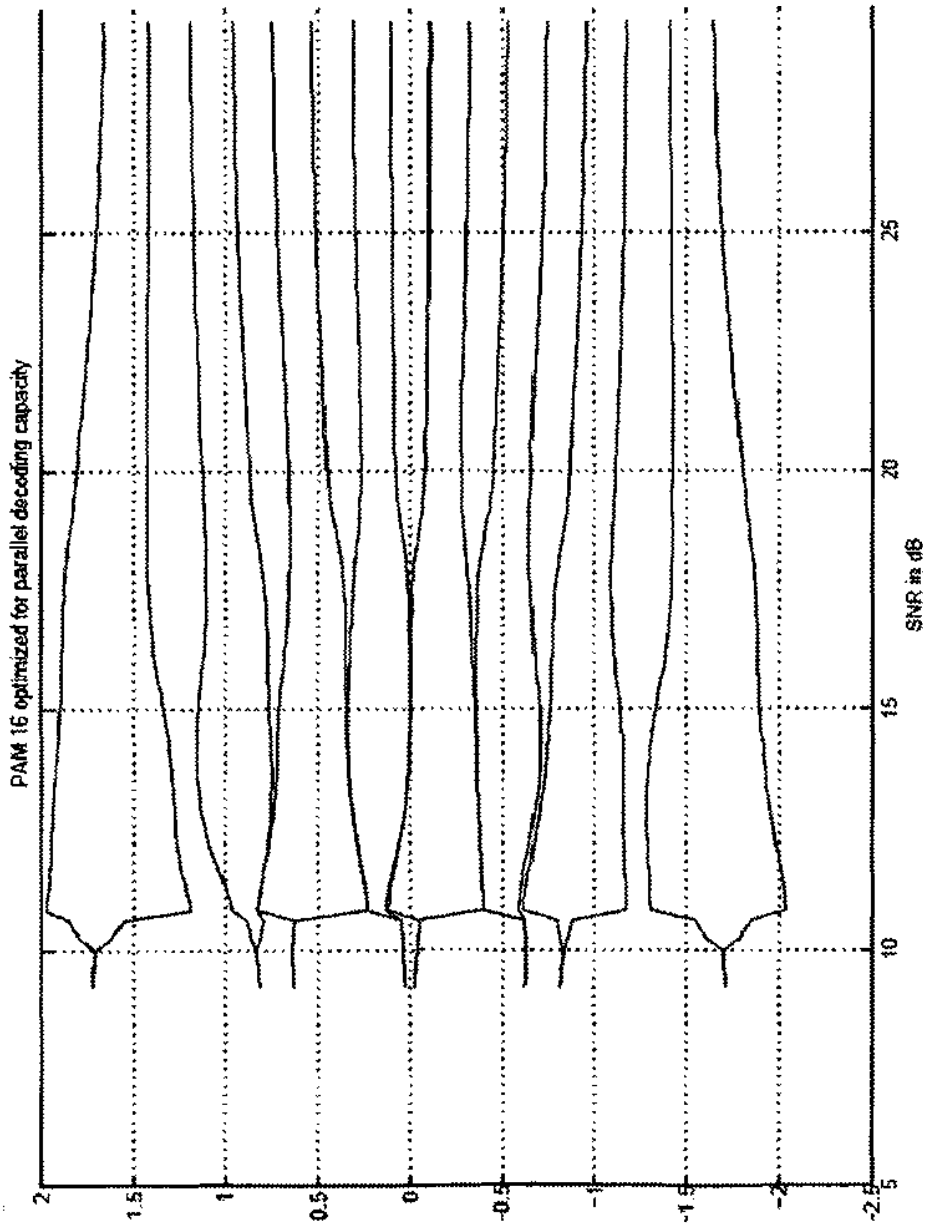

PAM 16 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14d

PAM-16 constellations optimized for joint capacity at different rates

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	8.52	11.94	15.25	18.60	22.12
x_0	-2.02	-1.96	-1.91	-1.85	-1.76
x_1	-1.16	-1.33	-1.40	-1.42	-1.42
x_2	-1.16	-1.10	-1.05	-1.10	-1.15
x_3	-0.90	-0.69	-0.82	-0.84	-0.90
x_4	-0.34	-0.69	-0.60	-0.62	-0.68
x_5	-0.34	-0.40	-0.43	-0.43	-0.47
x_6	-0.34	-0.17	-0.24	-0.26	-0.28
x_7	-0.34	-0.17	-0.09	-0.08	-0.09
x_8	0.34	0.17	0.09	0.08	0.09
x_9	0.34	0.17	0.24	0.26	0.28
x_{10}	0.34	0.40	0.43	0.43	0.47
x_{11}	0.34	0.69	0.60	0.62	0.68
x_{12}	0.90	0.69	0.82	0.84	0.90
x_{13}	1.16	1.10	1.05	1.10	1.15
x_{14}	1.16	1.33	1.40	1.42	1.42
x_{15}	2.02	1.96	1.91	1.85	1.76

FIG. 15a

PAM-16 constellations optimized for parallel decoding capacity at different

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	9.00	12.25	15.42	18.72	22.13
x_0	-1.72	-1.98	-1.89	-1.84	-1.75
x_1	-1.72	-1.29	-1.36	-1.42	-1.42
x_2	-0.81	1.94	1.89	1.84	1.75
x_3	-0.81	-1.17	-1.14	-1.11	-1.15
x_4	1.72	-0.38	-0.35	-0.40	-0.47
x_5	1.72	-0.65	-0.70	-0.65	-0.68
x_6	-0.62	-0.38	-0.34	-0.29	-0.28
x_7	-0.62	-0.68	-0.76	-0.83	-0.90
x_8	0.62	1.09	1.13	1.11	1.15
x_9	0.62	0.76	0.76	0.84	0.90
x_{10}	0.02	1.26	1.35	1.42	1.42
x_{11}	0.02	0.76	0.70	0.65	0.68
x_{12}	0.81	0.06	0.00	0.05	0.09
x_{13}	0.81	0.29	0.34	0.29	0.28
x_{14}	-0.02	0.06	0.00	-0.05	-0.09
x_{15}	-0.02	0.29	0.35	0.40	0.47

FIG. 15b

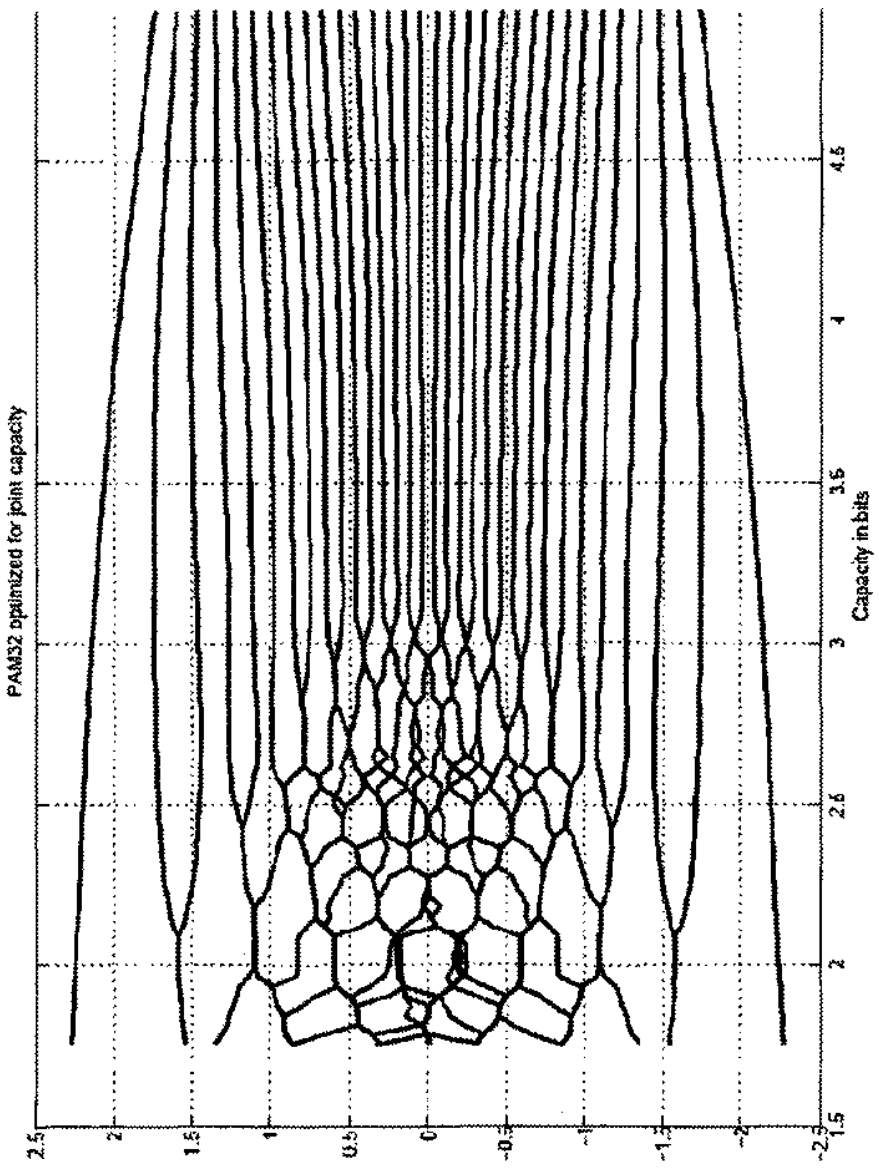

PAM32 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16a

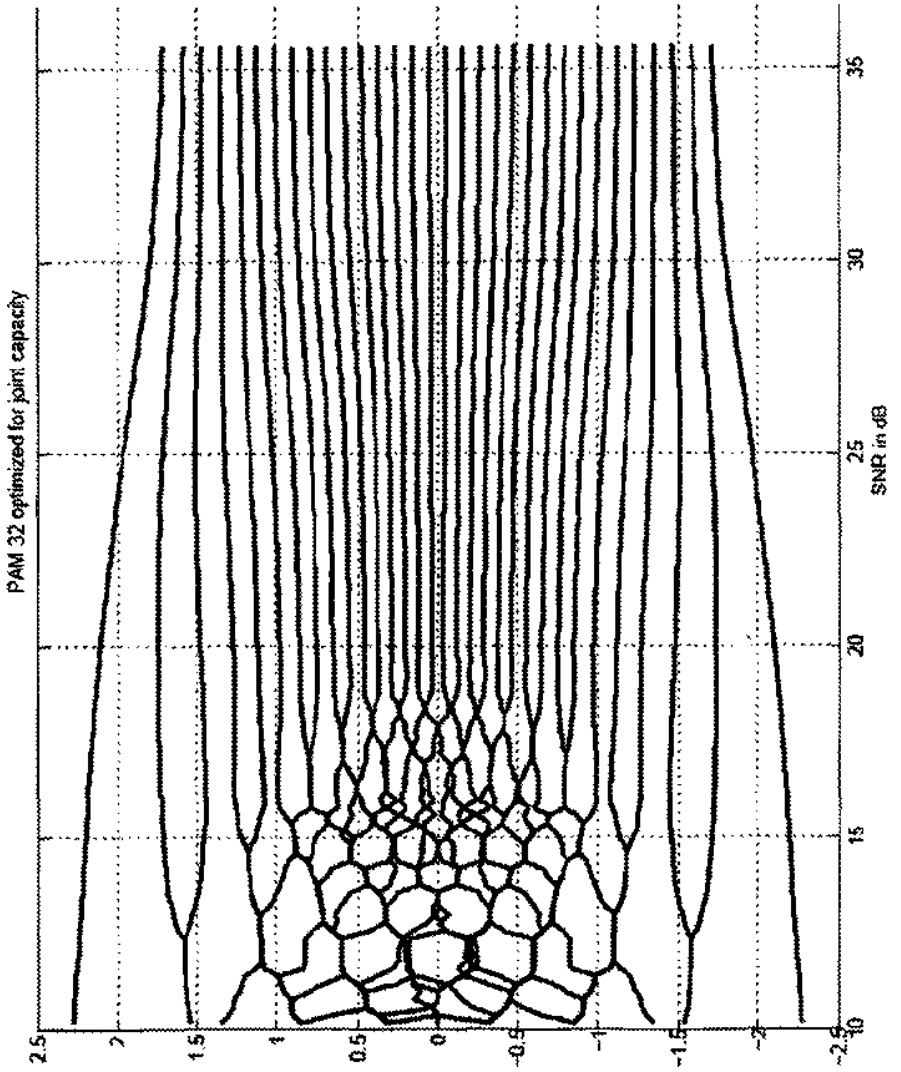

PAM 32 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

FIG. 16b

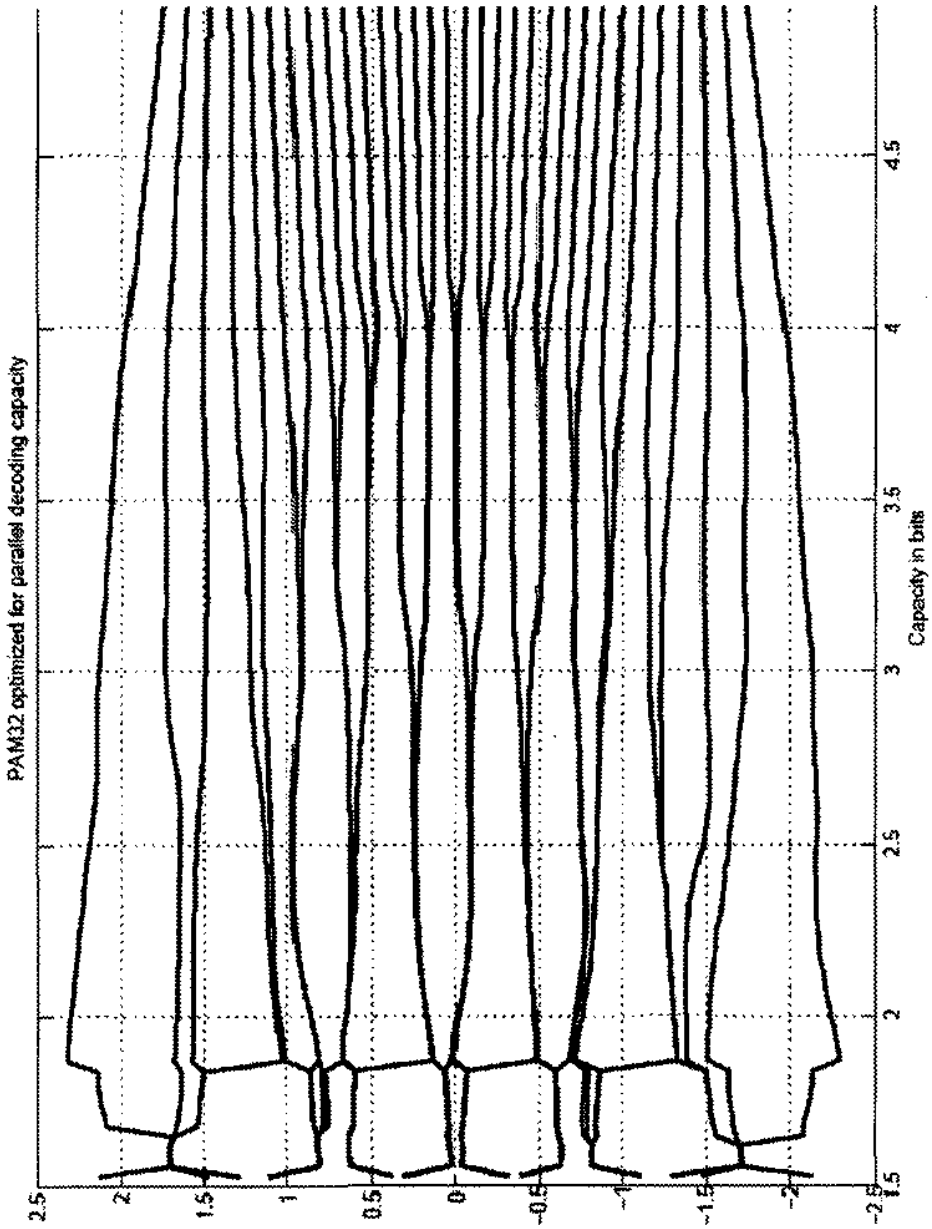

PAM32 optimized for parallel decoding capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16c

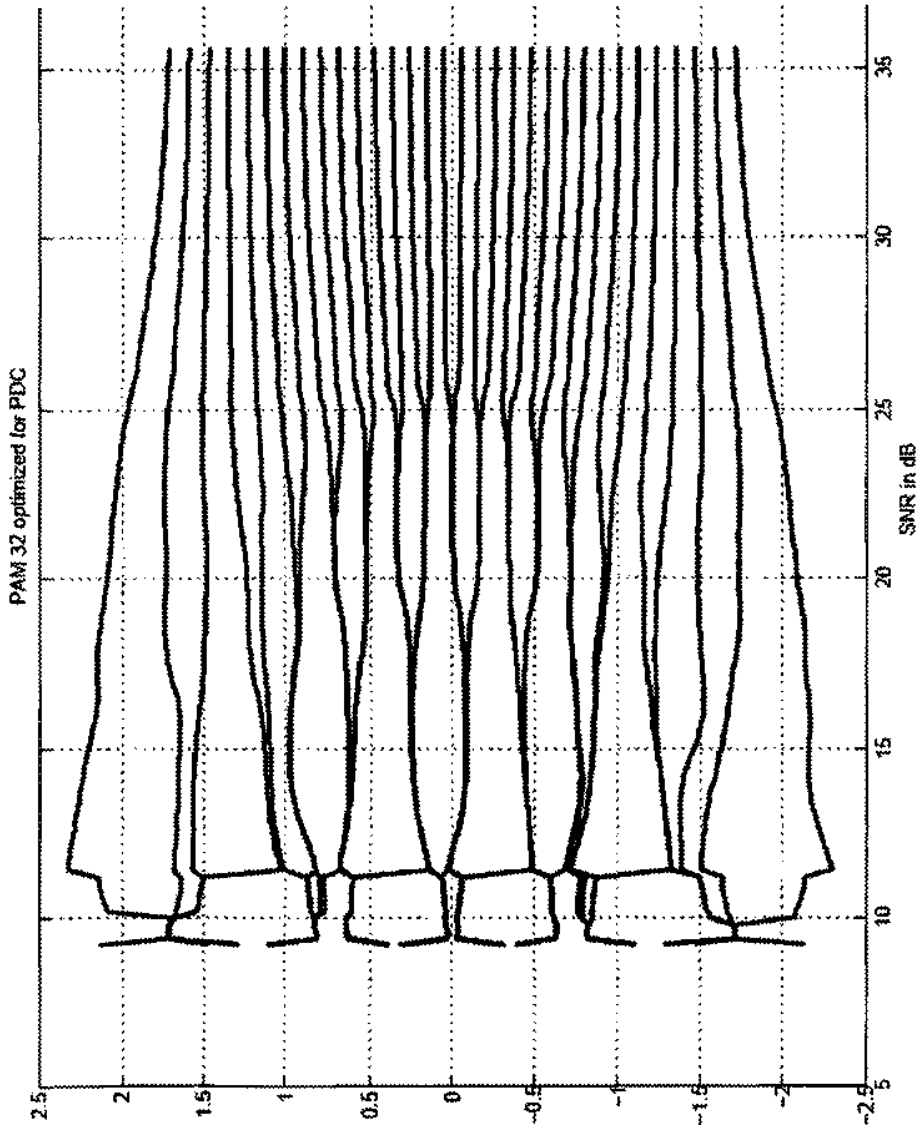

PAM 32 optimized for PDC
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

FIG. 16d

PAM-32 constellations optimized for joint capacity at different rates

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	11.83	15.05	18.23	21.42	24.69	28.19
x_0	-2.25	-2.19	-2.14	-2.07	-1.97	-1.85
x_1	-1.58	-1.71	-1.74	-1.74	-1.72	-1.66
x_2	-1.58	-1.46	-1.46	-1.49	-1.51	-1.50
x_3	-1.10	-1.23	-1.27	-1.29	-1.33	-1.35
x_4	-1.10	-1.13	-1.11	-1.13	-1.17	-1.21
x_5	-1.10	-0.90	-0.98	-0.99	-1.02	-1.08
x_6	-0.83	-0.90	-0.85	-0.87	-0.90	-0.95
x_7	-0.60	-0.75	-0.75	-0.76	-0.78	-0.84
x_8	-0.60	-0.58	-0.63	-0.65	-0.67	-0.73
x_9	-0.60	-0.58	-0.57	-0.56	-0.57	-0.62
x_{10}	-0.60	-0.49	-0.42	-0.46	-0.48	-0.52
x_{11}	-0.24	-0.29	-0.40	-0.38	-0.39	-0.42
x_{12}	-0.21	-0.28	-0.24	-0.29	-0.30	-0.32
x_{13}	-0.20	-0.28	-0.24	-0.21	-0.21	-0.23
x_{14}	-0.20	-0.09	-0.09	-0.12	-0.13	-0.14
x_{15}	-0.16	-0.00	-0.07	-0.04	-0.04	-0.05
x_{16}	0.16	0.00	0.07	0.04	0.04	0.05
x_{17}	0.19	0.09	0.09	0.12	0.13	0.14
x_{18}	0.21	0.28	0.24	0.21	0.21	0.23
x_{19}	0.22	0.28	0.24	0.29	0.30	0.32
x_{20}	0.23	0.28	0.41	0.38	0.39	0.42
x_{21}	0.60	0.49	0.42	0.46	0.48	0.52
x_{22}	0.60	0.58	0.57	0.56	0.57	0.62
x_{23}	0.60	0.58	0.62	0.65	0.67	0.73
x_{24}	0.60	0.75	0.75	0.76	0.78	0.84
x_{25}	0.83	0.90	0.85	0.87	0.90	0.95
x_{26}	1.10	0.90	0.98	0.99	1.02	1.08
x_{27}	1.10	1.13	1.11	1.13	1.17	1.21
x_{28}	1.10	1.23	1.27	1.29	1.33	1.35
x_{29}	1.58	1.46	1.46	1.49	1.51	1.50
x_{30}	1.58	1.71	1.74	1.74	1.72	1.66
x_{31}	2.25	2.19	2.14	2.07	1.97	1.85

FIG. 17a

PAM-32 constellations optimized for parallel decoding capacity at different

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	12.21	15.27	18.42	21.52	24.79	28.20
x_0	-2.25	-2.16	-2.14	-2.05	-1.97	-1.85
x_1	-1.52	-1.64	-1.75	-1.74	-1.72	-1.66
x_2	2.30	2.19	-1.31	2.05	1.97	-1.35
x_3	-1.39	-1.48	-1.43	-1.49	-1.51	-1.49
x_4	1.56	1.54	2.14	-0.96	-1.03	1.85
x_5	-1.31	-1.23	1.75	-1.15	-1.17	1.66
x_6	1.67	1.65	-1.07	-0.91	-0.90	-1.21
x_7	-1.31	-1.24	-1.04	-1.28	-1.33	-1.08
x_8	-0.48	-0.43	-0.36	-0.17	-0.17	-0.42
x_9	-0.72	-0.76	-0.36	-0.34	-0.31	-0.52
x_{10}	-0.48	-0.43	-0.62	-0.17	-0.15	-0.73
x_{11}	-0.73	-0.76	-0.62	-0.34	-0.35	-0.62
x_{12}	-0.48	-0.42	-0.29	-0.71	-0.67	-0.33
x_{13}	-0.76	-0.86	-0.29	-0.52	-0.55	-0.23
x_{14}	-0.48	-0.42	-0.77	-0.72	-0.77	-0.84
x_{15}	-0.76	-0.86	-0.77	-0.52	-0.48	-0.96
x_{16}	0.87	0.98	1.07	1.49	1.51	1.21
x_{17}	0.66	0.63	1.04	1.28	1.33	1.08
x_{18}	0.87	0.98	0.77	1.74	1.72	0.84
x_{19}	0.66	0.63	0.77	1.15	1.17	0.96
x_{20}	1.07	1.13	1.31	0.72	0.77	1.35
x_{21}	0.66	0.59	1.43	0.91	0.90	1.49
x_{22}	1.05	1.10	0.62	0.71	0.67	0.73
x_{23}	0.66	0.60	0.62	0.96	1.03	0.62
x_{24}	-0.01	-0.08	0.02	0.00	0.01	0.05
x_{25}	0.17	0.25	0.02	0.17	0.15	0.14
x_{26}	-0.01	-0.08	0.29	0.00	-0.01	0.33
x_{27}	0.17	0.25	0.29	0.17	0.17	0.23
x_{28}	-0.01	-0.08	-0.02	0.52	0.48	-0.05
x_{29}	0.17	0.25	-0.02	0.34	0.35	-0.14
x_{30}	-0.01	-0.08	0.36	0.52	0.55	0.42
x_{31}	0.17	0.25	0.36	0.34	0.31	0.52

FIG. 17b

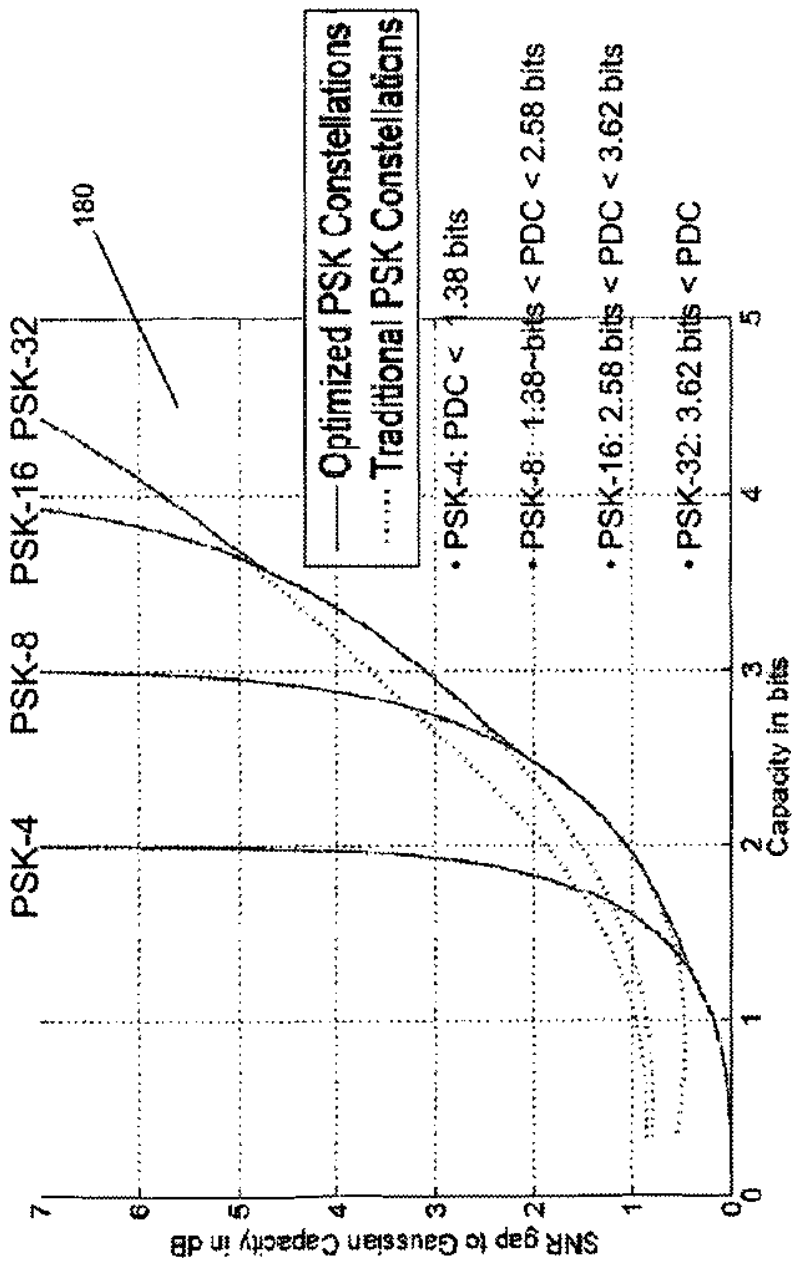

180
PSK-4
PSK-8
PSK-16
PSK-32
Optimized PSK Constellations
Traditional PSK Constellations
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38~bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
SNR gap to Gaussian Capacity in dB
Capacity in bits
0
1
2
3
4
5
6
7

FIG. 18

Points location in degrees
0
50
100
150
200
250
300
350
SNR in dB
0
2
4
6
8
10
12
14
16
18
20

FIG. 19

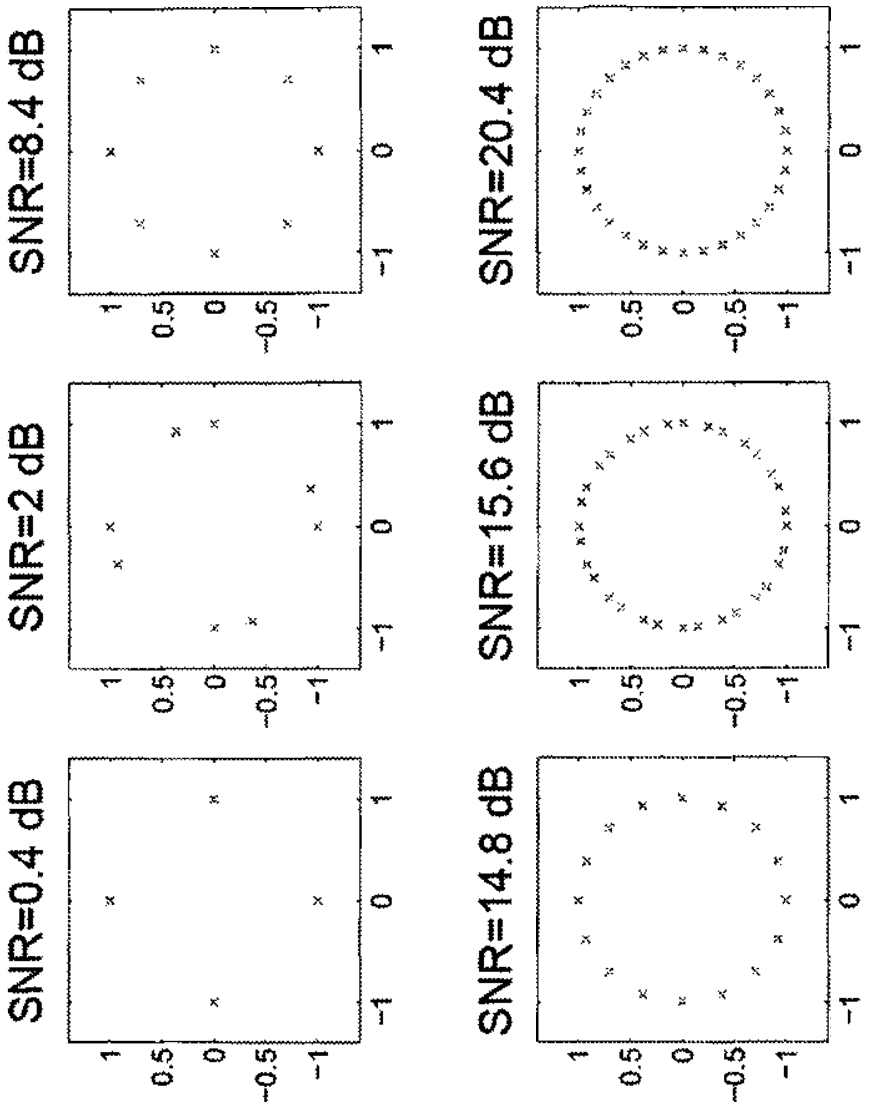


FIG. 20

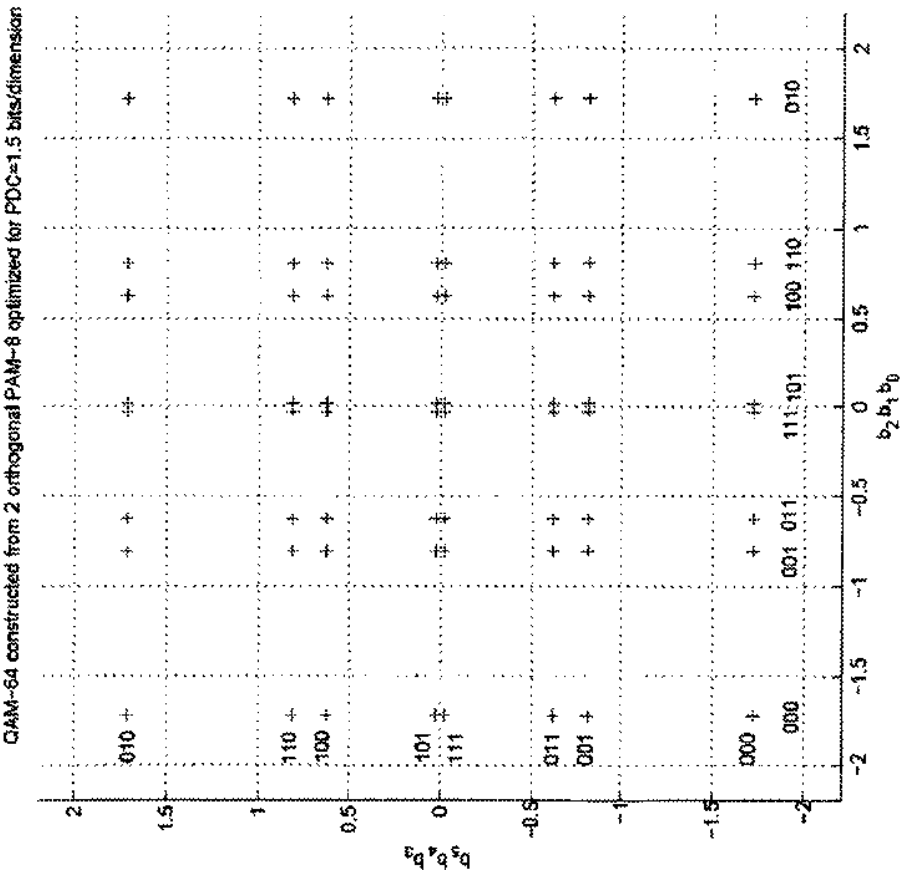

QAM-64 constructed from 2 orthogonal PAM-8 optimized for PDC=1.5 bits/dimension
$b_5 b_4 b_3$
$b_2 b_1 b_0$
010
110
100
101
111
011
001
000
001
011
111
101
100
110
010

FIG. 21

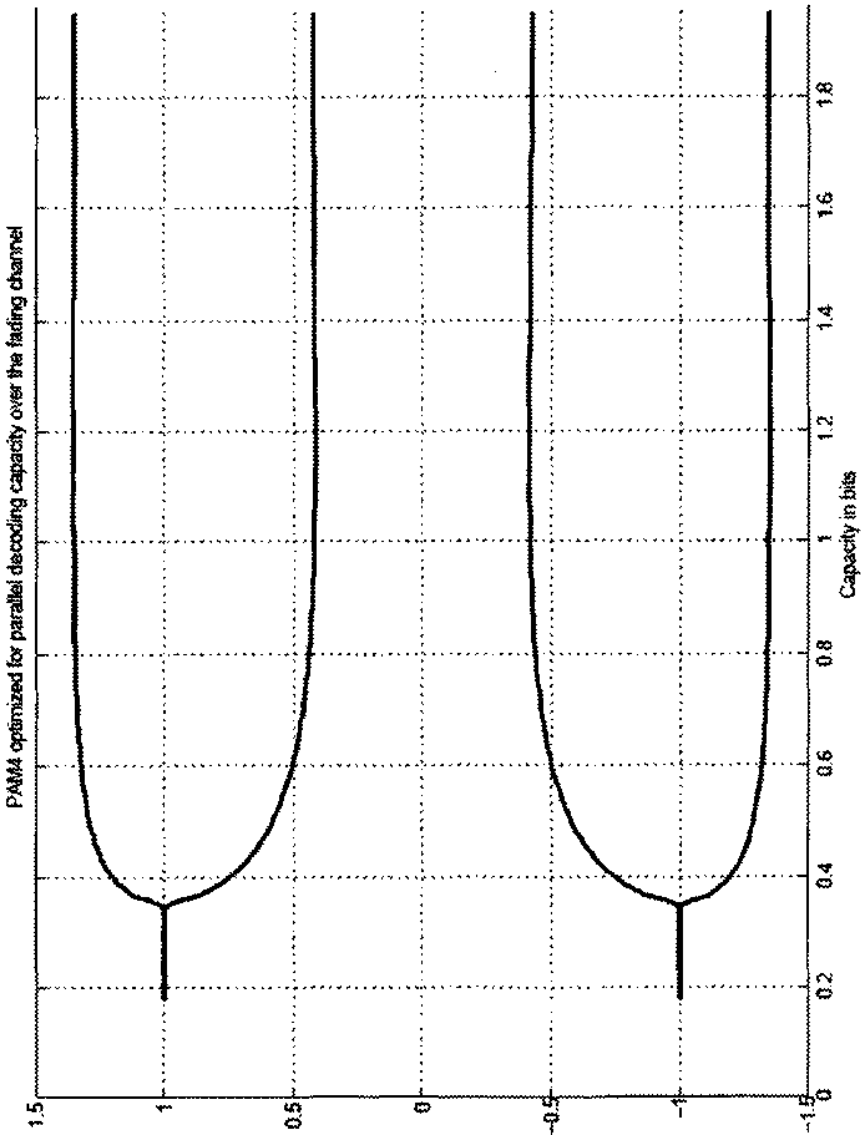

PAM4 optimized for parallel decoding capacity over the fading channel
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 22a

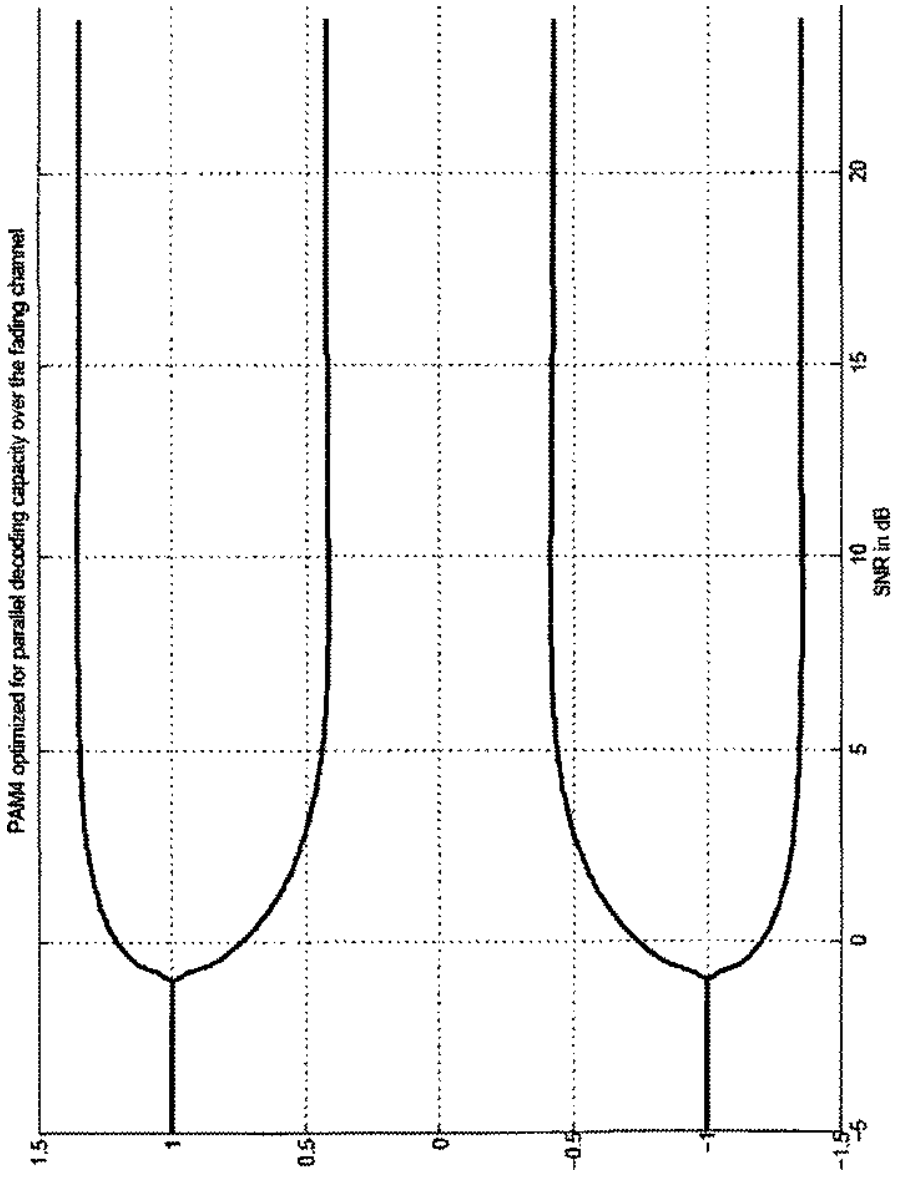

PAM4 optimized for parallel decoding capacity over the fading channel
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20

FIG. 22b

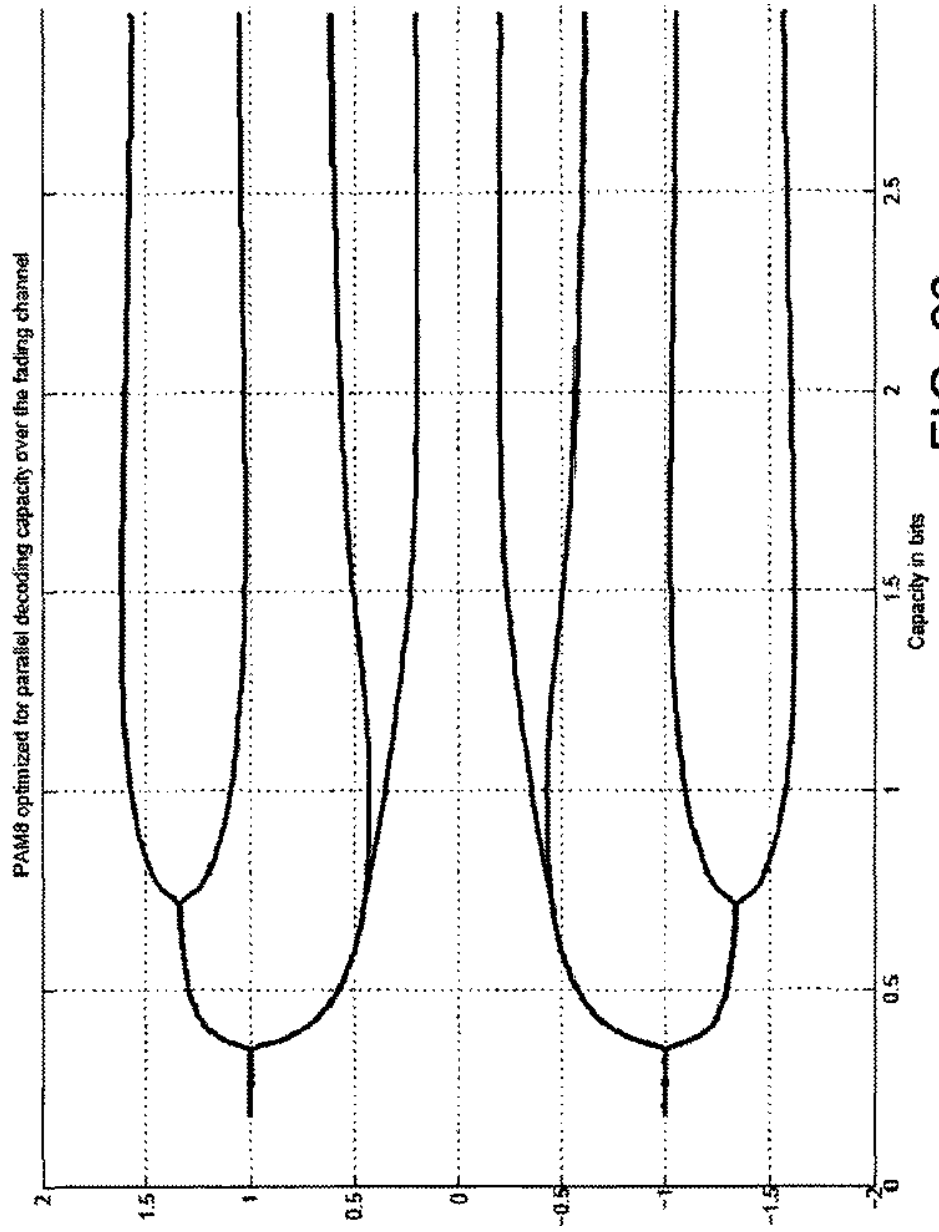

PAM8 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 23a

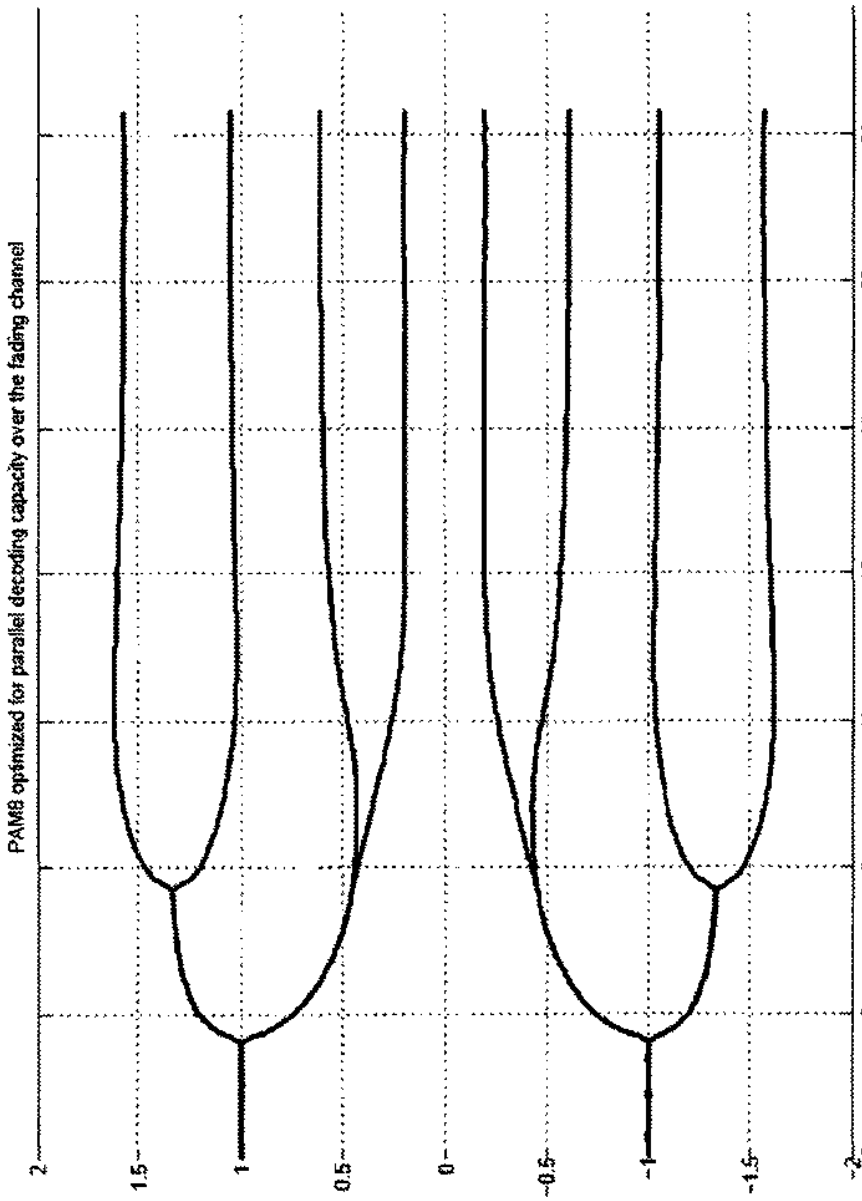

PAM8 optimized for parallel decoding capacity over the fading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30

FIG. 23b

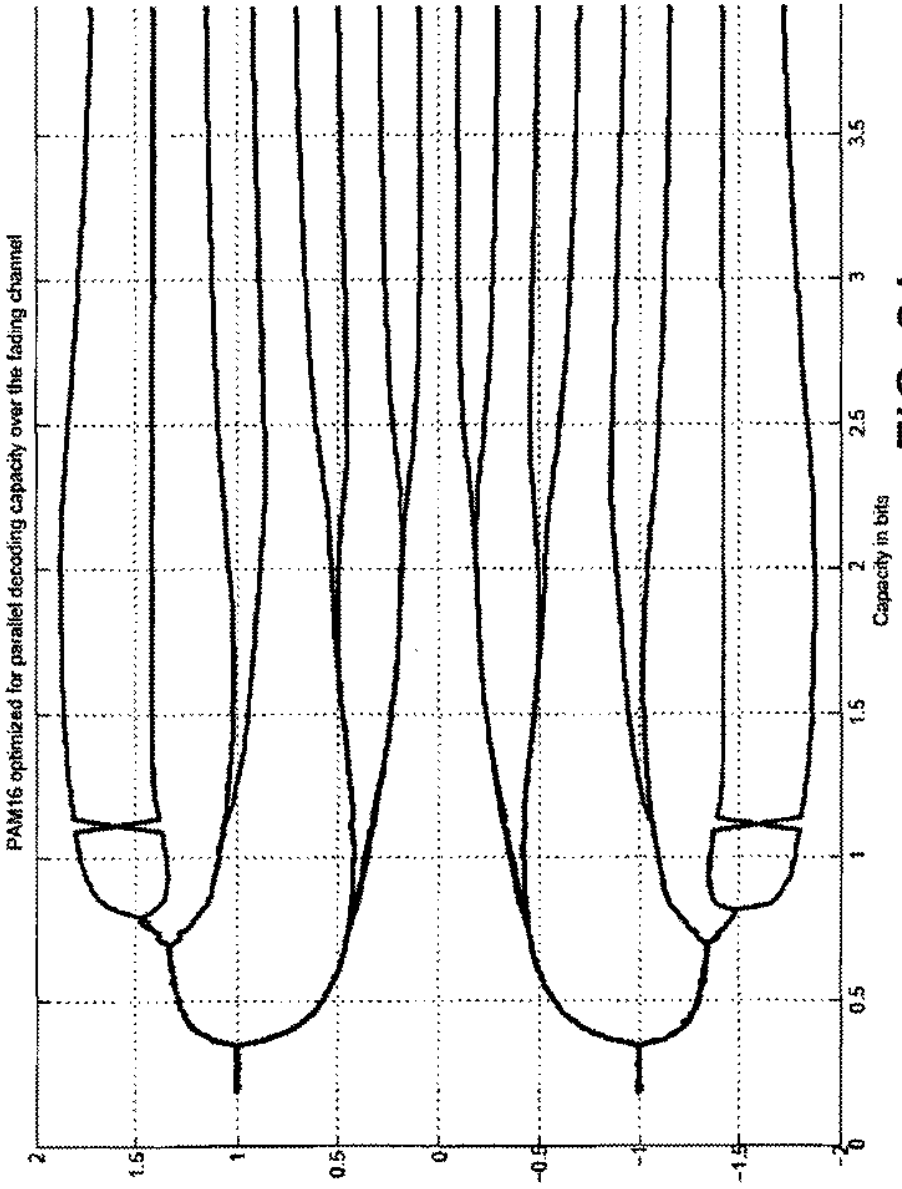

PAM16 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

FIG. 24a

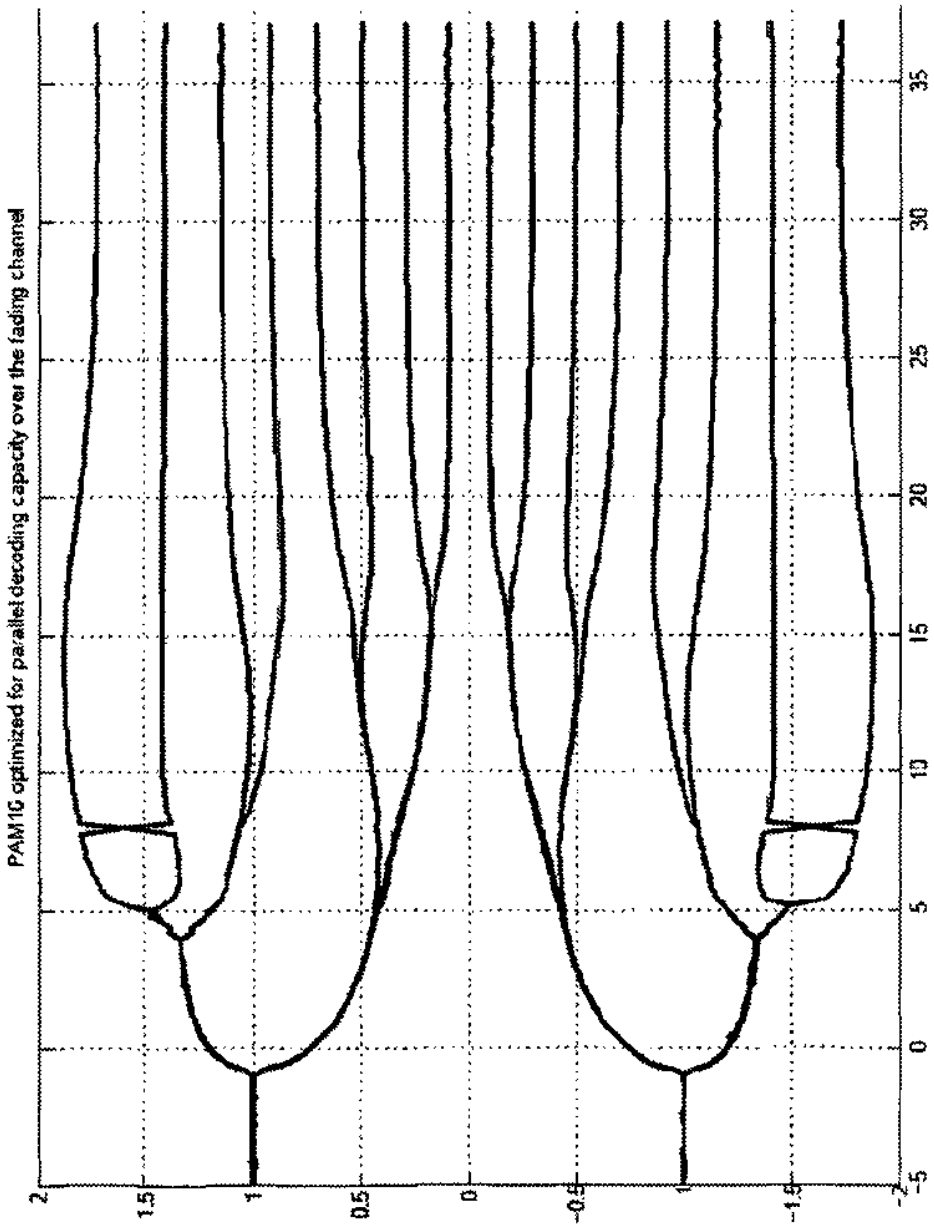

PAM16 optimized for parallel decoding capacity over the fading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
35
30
25
20
15
10
5
0
-5

FIG. 24b

REFERENCES CITED IN THE DESCRIPTION

This list of references cited by the applicant is for the reader's convenience only. It does not form part of the European patent document. Even though great care has been taken in compiling the references, errors or omissions cannot be excluded and the EPO disclaims all liability in this regard.

Non-patent literature cited in the description

- **D. SOMMER ; G. P. FETTWEIS.** Signal shaping by non-uniform QAM for AWGN channels and applications using turbo coding. *Proc. 2000 ITG Conf. Source and Channel Coding,* 2000, 81-86 **[0006]**
- **R. DE GAUDENZI ; A. GUILLEN I FABREGAS ; A. MARTINEZ.** Performance analysis of turbo-coded APSK modulations over nonlinear satellite channels. *IEEE Transactions on Wireless Communications,* September 2006, vol. 5 (9), 2396-2407 **[0006]**

(11) TR 2019 05158 T4

(12) AVRUPA PATENT FASİKÜLÜ TÜRKÇE ÇEVİRİSİ

(21) Başvuru Numarası
2019/05158

(51) Buluşun tasnif sınıfları
H04J 11/00 ***H04L 1/00*** ***H04L 27/34***

(96) Başvuru Tarihi
2008/06/05

(96) EP Başvuru No
EP08795885.6

(97) EP Yayın No
EP2153561B1

(97) EP Yayın Tarihi
2019/01/09

(30) Rüçhan Bilgileri (32) (33) (31)
2007/06/05 US 933319 P

(74) Vekil
FİLİZ ARK (MUMCUOĞLU VE ARK PATENT VE MARKA DAN. LTD. ŞTİ.)
MUTLUKENT MAH. MUTLUKÖY SİTESİ 1943. CAD. NO:10 06800 - ÇANKAYA/ANKARA

(73) Patent Sahibi
CONSTELLATION DESIGNS, INC.
1563 Calle Patricia Pacific Palisades, Ca 90272 / A.B.D.

(72) Buluşu Yapanlar
CHRISTOPHER RICHARD JONES
1563 Calle Patricia Pacific Palisades, Ca 90272 / A.B.D.

MAGED F. BARSOUM
1468 Kinsport Lane San Jose, Ca 95120 / A.B.D.

(54) Buluş Başlığı

Kapasitesi optimize edilmiş yıldız kümeleri ile sinyalleme yöntemi ve aparatı.

(57) Özet

Benzeri bir SNR bandı içinde çalışan klasik yıldız kümeleri ile karşılaştırıldığında artan kapasiteye sahip geometrik olarak şekillendirilen yıldız kümeleri kullanan iletişim sistemleri açıklanır. Birkaç düzenlemede, geometrik olarak şekillendirilen yıldız kümesi, paralel kod çözme kapasitesi veya ortak kapasite gibi bir kapasite ölçümüne göre optimize edilir. Birçok düzenlemede, bir iletişim sistemi içindeki vericiler ve alıcılara yönelik bir üretici yazılımı yükseltmesinin parçası olarak bir klasik yıldız kümesinin yerini almak üzere bir kapasitesi optimize edilmiş geometrik olarak şekillendirilen yıldız kümesi kullanılabilir. Bir dizi düzenlemede, geometrik olarak şekillendirilen yıldız kümesi bir Toplanır Beyaz Gauss Gürültü kanalı veya bir sönümlü kanal için optimize edilir. Birçok düzenlemede, iletişim uyarlanabilir oran kodlamasını kullanır ve kod oranı değiştikçe geometrik olarak şekillendirilen yıldız kümesi içindeki noktaların konumu değişir. Buluşun bir düzenlemesi bir iletişim kanalı aracılığıyla bir alıcıya sinyalleri iletmek üzere konfigüre edilen bir verici içerir, burada verici, genişletilmiş bir çıktı kodlanmış bit oranında kullanıcı bitleri ve çıktı kodlanan bitleri almak üzere konfigüre edilen bir kodlayıcıyı, bir simge yıldız kümesinde simgeler ile kodlanmış bitleri eşleştirmek üzere konfigüre edilen bir eşleştiriciyi, eşleştirici tarafından oluşturulan simgeleri kullanarak iletişim kanalı aracılığıyla iletime yönelik bir sinyali oluşturmak üzere konfigüre edilen bir modülatörü içerir. Ayrıca, alıcı iletişim kanalı aracılığıyla alınan sinyali demodüle etmek üzere konfigüre edilen bir demodülatör, demodüle edilen sinyalden olasılıkları tahmin etmek üzere konfigüre edilen bir eşleştirme kaldırıcı, eşleştirme kaldırıcı tarafından oluşturulan olasılıklardan kodu çözülmüş bitleri tahmin etmek üzere konfigüre edilen bir kod çözücü içerir. Ayrıca, simge yıldız kümesi, dmin değerini maksimize eden bir sinyal yıldız kümesi ile karşılaştırıldığında azalan bir sinyal-gürültü oranında belirli bir kapasite sağlayan bir kapasitesi optimize edilmiş geometrik olarak aralıklı simge yıldız kümesidir.

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

PCT

(43) International Publication Date
11 December 2008 (11.12.2008)

(10) International Publication Number
WO 2008/151308 A1

(51) International Patent Classification:
H04J 11/00 (2006.01)

(21) International Application Number:
PCT/US2008/065994

(22) International Filing Date: 5 June 2008 (05.06.2008)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
60/933,519 5 June 2007 (05.06.2007) US

(71) Applicants and
(72) Inventors: **BARSOUM, Maged, F.** [US/US]; 374 Durango Aisle, Irvine, CA 92606 (US). **JONES, Christopher, Richard** [US/US]; 1564 Calle Patricia, Pacific Palisades, CA 90272 (US).

(74) Agent: **BAILEY, David, L.**; Kaufi, Pomeroy, Peck & Bailey LLP, P.O. Box 19152, Irvine, CA 92623 (US).

(81) Designated States *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, SV, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.

(84) Designated States *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HR, HU, IE, IS, IT, LT, LU, LV, MC, MT, NL, NO, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report*
— *before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments*

(54) Title: DESIGN METHODOLOGY AND METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS

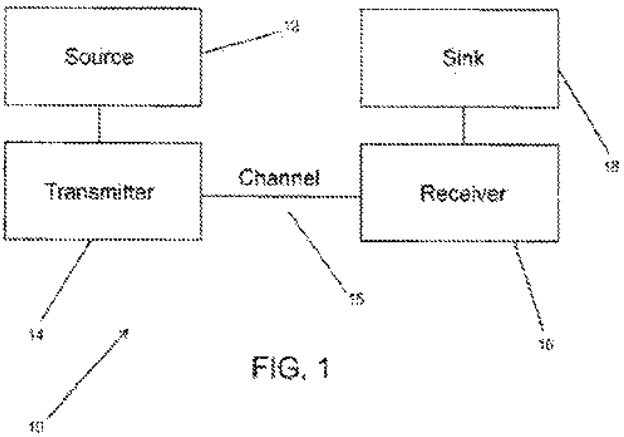


FIG. 1

(57) Abstract: Communication systems are described that use geometrically shaped constellations that have increased capacity compared to conventional constellations operating within a similar SNR band. In several embodiments, the geometrically shaped is optimized based upon a capacity measure such as parallel decoding capacity or joint capacity. In many embodiments, a capacity optimized geometrically shaped constellation can be used to replace a conventional constellation as part of a firmware upgrade to transmitters and receivers within a communication system. In a number of embodiments, the geometrically shaped constellation is optimized for an Additive White Gaussian Noise channel or a fading channel. In numerous embodiments, the communication uses adaptive rate encoding and the location of points within the geometrically shaped constellation changes as the code rate changes. One embodiment of the invention includes a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter includes a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate, a mapper configured to map encoded bits to symbols in a symbol constellation, a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper. In addition, the receiver includes a demodulator configured to demodulate the received signal via the communication channel, a demapper configured to estimate likelihoods from the demodulated signal, a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper. Furthermore, the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

WO 2008/151308 A1

Design Methodology and Method and Apparatus for Signaling with Capacity Optimized Constellations

STATEMENT OF FEDERALLY SPONSORED RESEARCH

[0001] This invention was made with Government support under contract NAS7-03001 awarded by NASA. The Government has certain rights in this invention.

BACKGROUND

[0002] The present invention generally relates to bandwidth and/or power efficient digital transmission systems and more specifically to the use of unequally spaced constellations having increased capacity.

[0003] The term "constellation" is used to describe the possible symbols that can be transmitted by a typical digital communication system. A receiver attempts to detect the symbols that were transmitted by mapping a received signal to the constellation. The minimum distance (d_{min}) between constellation points is indicative of the capacity of a constellation at high signal-to-noise ratios (SNRs). Therefore, constellations used in many communication systems are designed to maximize d_{min}. Increasing the dimensionality of a constellation allows larger minimum distance for constant constellation energy per dimension. Therefore, a number of multi-dimensional constellations with good minimum distance properties have been designed.

[0004] Communication systems have a theoretical maximum capacity, which is known as the Shannon limit. Many communication systems attempt to use codes to increase the capacity of a communication channel. Significant coding gains have been achieved using coding techniques such as turbo codes and LDPC codes. The coding gains achievable using any coding technique are limited by the constellation of the communication system. The Shannon limit can be thought of as being based upon a theoretical constellation known as a Gaussian distribution, which is an infinite constellation where symbols at the center of the constellation are transmitted more frequently than symbols at the edge of the constellation. Practical constellations are finite and transmit symbols with equal likelihoods, and therefore have capacities that are less than the Gaussian capacity. The capacity of a constellation is thought to represent a limit on the gains that can be achieved using coding when using that constellation..

[0005] Prior attempts have been made to develop unequally spaced constellations. For example, a system has been proposed that uses unequally spaced constellations that are optimized to minimize the error rate of an uncoded system. Another proposed system uses a constellation with equiprobable but unequally spaced symbols in an attempts to mimic a Gaussian distribution.

[0006] Other approaches increases the dimensionality of a constellation or select a new symbol to be transmitted taking into consideration previously transmitted symbols. However, these constellation were still designed based on a minimum distance criteria..

SUMMARY OF THE INVENTION

[0007] Systems and methods are described for constructing a modulation such that the constrained capacity between a transmitter and a receiver approaches the Gaussian channel capacity limit first described by Shannon [ref Shannon 1948]. Traditional communications systems employ modulations that leave a significant gap to Shannon Gaussian capacity. The modulations of the present invention reduce, and in some cases, nearly eliminate this gap. The invention does not require specially designed coding mechanisms that tend to transmit some points of a modulation more frequently than others but rather provides a method for locating points (in a one or multiple dimensional space) in order to maximize capacity between the input and output of a bit or symbol mapper and demapper respectively. Practical application of the method allows systems to transmit data at a given rate for less power or to transmit data at a higher rate for the same amount of power.

[0008] One embodiment of the invention includes a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter, includes a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate, a mapper configured to map encoded bits to symbols in a symbol constellation, a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper. In addition, the receiver includes a demodulator configured to demodulate the received signal via the communication channel, a demapper configured to estimate likelihoods from the demodulated signal, a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper. Furthermore, the symbol

constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

[0009] A further embodiment of the invention includes encoding the bits of user information using a coding scheme, mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}, modulating the symbols in accordance with a modulation scheme, transmitting the modulated signal via the communication channel, receiving a modulated signal, demodulating the modulated signal in accordance with the modulation scheme, demapping the demodulated signal using the geometrically shaped signal constellation to produce likelihoods, and decoding the likelihoods to obtain an estimate of the decoded bits.

[0010] Another embodiment of the invention includes selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0011] A still further embodiment of the invention includes selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0012] Still another embodiment of the invention includes selecting an appropriate constellation size and a desired SNR, and optimizing the location of the points of the constellation to maximize a capacity measure of the constellation.

[0013] A yet further embodiment of the invention includes obtaining a geometrically shaped PAM constellation with a constellation size that is the square root of said given constellation size, where the geometrically shaped PAM constellation has a capacity greater than that of a PAM constellation that maximizes d_{min}, creating an orthogonalized PAM constellation using the

geometrically shaped PAM constellation, and combining the geometrically shaped PAM constellation and the orthogonalized PAM constellation to produce a geometrically shaped QAM constellation.

[0014] Another further embodiment of the invention includes transmitting information over a channel using a geometrically shaped symbol constellation, and modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate.

BRIEF DESCRIPTION OF DRAWINGS

[0015] FIG. 1 is a conceptual illustration of a communication system in accordance with an embodiment of the invention.

[0016] FIG. 2 is a conceptual illustration of a transmitter in accordance with an embodiment of the invention.

[0017] FIG. 3 is a conceptual illustration of a receiver in accordance with an embodiment of the invention.

[0018] FIG. 4a is a conceptual illustration of the joint capacity of a channel.

[0019] FIG. 4b is a conceptual illustration of the parallel decoding capacity of a channel.

[0020] FIG. 5 is a flow chart showing a process for obtaining a constellation optimized for capacity for use in a communication system having a fixed code rate and modulation scheme in accordance with an embodiment of the invention.

[0021] FIG. 6a is a chart showing a comparison of Gaussian capacity and PD capacity for traditional PAM-2,4,8,16,32.

[0022] FIG. 6b is a chart showing a comparison between Gaussian capacity and joint capacity for traditional PAM-2,4,8,16,32.

[0023] FIG. 7 is a chart showing the SNR gap to Gaussian capacity for the PD capacity and joint capacity of traditional PAM-2,4,8,16,32 constellations.

[0024] FIG. 8a is a chart comparing the SNR gap to Gaussian capacity of the PD capacity for traditional and optimized PAM-2,4,8,16,32 constellations.

[0025] FIG. 8b is a chart comparing the SNR gap to Gaussian capacity of the joint capacity for traditional and optimized PAM-2,4,8,16,32 constellations.

[0026] FIG. 9 is a chart showing Frame Error Rate performance of traditional and PD capacity optimized PAM-32 constellations in simulations involving several different length LDPC codes.

[0027] FIGS. 10a – 10d are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

[0028] FIGS. 11a and 11b are design tables of PD capacity and joint capacity optimized PAM-4 constellations in accordance with embodiments of the invention.

[0029] FIGS. 12a – 12d are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

[0030] FIGS. 13a and 13b are design tables of PD capacity and joint capacity optimized PAM-8 constellations in accordance with embodiments of the invention.

[0031] FIGS. 14a – 14d are locus plots showing the location of constellation points of a PAM-16 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

[0032] FIGS. 15a and 15b are design tables of PD capacity and joint capacity optimized PAM-16 constellations in accordance with embodiments of the invention.

[0033] FIGS. 16a – 16d are locus plots showing the location of constellation points of a PAM-32 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

[0034] FIGS. 17a and 17b are design tables of PD capacity and joint capacity optimized PAM-32 constellations in accordance with embodiments of the invention.

[0035] FIG. 18 is a chart showing the SNR gap to Gaussian capacity for traditional and capacity optimized PSK constellations.

[0036] FIG. 19 is a chart showing the location of constellation points of PD capacity optimized PSK-32 constellations.

[0037] FIG. 20 is a series of PSK-32 constellations optimized for PD capacity at different SNRs in accordance with embodiments of the invention.

[0038] FIG. 21 illustrates a QAM-64 constructed from orthogonal Cartesian product of two PD optimized PAM-8 constellations in accordance with an embodiment of the invention.

[0039] FIGS. 22a and 22b are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

[0040] FIGS. 23a and 23b are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

[0041] FIGS. 24a and 24b are locus plots showing the location of constellation points of a PAM-16 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

DETAILED DESCRIPTION OF THE INVENTION

[0042] Turning now to the drawings, communication systems in accordance with embodiments of the invention are described that use signal constellations, which have unequally spaced (i.e. 'geometrically' shaped) points. In several embodiments, the locations of geometrically shaped points are designed to provide a given capacity measure at a reduced signal-to-noise ratio (SNR) compared to the SNR required by a constellation that maximizes d_{min}. In many embodiments, the constellations are selected to provide increased capacity at a predetermined range of channel signal-to-noise ratios (SNR). Capacity measures that can be used in the selection of the location of constellation points include, but are not limited to, parallel decode (PD) capacity and joint capacity.

[0043] In many embodiments, the communication systems utilize capacity approaching codes including, but not limited to, LDPC and Turbo codes. As is discussed further below, direct optimization of the constellation points of a communication system utilizing a capacity approaching channel code, can yield different constellations depending on the SNR for which they are optimized. Therefore, the same constellation is unlikely to achieve the same coding gains applied across all code rates; that is, the same constellation will not enable the best possible performance across all rates. In many instances, a constellation at one code rate can achieve

gains that cannot be achieved at another code rate. Processes for selecting capacity optimized constellations to achieve increased coding gains based upon a specific coding rate in accordance with embodiments of the invention are described below. In a number of embodiments, the communication systems can adapt location of points in a constellation in response to channel conditions, changes in code rate and/or to change the target user data rate.

COMMUNICATION SYSTEMS

[0044] A communication system in accordance with an embodiment of the invention is shown in FIG. 1. The communication system 10 includes a source 12 that provides user bits to a transmitter 14. The transmitter transmits symbols over a channel to a receiver 16 using a predetermined modulation scheme. The receiver uses knowledge of the modulation scheme, to decode the signal received from the transmitter. The decoded bits are provided to a sink device that is connected to the receiver.

[0045] A transmitter in accordance with an embodiment of the invention is shown in FIG. 2. The transmitter 14 includes a coder 20 that receives user bits from a source and encodes the bits in accordance with a predetermined coding scheme. In a number of embodiments, a capacity approaching code such as a turbo code or a LDPC code is used. In other embodiments, other coding schemes can be used to providing a coding gain within the communication system. A mapper 22 is connected to the coder. The mapper maps the bits output by the coder to a symbol within a geometrically distributed signal constellation stored within the mapper. The mapper provides the symbols to a modulator 24, which modulates the symbols for transmission via the channel.

[0046] A receiver in accordance with an embodiment of the invention is illustrated in FIG. 3. The receiver 16 includes a demodulator 30 that demodulates a signal received via the channel to obtain symbol or bit likelihoods. The demapper uses knowledge of the geometrically shaped symbol constellation used by the transmitter to determine these likelihoods. The demapper 32 provides the likelihoods to a decoder 34 that decodes the encoded bit stream to provide a sequence of received bits to a sink.

GEOMETRICALLY SHAPED CONSTELLATIONS

[0047] Transmitters and receivers in accordance with embodiments of the invention utilize geometrically shaped symbol constellations. In several embodiments, a geometrically shaped symbol constellation is used that optimizes the capacity of the constellation. Various geometrically shaped symbol constellations that can be used in accordance with embodiments of the invention, techniques for deriving geometrically shaped symbol constellations are described below.

SELECTION OF A GEOMETRICALLY SHAPED CONSTELLATIONS

[0048] Selection of a geometrically shaped constellation for use in a communication system in accordance with an embodiment of the invention can depend upon a variety of factors including whether the code rate is fixed. In many embodiments, a geometrically shaped constellation is used to replace a conventional constellation (i.e. a constellation maximized for d_{min}) in the mapper of transmitters and the demapper of receivers within a communication system. Upgrading a communication system involves selection of a constellation and in many instances the upgrade can be achieved via a simple firmware upgrade. In other embodiments, a geometrically shaped constellation is selected in conjunction with a code rate to meet specific performance requirements, which can for example include such factors as a specified bit rate, a maximum transmit power. Processes for selecting a geometric constellation when upgrading existing communication systems and when designing new communication systems are discussed further below.

UPGRADING EXISTING COMMUNICATION SYSTEMS

[0049] A geometrically shaped constellation that provides a capacity, which is greater than the capacity of a constellation maximized for d_{min}, can be used in place of a conventional constellation in a communication system in accordance with embodiments of the invention. In many instances, the substitution of the geometrically shaped constellation can be achieved by a firmware or software upgrade of the transmitters and receivers within the communication system. Not all geometrically shaped constellations have greater capacity than that of a constellation maximized for d_{min}. One approach to selecting a geometrically shaped constellation

having a greater capacity than that of a constellation maximized for d_{min} is to optimize the shape of the constellation with respect to a measure of the capacity of the constellation for a given SNR. Capacity measures that can be used in the optimization process can include, but are not limited to, joint capacity or parallel decoding capacity.

JOINT CAPACITY AND PARALLEL DECODING CAPACITY

[0050] A constellation can be parameterized by the total number of constellation points, M, and the number of real dimensions, N_{dim}. In systems where there are no belief propagation iterations between the decoder and the constellation demapper, the constellation demapper can be thought of as part of the channel. A diagram conceptually illustrating the portions of a communication system that can be considered part of the channel for the purpose of determining PD capacity is shown in FIG. 4a. The portions of the communication system that are considered part of the channel are indicated by the ghost line 40. The capacity of the channel defined as such is the parallel decoding (PD) capacity, given by:

$$C_{PD} = \sum_{i=0}^{i-1} I(X_i; Y)$$

where X_i is the ith bit of the i-bits transmitted symbol, and Y is the received symbol, and $I(A;B)$ denotes the mutual information between random variables A and B.

[0051] Expressed another way, the PD capacity of a channel can be viewed in terms of the mutual information between the output bits of the encoder (such as an LDPC encoder) at the transmitter and the likelihoods computed by the demapper at the receiver. The PD capacity is influenced by both the placement of points within the constellation and by the labeling assignments.

[0052] With belief propagation iterations between the demapper and the decoder, the demapper can no longer be viewed as part of the channel, and the joint capacity of the constellation becomes the tightest known bound on the system performance. A diagram conceptually illustrating the portions of a communication system that are considered part of the channel for the purpose of determining the joint capacity of a constellation is shown in FIG. 4b.

The portions of the communication system that are considered part of the channel are indicated by the ghost line 42. The joint capacity of the channel is given by:

$$C_{JOINT} = I(X;Y)$$

[0053] Joint capacity is a description of the achievable capacity between the input of the mapper on the transmit side of the link and the output of the channel (including for example AWGN and Fading channels). Practical systems must often 'demap' channel observations prior to decoding. In general, the step causes some loss of capacity. In fact it can be proven that $C_Q \geq C_{JOINT} \geq C_{PD}$. That is, C_{JOINT} upper bounds the capacity achievable by C_{PD}. The methods of the present invention are motivated by considering the fact that practical limits to a given communication system capacity are limited by C_{JOINT} and C_{PD}. In several embodiments of the invention, geometrically shaped constellations are selected that maximize these measures.

SELECTING A CONSTELLATION HAVING AN OPTIMAL CAPACITY

[0054] Geometrically shaped constellations in accordance with embodiments of the invention can be designed to optimize capacity measures including, but not limited to PD capacity or joint capacity. A process for selecting the points, and potentially the labeling, of a geometrically shaped constellation for use in a communication system having a fixed code rate in accordance with an embodiment of the invention is shown in FIG. 5. The process 50 commences with the selection (52) of an appropriate constellation size M and a desired capacity per dimension η. In the illustrated embodiment, the process involves a check (52) to ensure that the constellation size can support the desired capacity. In the event that the constellation size could support the desired capacity, then the process iteratively optimizes the M-ary constellation for the specified capacity. Optimizing a constellation for a specified capacity often involves an iterative process, because the optimal constellation depends upon the SNR at which the communication system operates. The SNR for the optimal constellation to give a required capacity is not known *a priori*. Throughout the description of the present invention SNR is defined as the ratio of the average constellation energy per dimension to the average noise energy per dimension. In most cases the capacity can be set to equal the target user bit rate per symbol per dimension. In some cases adding some implementation margin on top of the target

user bit rate could result in a practical system that can provide the required user rate at a lower rate. The margin is code dependent. The following procedure could be used to determine the target capacity that includes some margin on top of the user rate. First, the code (e.g. LDPC or Turbo) can be simulated in conjunction with a conventional equally spaced constellation. Second, from the simulation results the actual SNR of operation at the required error rate can be found. Third, the capacity of the conventional constellation at that SNR can be computed. Finally, a geometrically shaped constellation can be optimized for that capacity.

[0055] In the illustrated embodiment, the iterative optimization loop involves selecting an initial estimate of the SNR at which the system is likely to operate (i.e. SNR_{in}). In several embodiments the initial estimate is the SNR required using a conventional constellation. In other embodiments, other techniques can be used for selecting the initial SNR. An *M*-ary constellation is then obtained by optimizing (56) the constellation to maximize a selected capacity measure at the initial SNR_{in} estimate. Various techniques for obtaining an optimized constellation for a given SNR estimate are discussed below.

[0056] The SNR at which the optimized *M*-ary constellation provides the desired capacity per dimension η (SNR_{out}) is determined (57). A determination (58) is made as to whether the SNR_{out} and SNR_{in} have converged. In the illustrated embodiment convergence is indicated by SNR_{out} equaling SNR_{in}. In a number of embodiments, convergence can be determined based upon the difference between SNR_{out} and SNR_{in} being less than a predetermined threshold. When SNR_{out} and SNR_{in} have not converged, the process performs another iteration selecting SNR_{out} as the new SNR_{in} (55). When SNR_{out} and SNR_{in} have converged, the capacity measure of the constellation has been optimized. As is explained in more detail below, capacity optimized constellation at low SNRs are geometrically shaped constellations that can achieve significantly higher performance gains (measured as reduction in minimum required SNR) than constellations that maximize d_{min}.

[0057] The process illustrated in FIG. 5 can maximize PD capacity or joint capacity of an *M*-ary constellation for a given SNR. Although the process illustrated in FIG. 5 shows selecting an *M*-ary constellation optimized for capacity, a similar process could be used that terminates upon generation of an *M*-ary constellation where the SNR gap to Gaussian capacity at a given capacity is a predetermined margin lower than the SNR gap of a conventional constellation, for example

0.5 db. Alternatively, other processes that identify M-ary constellations having capacity greater than the capacity of a conventional constellation can be used in accordance with embodiments of the invention. A geometrically shaped constellation in accordance with embodiments of the invention can achieve greater capacity than the capacity of a constellation that maximizes d_{min} without having the optimal capacity for the SNR range within which the communication system operates.

[0058] We note that constellations designed to maximize joint capacity may also be particularly well suited to codes with symbols over GF(q) , or with multi-stage decoding.. Conversely constellations optimized for PD capacity could be better suited to the more common case of codes with symbols over GF(2)

OPTIMIZING THE CAPACITY OF AN M-ARY CONSTELLATION AT A GIVEN SNR

[0059] Processes for obtaining a capacity optimized constellation often involve determining the optimum location for the points of an M-ary constellation at a given SNR. An optimization process, such as the optimization process 56 shown in FIG. 5, typically involves unconstrained or constrained non-linear optimization. Possible objective functions to be maximized are the Joint or PD capacity functions. These functions may be targeted to channels including but not limited to Additive White Gaussian Noise (AWGN) or Rayleigh fading channels. The optimization process gives the location of each constellation point identified by its symbol labeling. In the case where the objective is joint capacity, point bit labelings are irrelevant meaning that changing the bit labelings doesn't change the joint capacity as long as the set of point locations remains unchanged..

[0060] The optimization process typically finds the constellation that gives the largest PD capacity or joint capacity at a given SNR. The optimization process itself often involves an iterative numerical process that among other things considers several constellations and selects the constellation that gives the highest capacity at a given SNR. In other embodiments, the constellation that requires the least SNR to give a required PD capacity or joint capacity can also be found. This requires running the optimization process iteratively as shown in FIG. 5.

[0061] Optimization constraints on the constellation point locations may include, but are not limited to, lower and upper bounds on point location, peak to average power of the resulting

constellation, and zero mean in the resulting constellation. It can be easily shown that a globally optimal constellation will have zero mean (no DC component). Explicit inclusion of a zero mean constraint helps the optimization routine to converge more rapidly. Except for cases where exhaustive search of all combinations of point locations and labelings is possible it will not necessarily always be the case that solutions are provably globally optimal. In cases where exhaustive search is possible, the solution provided by the non-linear optimizer is in fact globally optimal.

[0062] The processes described above provide examples of the manner in which a geometrically shaped constellation having an increased capacity relative to a conventional capacity can be obtained for use in a communication system having a fixed code rate and modulation scheme. The actual gains achievable using constellations that are optimized for capacity compared to conventional constellations that maximize d_{min} are considered below.

GAINS ACHIEVED BY OPTIMIZED GEOMETRICALLY SPACED CONSTELLATIONS

[0063] The ultimate theoretical capacity achievable by any communication method is thought to be the Gaussian capacity, C_G which is defined as:

$$C_G = \frac{1}{2}log_2(1 + SNR)$$

[0064] Where signal-to-noise (SNR) is the ratio of expected signal power to expected noise power. The gap that remains between the capacity of a constellation and C_G can be considered a measure of the quality of a given constellation design.

[0065] The gap in capacity between a conventional modulation scheme in combination with a theoretically optimal coder can be observed with reference to FIGS. 6a and 6b. FIG. 6a includes a chart 60 showing a comparison between Gaussian capacity and the PD capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}. Gaps 62 exist between the plot of Gaussian capacity and the PD capacity of the various PAM constellations. FIG. 6b includes a chart 64 showing a comparison between Gaussian capacity and the joint capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}. Gaps 66 exist between the plot of Gaussian capacity and the joint capacity of the various PAM constellations. These

gaps in capacity represent the extent to which conventional PAM constellations fall short of obtaining the ultimate theoretical capacity i.e. the Gaussian capacity,.

[0066] In order to gain a better view of the differences between the curves shown in FIGS. 6a and 6b at points close to the Gaussian capacity, the SNR gap to Gaussian capacity for different values of capacity for each constellation are plotted in FIG. 7. It is interesting to note from the chart 70 in FIG. 7 that (unlike the joint capacity) at the same SNR, the PD capacity does not necessarily increase with the number of constellation points. As is discussed further below, this is not the case with PAM constellations optimized for PD capacity.

[0067] FIGS. 8a and 8b summarize performance of constellations for PAM-4, 8, 16, and 32 optimized for PD capacity and joint capacity (it should be noted that BPSK is the optimal PAM-2 constellation at all code rates). The constellations are optimized for PD capacity and joint capacity for different target user bits per dimension (i.e. code rates). The optimized constellations are different depending on the target user bits per dimension, and also depending on whether they have been designed to maximize the PD capacity or the joint capacity. All the PD optimized PAM constellations are labeled using a gray labeling which is not always the binary reflective gray labeling. It should be noted that not all gray labels achieve the maximum possible PD capacity even given the freedom to place the constellation points anywhere on the real line. FIG. 8a shows the SNR gap for each constellation optimized for PD capacity. FIG. 8b shows the SNR gap to Gaussian capacity for each constellation optimized for joint capacity. Again, it should be emphasized that each '+' on the plot represents a different constellation.

[0068] Referring to FIG. 8a, the coding gain achieved using a constellation optimized for PD capacity can be appreciated by comparing the SNR gap at a user bit rate per dimension of 2.5 bits for PAM-32. A user bit rate per dimension of 2.5 bits for a system transmitting 5 bits per symbol constitutes a code rate of 1/2. At that code rate the constellation optimized for PD capacity provides an additional coding gain of approximately 1.5 dB when compared to the conventional PAM-32 constellation.

[0069] The SNR gains that can be achieved using constellations that are optimized for PD capacity can be verified through simulation. The results of a simulation conducted using a rate 1/2 LDPC code in conjunction with a conventional PAM-32 constellation and in conjunction with a PAM-32 constellation optimized for PD capacity are illustrated in FIG. 9. A chart 90

includes plots of Frame Error Rate performance of the different constellations with respect to SNR and using different length codes (i.e. k=4,096 and k=16,384). Irrespective of the code that is used, the constellation optimized for PD capacity achieves a gain of approximately 1.3 dB, which closely approaches the gain predicted from FIG. 8a.

CAPACITY OPTIMIZED PAM CONSTELLATIONS

[0070] Using the processes outlined above, locus plots of PAM constellations optimized for capacity can be generated that show the location of points within PAM constellations versus SNR. Locus plots of PAM-4, 8, 16, and 32 constellations optimized for PD capacity and joint capacity and corresponding design tables at various typical user bit rates per dimension are illustrated in FIGS. 10a – 17b. The locus plots and design tables show PAM-4,8,16,32 constellation point locations and labelings from low to high SNR corresponding to a range of low to high spectral efficiency.

[0071] In FIG.10a, a locus plot 100 shows the location of the points of PAM-4 constellations optimized for Joint capacity plotted against achieved capacity. A similar locus plot 105 showing the location of the points of Joint capacity optimized PAM-4 constellations plotted against SNR is included in FIG. 10b. In FIG. 10c, the location of points for PAM-4 optimized for PD capacity is plotted against achievable capacity and in FIG. 10d the location of points for PAM-4 for PD capacity is plotted against SNR. At low SNRs, the PD capacity optimized PAM-4 constellations have only 2 unique points, while the Joint optimized constellations have 3. As SNR is increased, each optimization eventually provides 4 unique points. This phenomenon is explicitly described in FIG. 11a and FIG. 11b where vertical slices of FIGS. 10ab and 10cd are captured in tables describing some PAM-4 constellations designs of interest. The SNR slices selected represent designs that achieve capacities = {0.5, 0.75, 1.0, 1.25, 1.5} bits per symbol (bps). Given that PAM-4 can provide at most $\log_2(4) = 2$ bps, these design points represent systems with information code rates R = {1/4, 3/8, 1/2, 5/8, 3/4} respectively.

[0072] FIGS. 12ab and 12cd present locus plots of PD capacity and joint capacity optimized PAM-8 constellation points versus achievable capacity and SNR. FIGS. 13a and 13b provide slices from these plots at SNRs corresponding to achievable capacities η = {0.5, 1.0, 1.5, 2.0, 2.5} bps. Each of these slices correspond to systems with code rate $R = \eta$ bps / $\log_2(8)$, resulting

in R ={1/6, 1/3, 1/2, 2/3, 5/6}. As an example of the relative performance of the constellations in these tables, consider FIG. 13b which shows a PD capacity optimized PAM-8 constellation optimized for SNR = 9.00 dB, or 1.5 bps. We next examine the plot provided in FIG. 8a and see that the gap of the optimized constellation to the ultimate, Gaussian, capacity (CG) is approximately 0.5 dB. At the same spectral efficiency, the gap of the traditional PAM-8 constellation is approximately 1.0 dB. The advantage of the optimized constellation is 0.5dB for the same rate (in this case R=1/2). This gain can be obtained by only changing the mapper and demapper in the communication system and leaving all other blocks the same.

[0073] Similar information is presented in FIGS. 14abcd, and 15ab which provide loci plots and *design* tables for PAM-16 PD capacity and joint capacity optimized constellations. Likewise FIGS. 16abcd, 17ab provide loci plots and design tables for PAM-32 PD capacity and joint capacity optimized constellations.

CAPACITY OPTIMIZED PSK CONSTELLATIONS

[0074] Traditional phase shift keyed (PSK) constellations are already quite optimal. This can be seen in the chart 180 comparing the SNR gaps of tradition PSK with capacity optimized PSK constellations shown in FIG. 18 where the gap between PD capacity and Gaussian capacity is plotted for traditional PSK-4,8,16,32 and for PD capacity optimized PSK-4,8,16,32.

[0075] The locus plot of PD optimized PSK-32 points across SNR is shown in FIG. 19, which actually characterizes all PSKs with spectral efficiency $\eta \leq 5$. This can be seen in FIG. 20. Note that at low SNR (0.4 dB) the optimal PSK-32 design is the same as traditional PSK-4, at SNR = 8.4 dB optimal PSK-32 is the same as traditional PSK-8, at SNR = 14.8 dB optimal PSK-32 is the same as traditional PSK-16, and finally at SNRs greater than 20.4 dB optimized PSK-32 is the same as traditional PSK-32. There are SNRs between these discrete points (for instance SNR = 2 and 15. dB) for which optimized PSK-32 provides superior PD capacity when compared to traditional PSK constellations.

[0076] We note now that the locus of points for PD optimized PSK-32 in FIG. 19 in conjunction with the gap to Gaussian capacity curve for optimized PSK-32 in FIG. 18 implies a potential design methodology. Specifically, the designer could achieve performance equivalent or better than that enabled by traditional PSK-4,8,16 by using only the optimized PSK-32 in

conjunction with a single tuning parameter that controlled where the constellation points should be selected from on the locus of FIG. 19. Such an approach would couple a highly rate adaptive channel code that could vary its rate, for instance, rate 4/5 to achieve and overall (code plus optimized PSK-32 modulation) spectral efficiency of 4 bits per symbol, down to 1/5 to achieve an overall spectral efficiency of 1 bit per symbol. Such an adaptive modulation and coding system could essentially perform on the optimal continuum represented by the rightmost contour of FIG. 18.

ADAPTIVE RATE DESIGN

[0077] In the previous example spectrally adaptive use of PSK-32 was described. Techniques similar to this can be applied for other capacity optimized constellations across the link between a transmitter and receiver. For instance, in the case where a system implements quality of service it is possible to instruct a transmitter to increase or decrease spectral efficiency on demand. In the context of the current invention a capacity optimized constellation designed precisely for the target spectral efficiency can be loaded into the transmit mapper in conjunction with a code rate selection that meets the end user rate goal. When such a modulation / code rate change occurred a message could propagated to the receiver so that the receiver, in anticipation of the change, could select a demapper / decoder configuration in order to match the new transmit-side configuration.

[0078] Conversely, the receiver could implement a quality of performance based optimized constellation / code rate pair control mechanism. Such an approach would include some form of receiver quality measure. This could be the receiver's estimate of SNR or bit error rate. Take for example the case where bit error rate was above some acceptable threshold. In this case, via a backchannel, the receiver could request that the transmitter lower the spectral efficiency of the link by swapping to an alternate capacity optimized constellation / code rate pair in the coder and mapper modules and then signaling the receiver to swap in the complementary pairing in the demapper / decoder modules.

GEOMETRICALLY SHAPED QAM CONSTELLATIONS

[0079] Quadrature amplitude modulation (QAM) constellations can be constructed by orthogonalizing PAM constellations into QAM inphase and quadrature components. Constellations constructed in this way can be attractive in many applications because they have low-complexity demappers.

[0080] In FIG. 21 we provide an example of a Quadrature Amplitude Modulation constellation constructed from a Pulse Amplitude Modulation constellation. The illustrated embodiment was constructed using a PAM-8 constellation optimized for PD capacity at user bit rate per dimension of 1.5 bits (corresponds to an SNR of 9.0 dB) (see FIG. 13b). The label-point pairs in this PAM-8 constellation are {(000, -1.72), (001, -0.81), (010, 1.72), (011,-0.62), (100, 0.62), (101, 0.02), (110, 0.81), (111, -0.02)}. Examination of FIG. 21 shows that the QAM constellation construction is achieved by replicating a complete set of PAM-8 points in the quadrature dimension for each of the 8 PAM-8 points in the in-phase dimension. Labeling is achieved by assigning the PAM-8 labels to the LSB range on the in-phase dimension and to the MSB range on the quadrature dimension. The resulting 8x8 outer product forms a highly structured QAM-64 for which very low-complexity de-mappers can be constructed. Due to the orthogonality of the in-phase and quadrature components the capacity characteristics of the resulting QAM-64 constellation are identical to that of the PAM-8 constellation on a per-dimension basis.

N-DIMENSIONAL CONSTELLATION OPTIMIZATION

[0081] Rather than designing constellations in 1-D (PAM for instance) and then extending to 2-D (QAM), it is possible to take direct advantage in the optimization step of the additional degree of freedom presented by an extra spatial dimension. In general it is possible to design N-dimensional constellations and associated labelings. The complexity of the optimization step grows exponentially in the number of dimensions as does the complexity of the resulting receiver de-mapper. Such constructions constitute embodiments of the invention and simply require more 'run-time' to produce.

CAPACITY OPTIMIZED CONSTELALTIONS FOR FADING CHANNELS

[0082] Similar processes to those outlined above can be used to design capacity optimized constellations for fading channels in accordance with embodiments of the invention. The processes are essentially the same with the exception that the manner in which capacity is calculated is modified to account for the fading channel. A fading channel can be described using the following equation:

$$Y = a(t).X+N$$

where X is the transmitted signal, N is an additive white Gaussian noise signal and $a(t)$ is the fading distribution, which is a function of time.

[0083] In the case of a fading channel, the instantaneous SNR at the receiver changes according to a fading distribution. The fading distribution is Rayleigh and has the property that the average SNR of the system remains the same as in the case of the AWGN channel, $E[Y^2]/E[N^2]$. Therefore, the capacity of the fading channel can be computed by taking the expectation of AWGN capacity, at a given average SNR, over the Rayleigh fading distribution of a that drives the distribution of the instantaneous SNR.

[0084] Many fading channels follow a Rayleigh distribution. FIGS. 22a – 24b are locus plots of PAM-4, 8, and 16 constellations that have been optimized for PD capacity on a Rayleigh fading channel. Locus plots versus user bit rate per dimension and versus SNR are provided. Similar processes can be used to obtain capacity optimized constellations that are optimized using other capacity measures, such as joint capacity, and/or using different modulation schemes.

WHAT IS CLAIMED IS:

1. A digital communication system, comprising:

a transmitter configured to transmit signals to a receiver via a communication channel;

wherein the transmitter, comprises:

a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate;

a mapper configured to map encoded bits to symbols in a symbol constellation;

a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper;

wherein the receiver, comprises:

a demodulator configured to demodulate the received signal via the communication channel;

a demapper configured to estimate likelihoods from the demodulated signal;

a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper; and

wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

2. The communication system of claim 1, wherein the geometrically spaced symbol constellation is capacity optimized subject to additional constraints.

3. The communication system of claim 1, wherein the geometrically shaped symbol constellation is optimized for capacity using parallel decode capacity.

4. The communication system of claim 1, wherein the geometrically shaped symbol constellation is optimized for capacity using joint capacity.

5. The communication system of claim 1, wherein the channel is an AWGN channel.

6. The communication system of claim 1, wherein the channel is a fading channel.

7. The communication system of claim 1, wherein the constellations points of the geometrically spaced symbol constellation are variable depending on the user rate.

8. A method of transmitting bits of user information over a communication channel, comprising:

encoding the bits of user information using a coding scheme;

mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min};

modulating the symbols in accordance with a modulation scheme;

transmitting the modulated signal via the communication channel;

receiving a modulated signal;

demodulating the modulated signal in accordance with the modulation scheme;

demapping the demodulated signal using the geometrically shaped signal constellation to produce likelihoods;

decoding the likelihoods to obtain an estimate of the decoded bits.

9. A method of selecting the points of a geometrically shaped symbol constellation used in a communication system, comprising:

selecting an appropriate constellation size and a desired capacity per dimension;

estimating an initial SNR at which the system is likely to operate; and

iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

10. A geometrically shaped symbol constellation used in a communication system, obtained by a process of:

selecting an appropriate constellation size and a desired capacity per dimension;

estimating an initial SNR at which the system is likely to operate; and

iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

11. A geometrically shaped symbol constellation used in a communication system, obtained by a process of:

selecting an appropriate constellation size and a desired SNR; and

optimizing the location of the points of the constellation to maximize a capacity measure of the constellation.

12. A method of transmitting information, comprising:

transmitting information over a channel using a geometrically shaped symbol constellation; and

modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate.

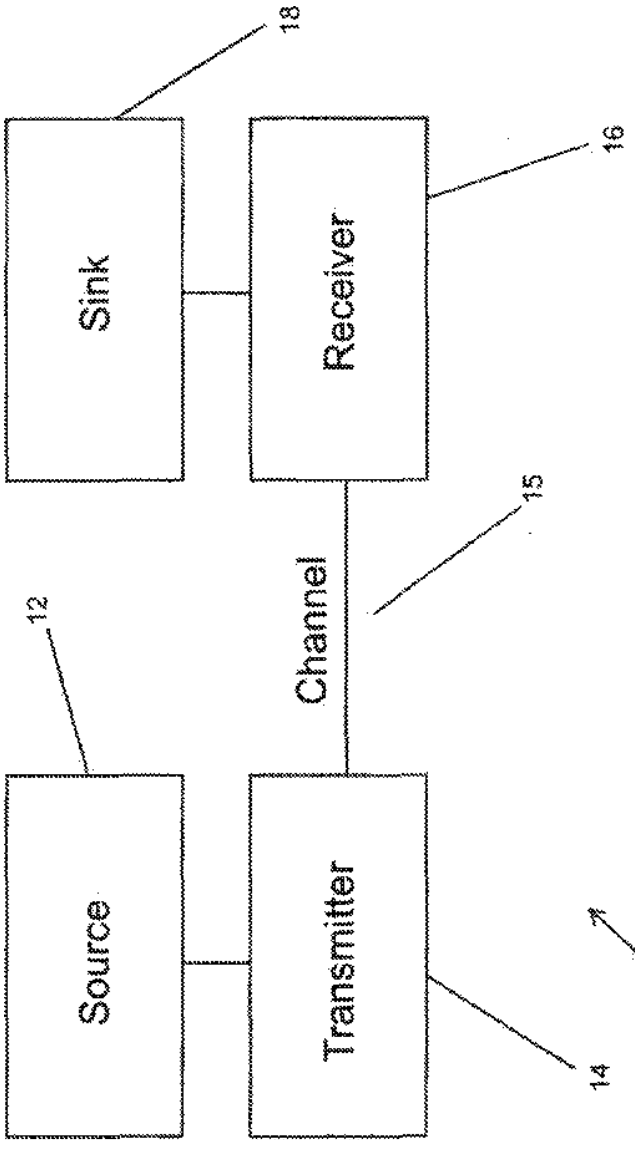


FIG. 1

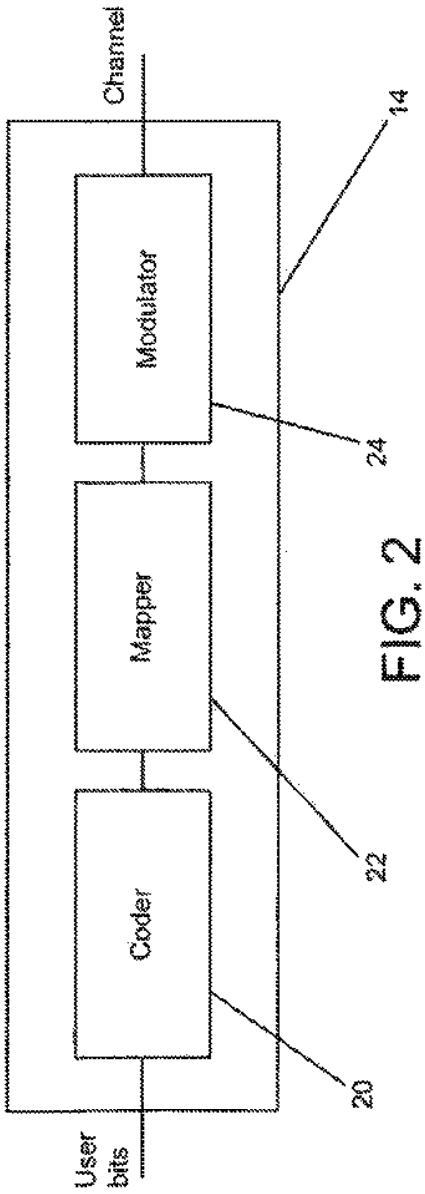

Channel
Modulator
Mapper
Coder
User bits
14
24
22
20

FIG. 2

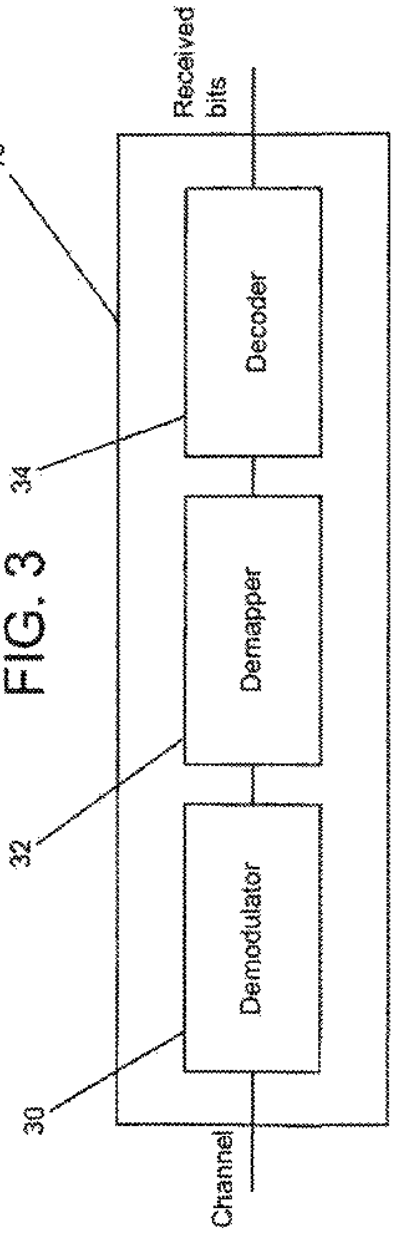

Received bits
Decoder
Demapper
Demodulator
Channel
16
34
32
30

FIG. 3

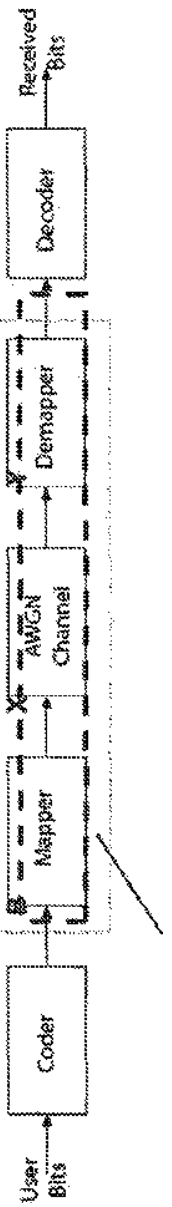

Parallel Decoding Capacity
User Bits
Coder
B
Mapper
X
AWGN Channel
Demapper
Decoder
Received Bits
40

FIG. 4a

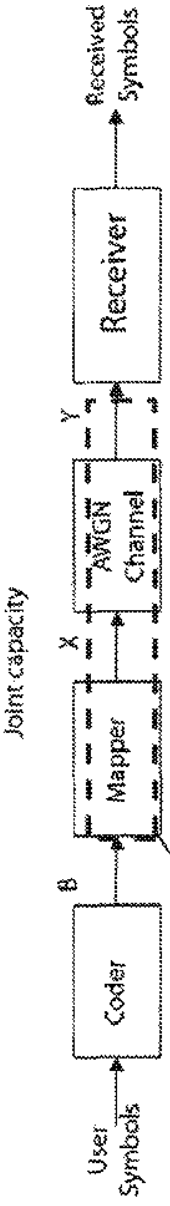

Joint capacity
User Symbols
Coder
B
Mapper
X
AWGN Channel
Y
Receiver
Received Symbols
42

FIG. 4b

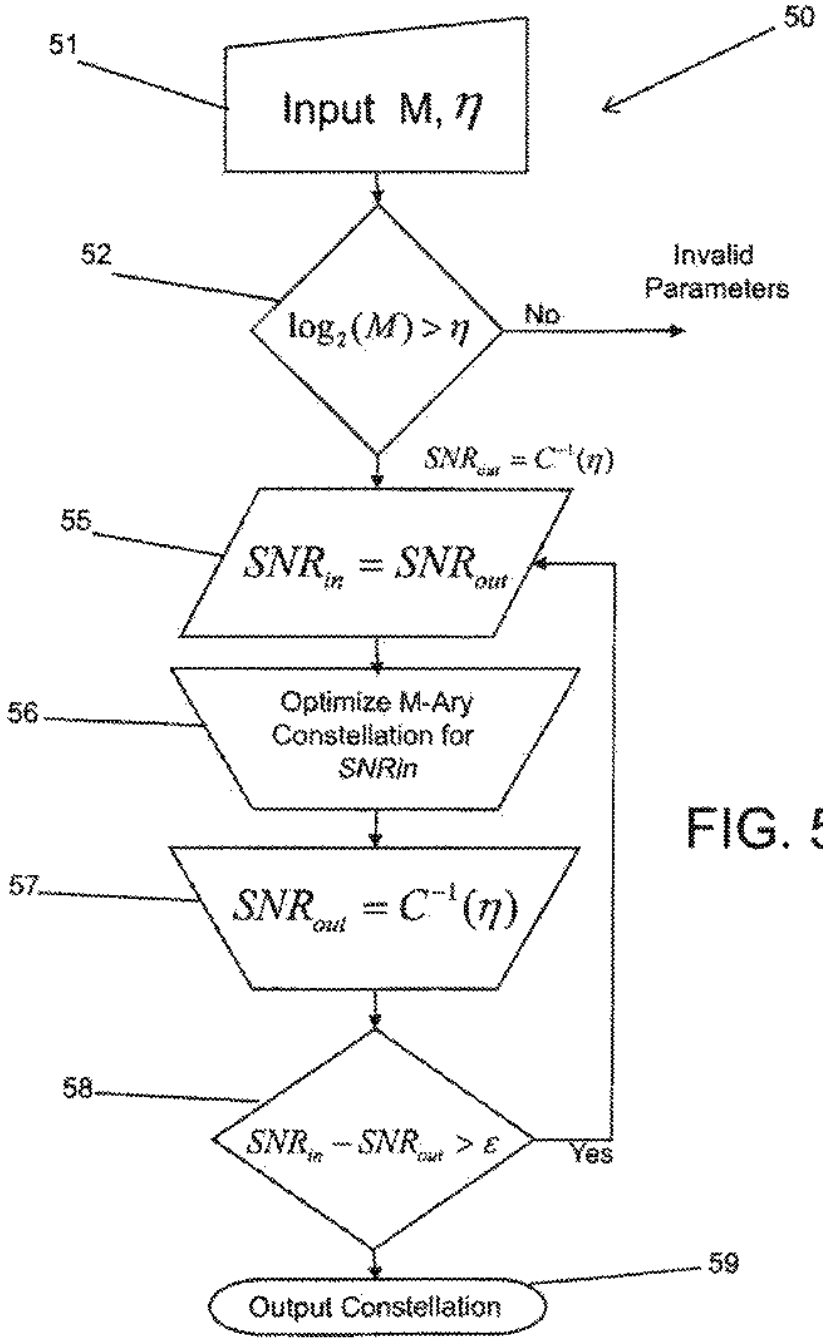

50
51
Input M, η
52
log2(M) > η
No
Invalid Parameters
SNRout = C-1(η)
55
SNRin = SNRout
56
Optimize M-Ary Constellation for SNRin
57
SNRout = C-1(η)
58
SNRin − SNRout > ε
Yes
59
Output Constellation

FIG. 5

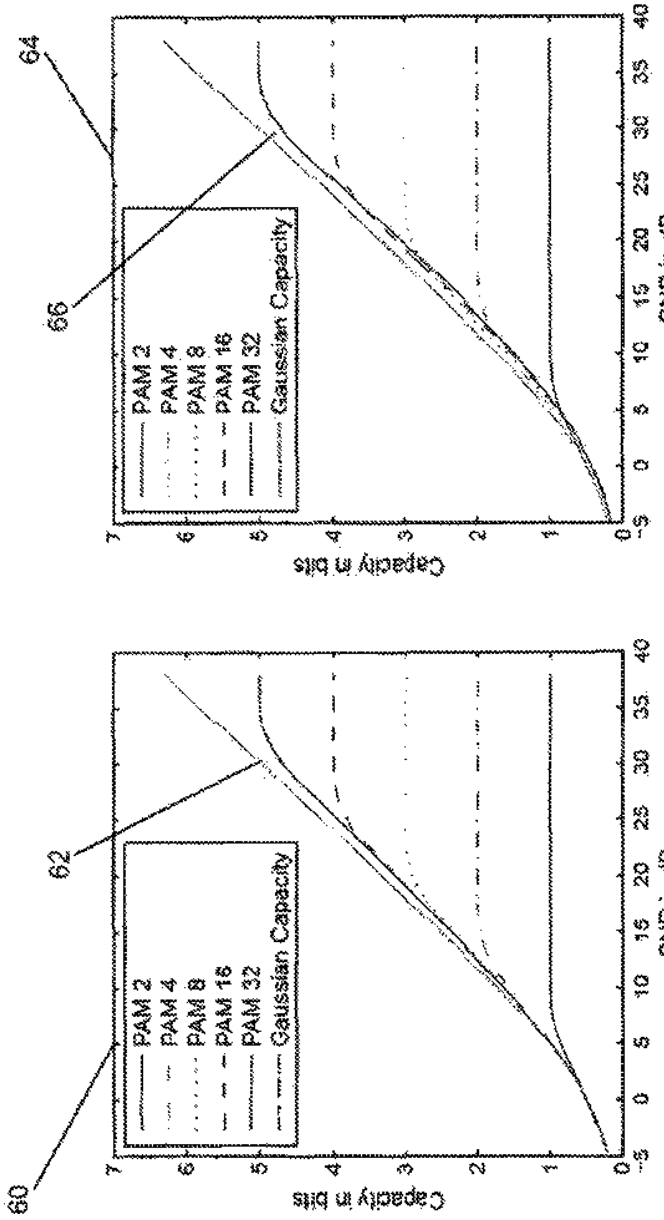

60
62
64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(a)
(b)

FIG. 6a

FIG. 6b

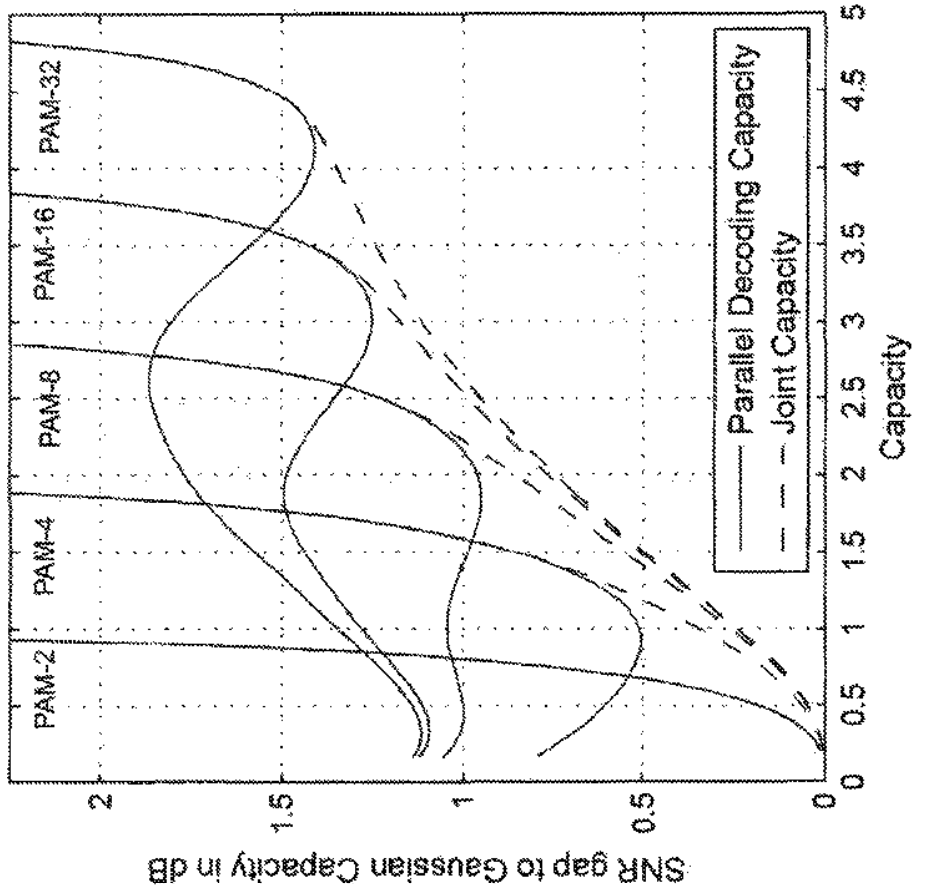

SNR gap to Gaussian Capacity in dB
Capacity
Parallel Decoding Capacity
Joint Capacity
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 7

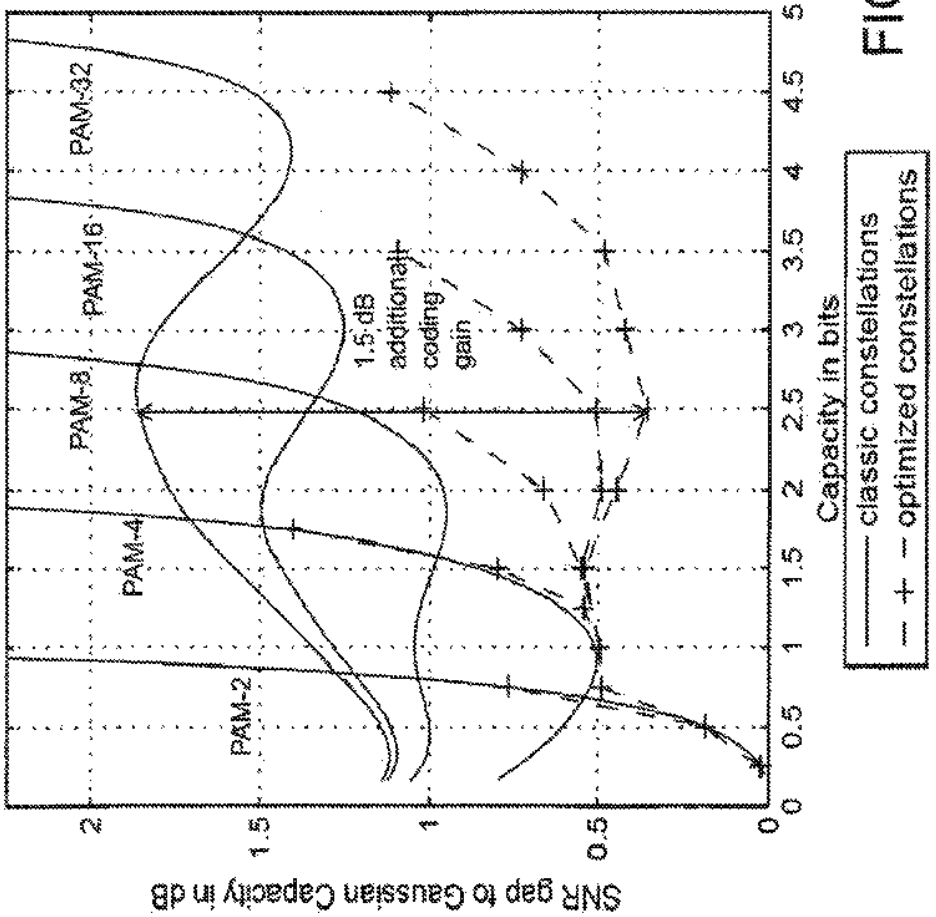

SNR gap to Gaussian Capacity in dB
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
1.5 dB additional coding gain
classic constellations
optimized constellations

FIG. 8a

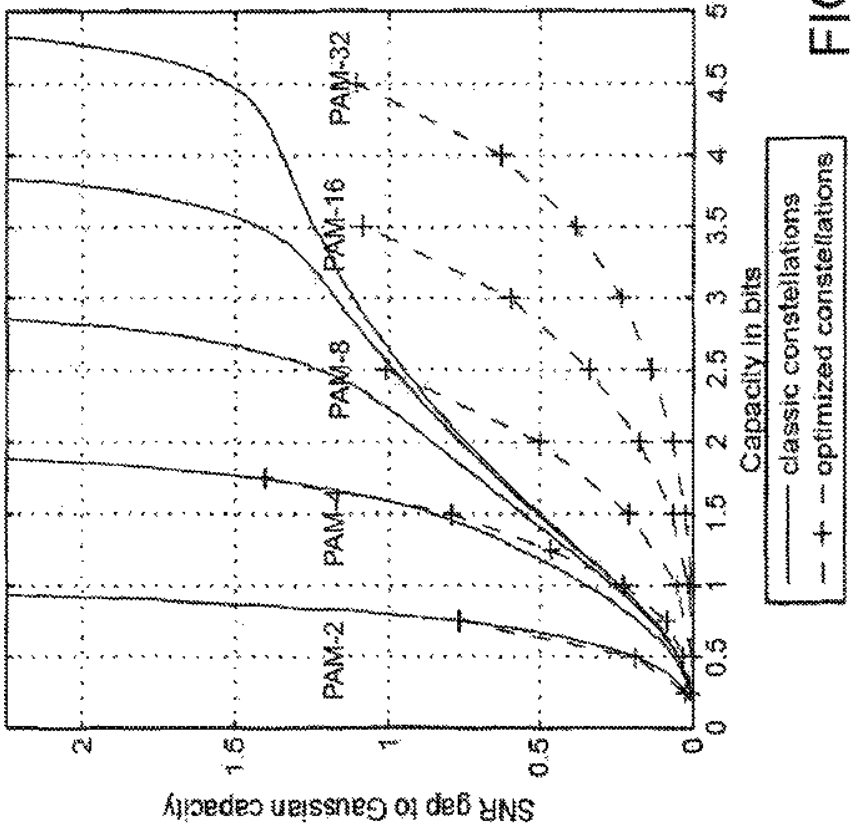

SNR gap to Gaussian capacity
Capacity in bits
classic constellations
optimized constellations
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8b

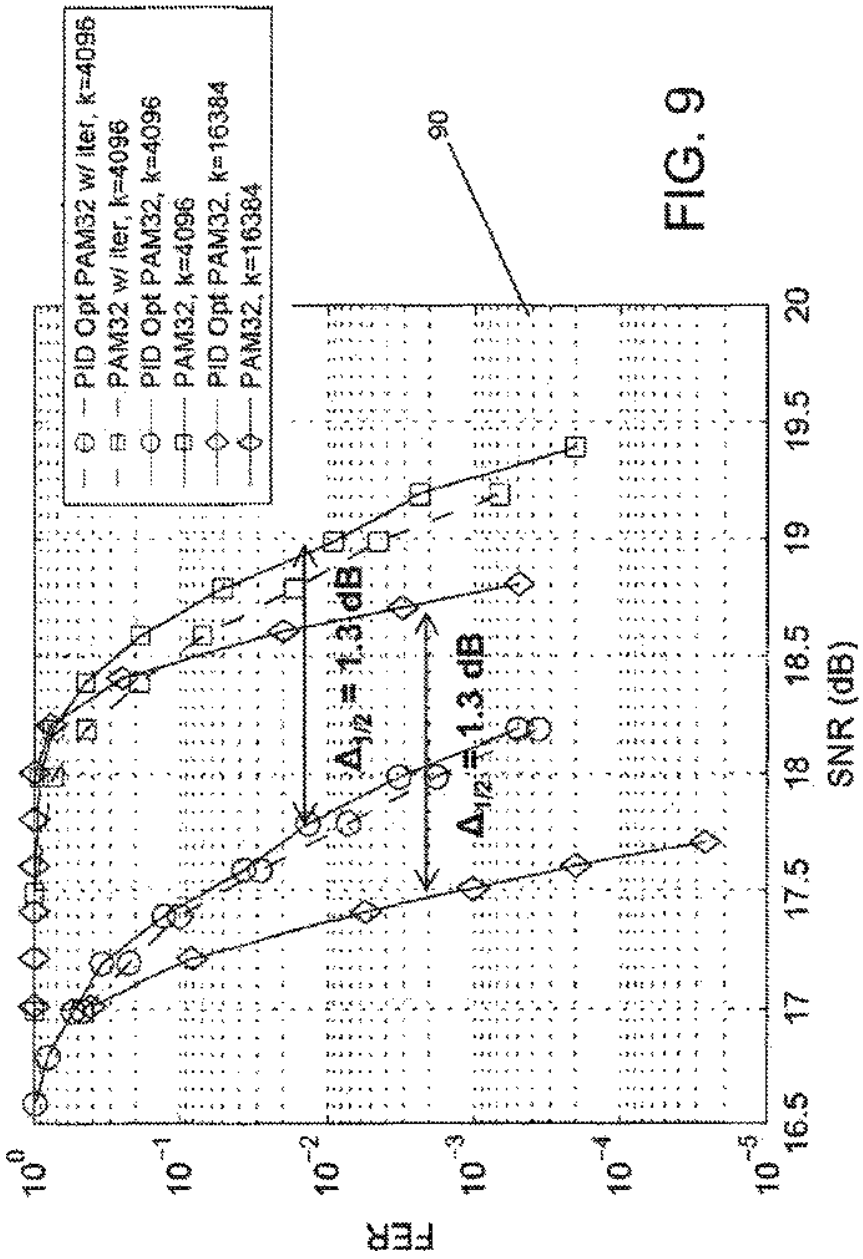

PID Opt PAM32 w/ iter, k=4096
PAM32 w/ iter, k=4096
PID Opt PAM32, k=4096
PAM32, k=4096
PID Opt PAM32, k=16384
PAM32, k=16384
90
$\Delta_{1/2}$ = 1.3 dB
$\Delta_{1/2}$ = 1.3 dB
FER
SNR (dB)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20

FIG. 9

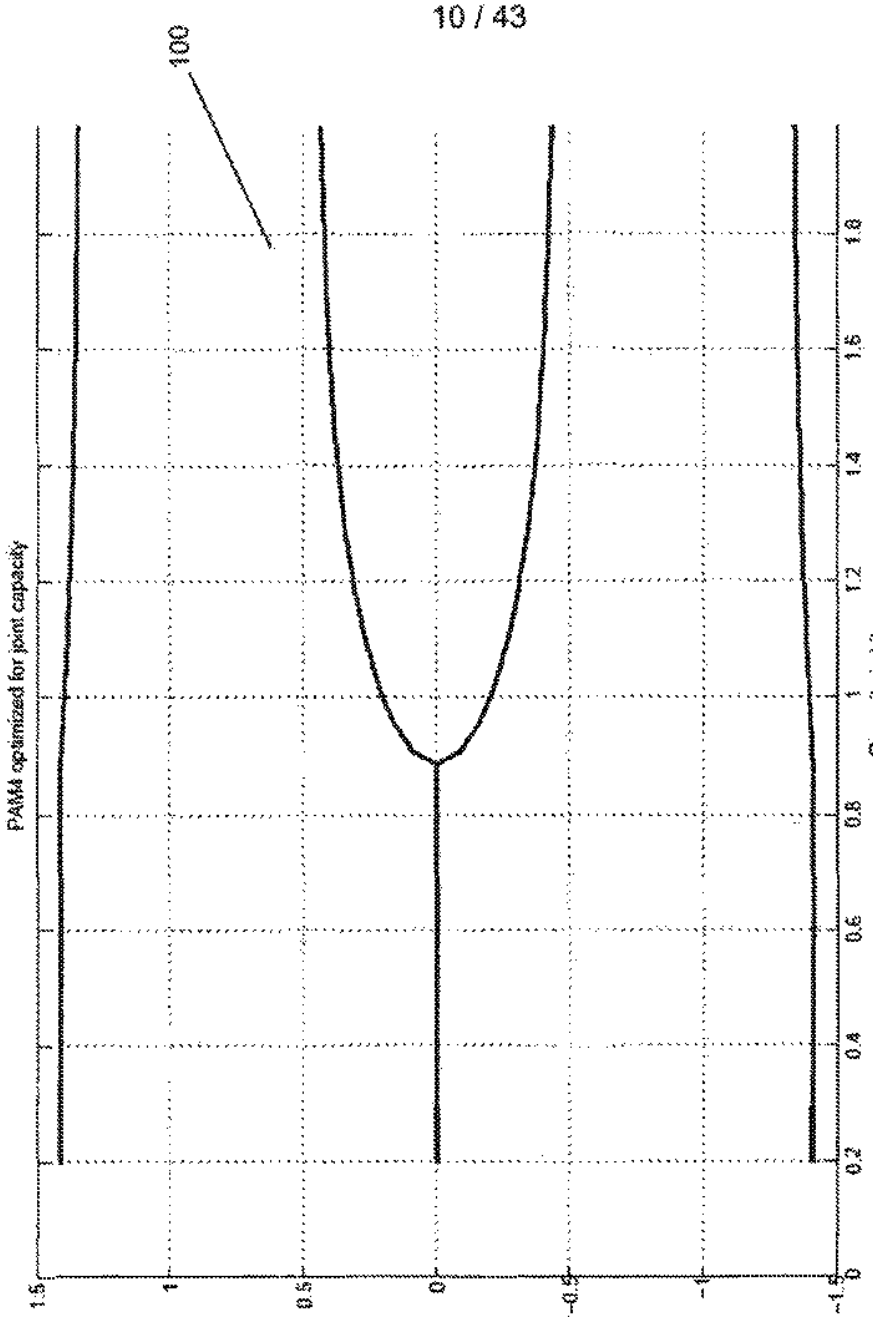

100
PAMA optimized for joint capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10a

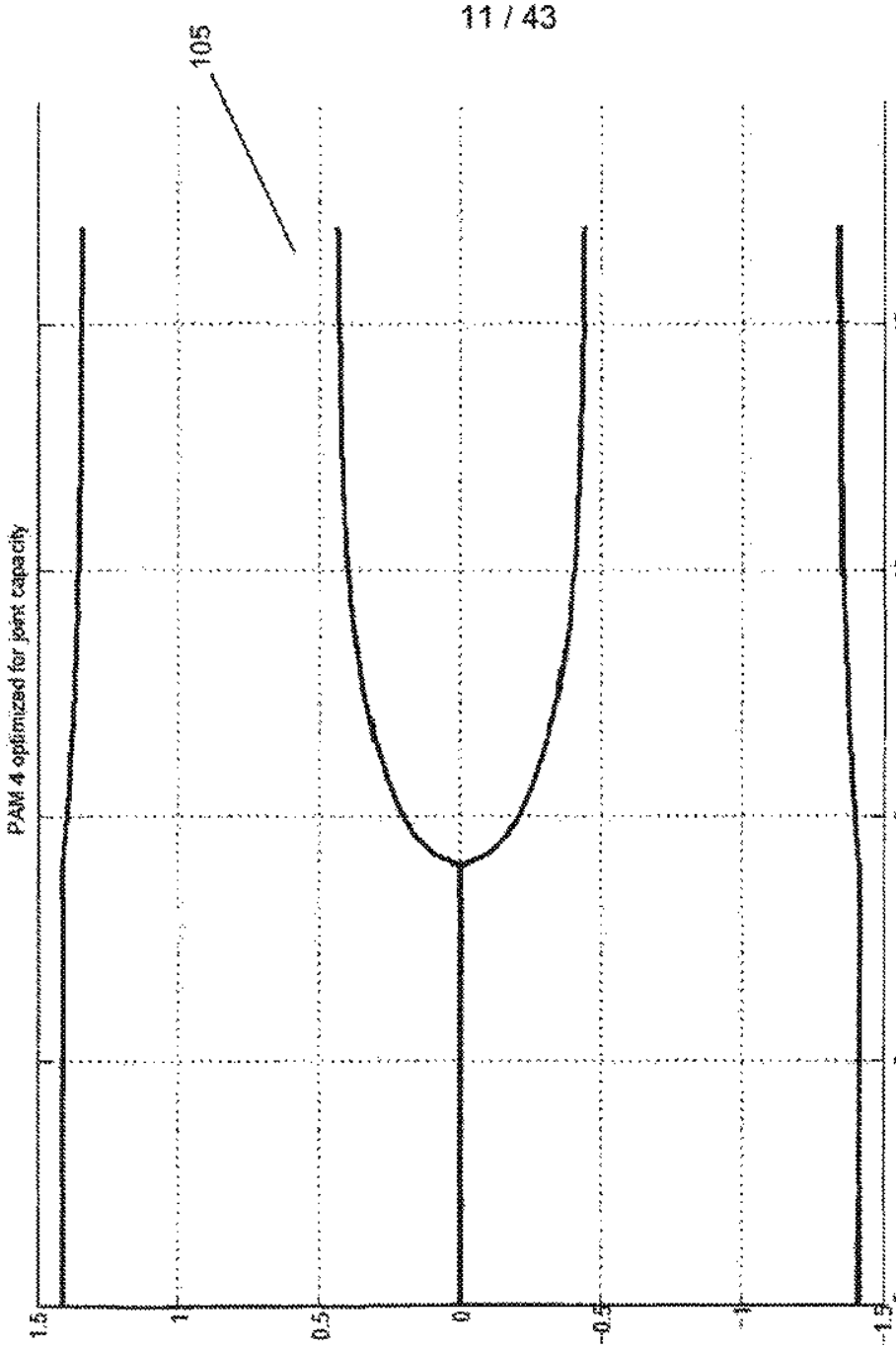

105
PAM 4 optimized for joint capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10b

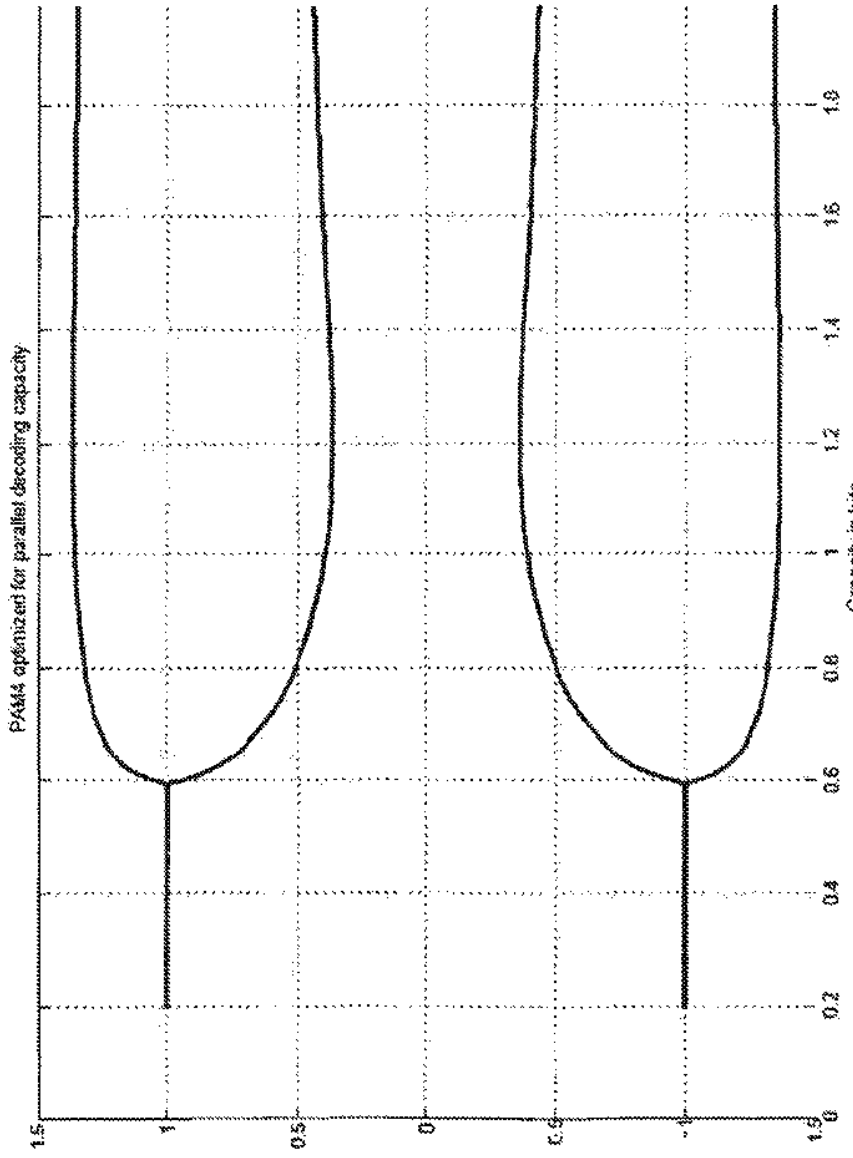

PAM4 optimized for parallel decoding capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10c

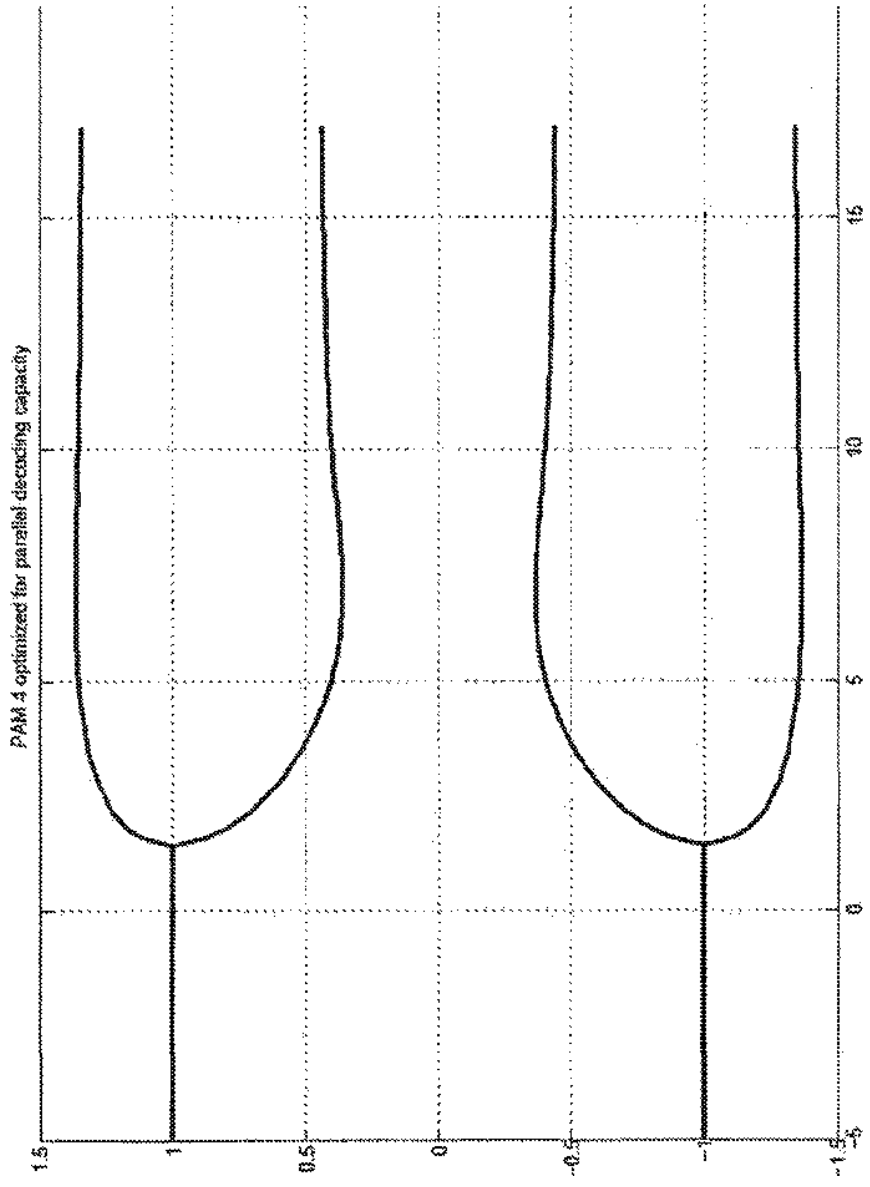

PAM 4 optimized for parallel decoding capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10d

PAM-4 constellations optimized for joint capacity at different rates

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.03	2.71	5.00	7.15	9.24
x_0	-1.41	-1.41	-1.40	-1.37	-1.36
x_1	0.00	0.00	-0.20	-0.33	-0.39
x_2	0.00	0.00	0.20	0.33	0.39
x_3	1.41	1.41	1.40	1.37	1.36

FIG. 11a

PAM-4 constellations optimized for parallel decoding capacity at different

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.19	3.11	5.26	7.22	9.25
x_0	-1.00	-1.30	-1.36	-1.37	-1.36
x_1	-1.00	-0.56	-0.39	-0.33	-0.39
x_2	1.00	1.30	1.36	0.33	1.36
x_3	1.00	0.56	0.39	1.37	0.39

FIG. 11b

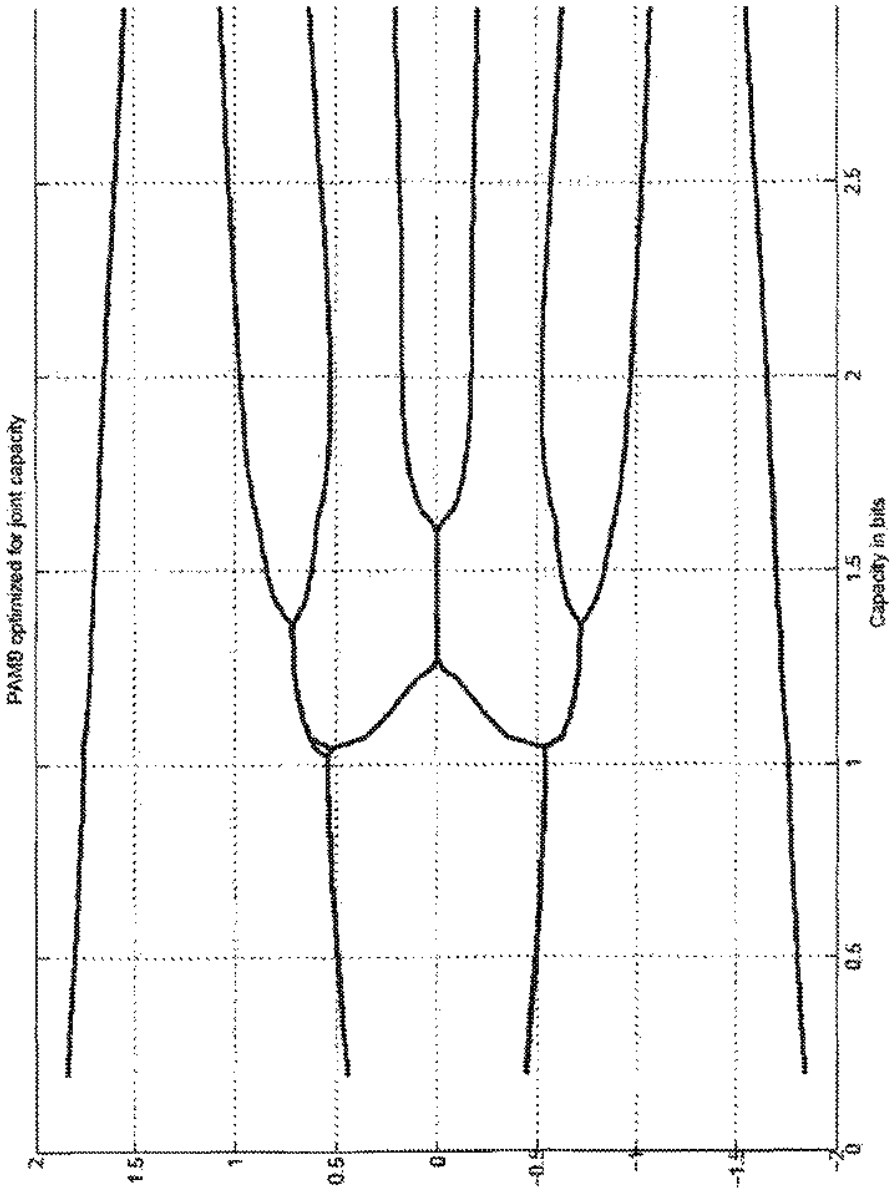

PAM9 optimized for joint capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12a

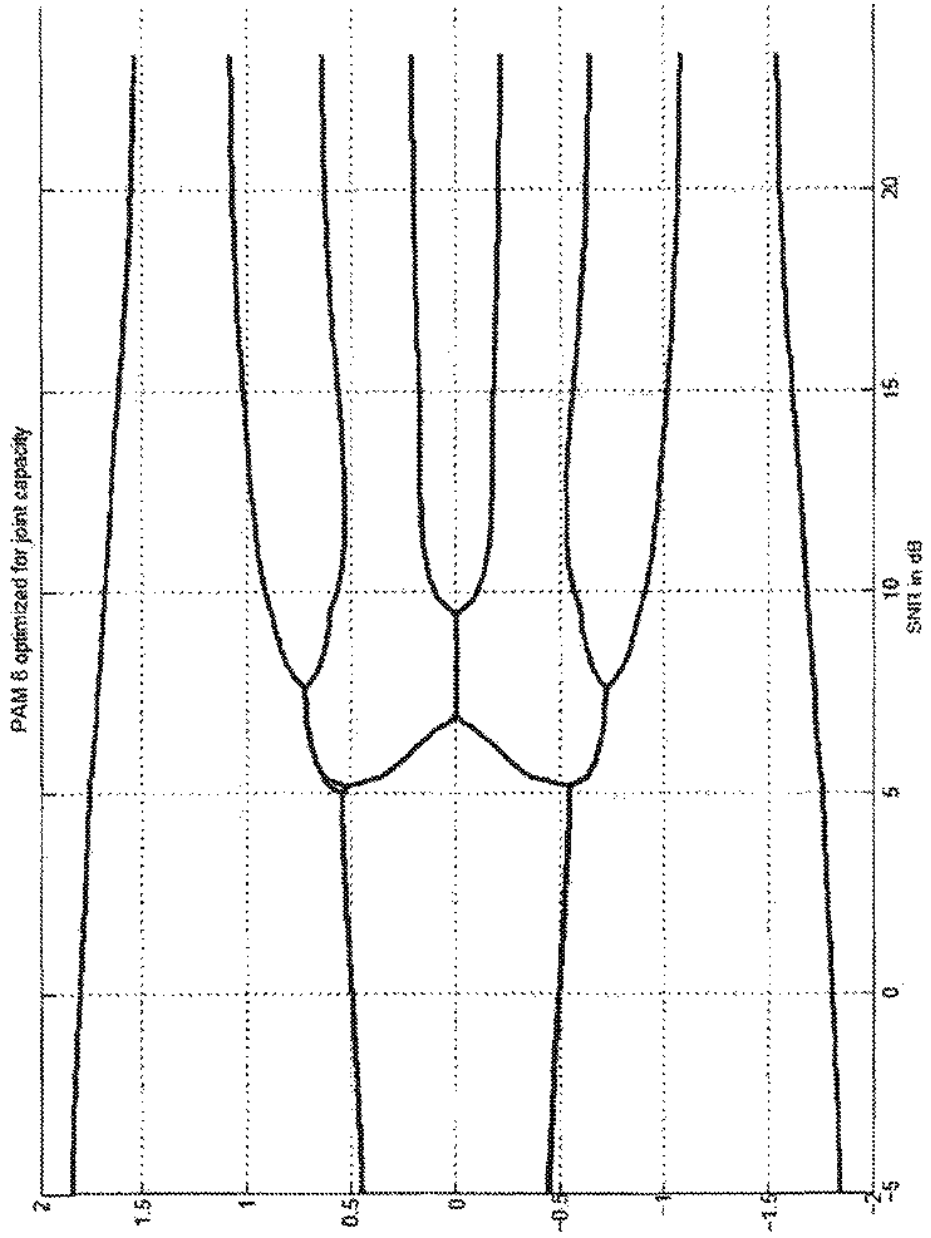

PAM 8 optimized for joint capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12b

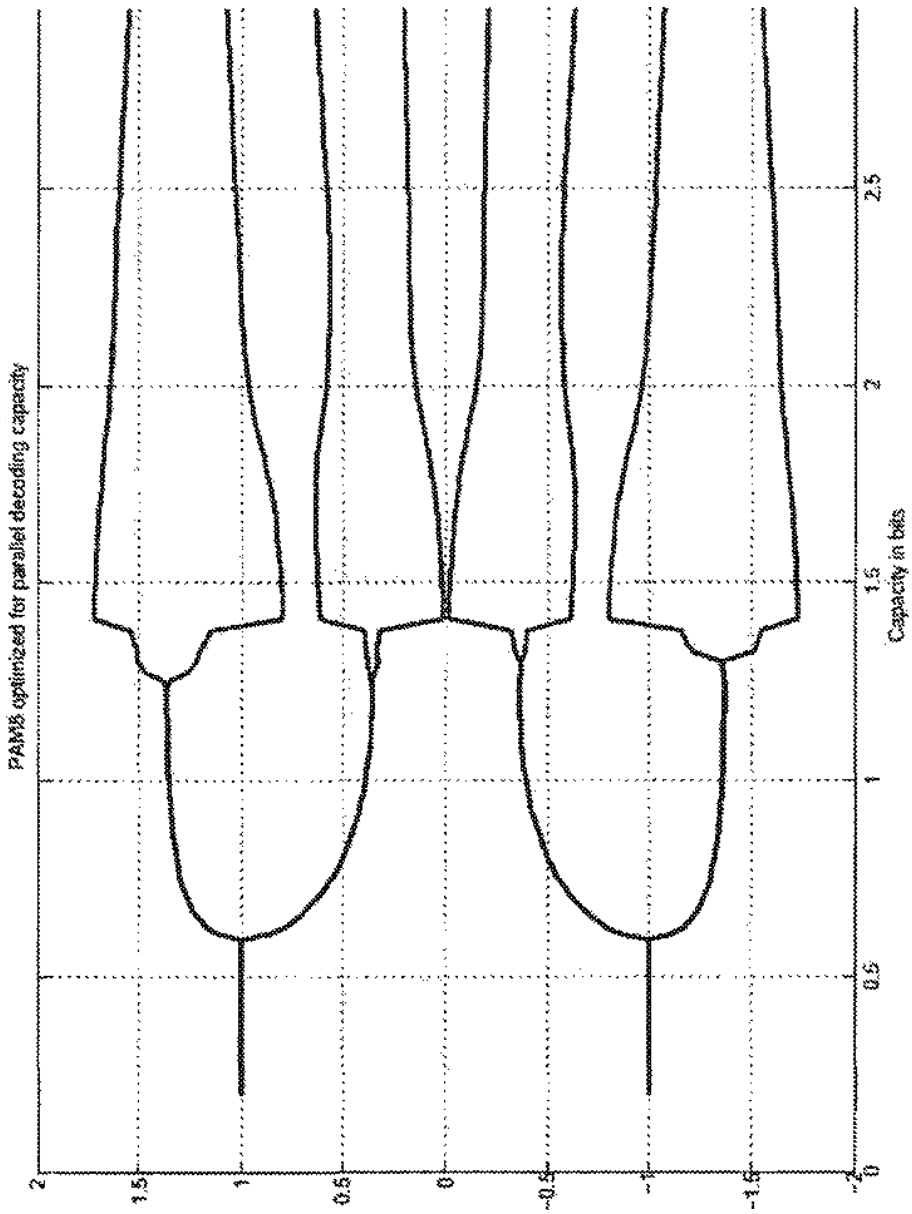

PAM5 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12c

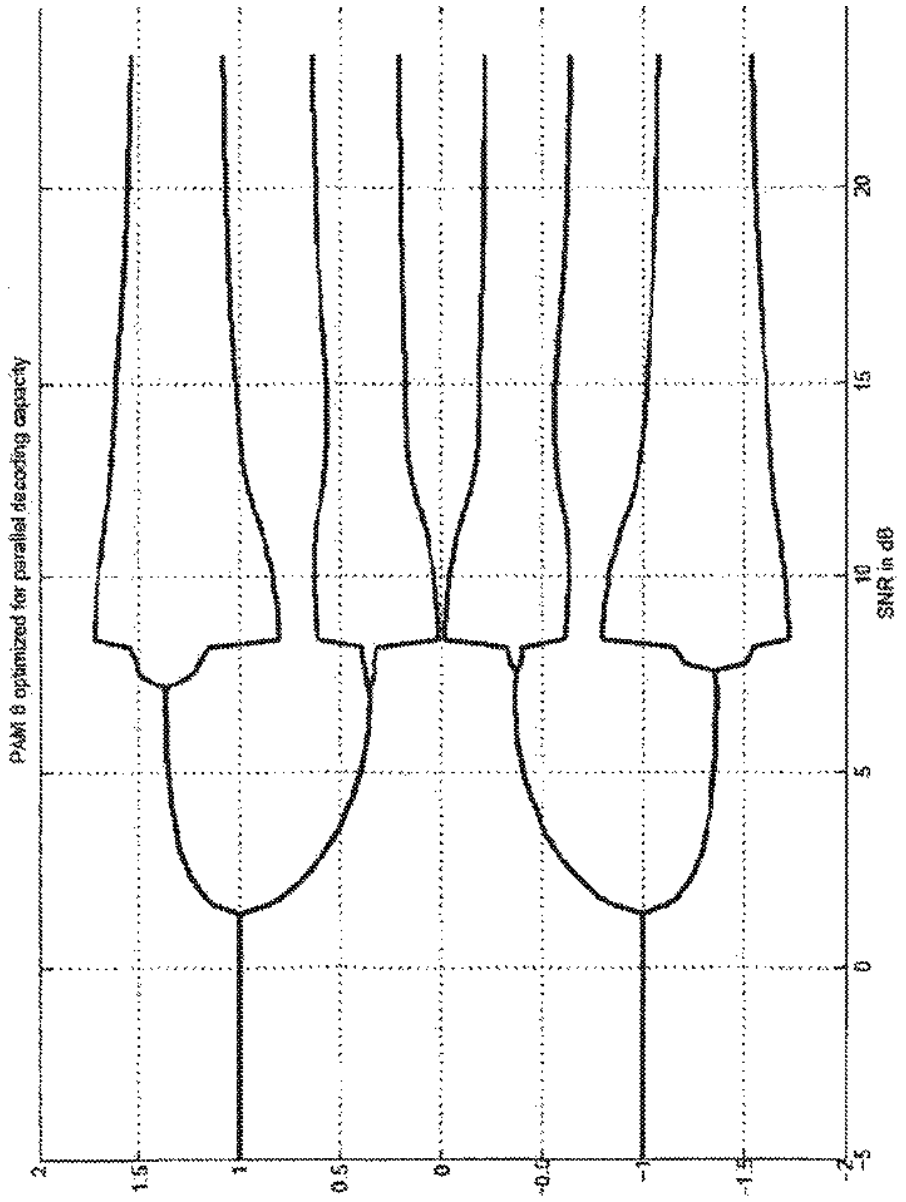

PAR 6 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12d

PAM-8 constellations optimized for joint capacity at different rates

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.00	4.82	8.66	12.26	15.93
x_0	-1.81	-1.76	-1.70	-1.66	-1.60
x_1	-0.50	-0.55	-0.84	-0.97	-1.03
x_2	-0.50	-0.55	-0.63	-0.53	-0.58
x_3	-0.50	-0.55	-0.00	-0.17	-0.19
x_4	0.50	0.55	0.00	0.17	0.19
x_5	0.50	0.55	0.63	0.53	0.58
x_6	0.50	0.55	0.84	0.97	1.03
x_7	1.81	1.76	1.70	1.66	1.60

FIG. 13a

PAM-8 constellations optimized for parallel decoding capacity at different

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.19	5.27	9.00	12.42	15.93
x_0	-1.00	-1.36	-1.72	-1.64	-1.60
x_1	-1.00	-1.36	-0.81	-0.97	-1.03
x_2	-1.00	-0.39	1.72	1.64	-0.19
x_3	-1.00	-0.39	-0.62	-0.58	-0.58
x_4	1.00	1.36	0.62	0.58	1.60
x_5	1.00	1.36	0.02	0.15	1.03
x_6	1.00	0.39	0.81	0.97	0.19
x_7	1.00	0.39	-0.02	-0.15	0.58

FIG. 13b

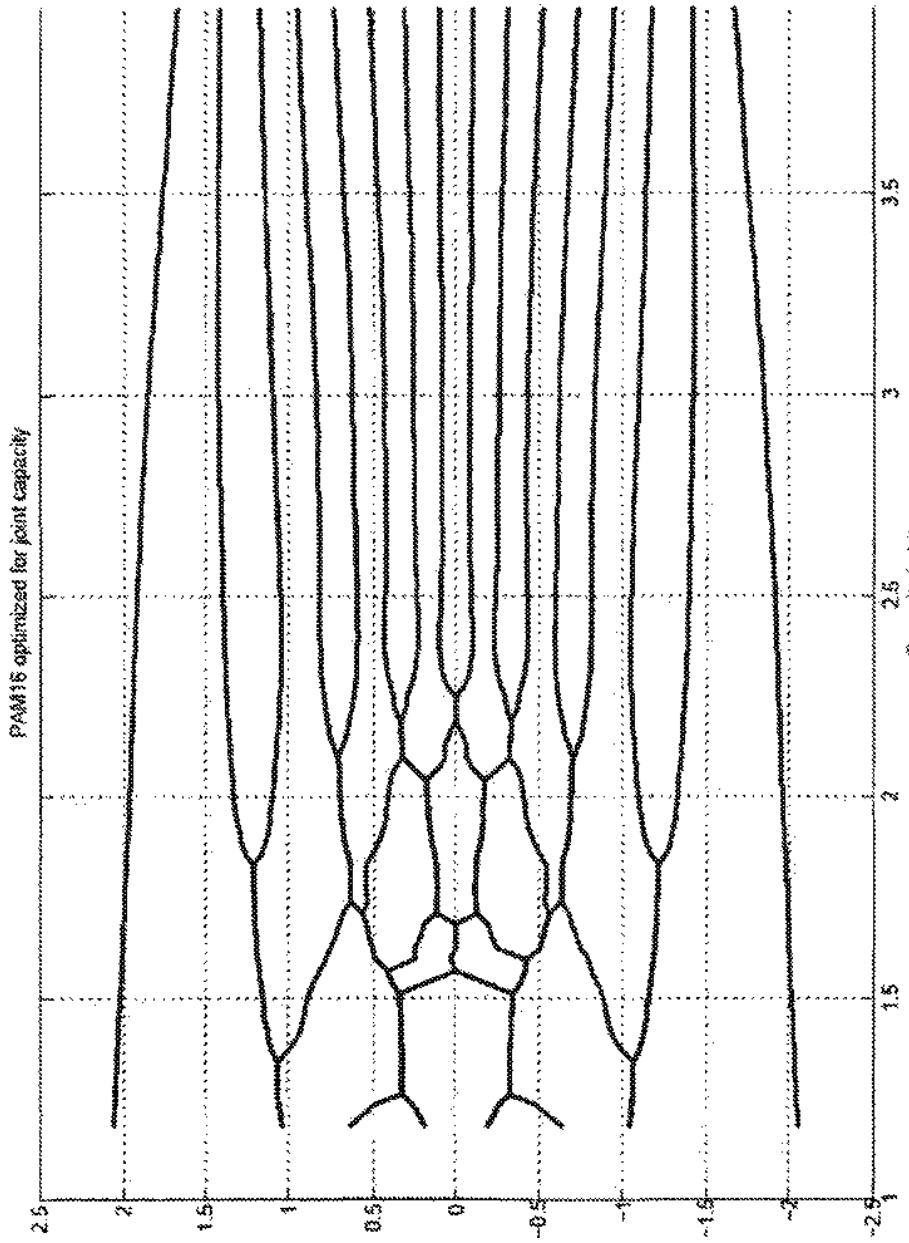

PAM16 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

FIG. 14a

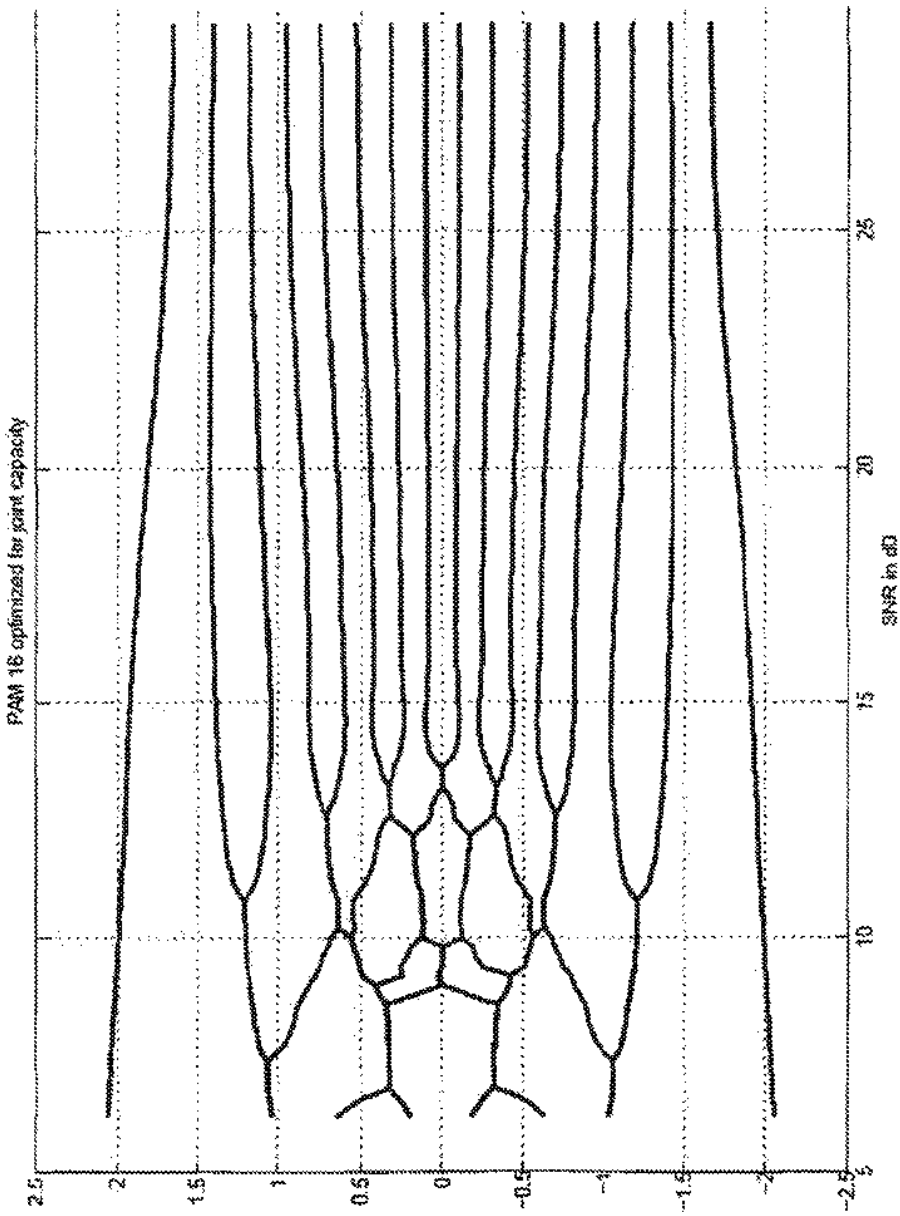

PAM 16 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14b

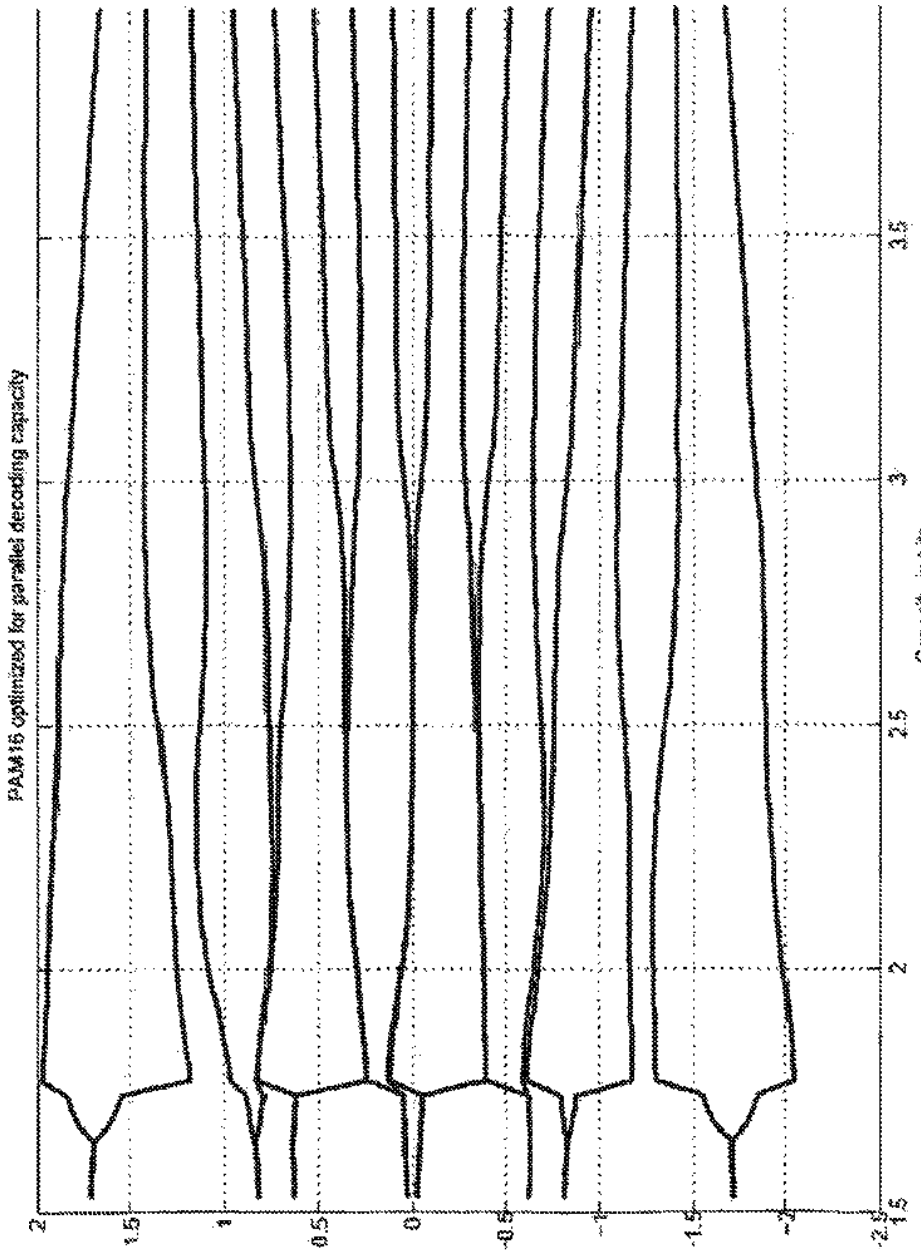

PAM16 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
1.5
2
2.5
3
3.5

FIG. 14c

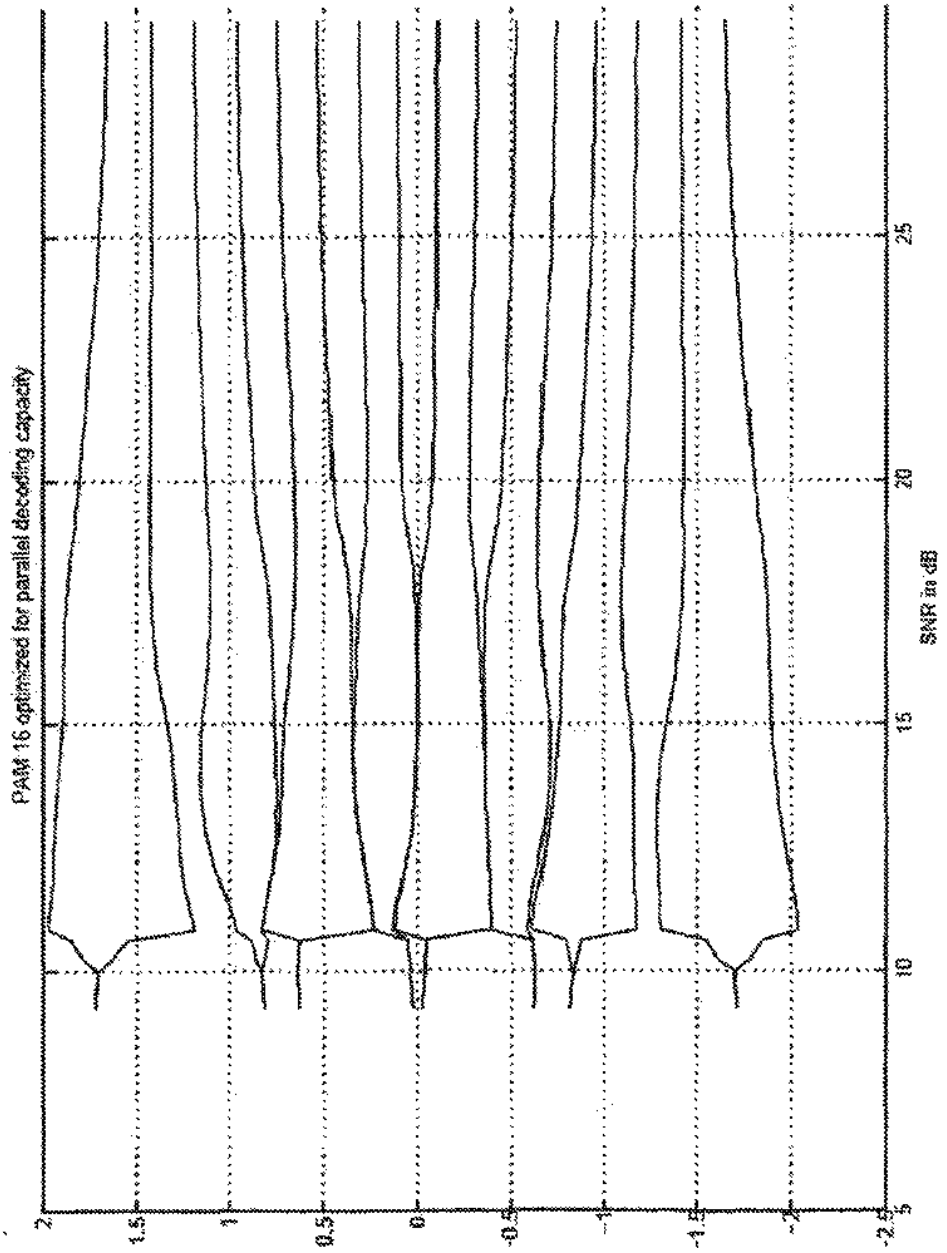

PAM 16 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14d

PAM-16 constellations optimized for joint capacity at different rates

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	8.52	11.94	15.25	18.60	22.12
x_0	-2.02	-1.96	-1.91	-1.85	-1.76
x_1	-1.16	-1.33	-1.40	-1.42	-1.42
x_2	-1.16	-1.10	-1.05	-1.10	-1.15
x_3	-0.90	-0.69	-0.82	-0.84	-0.90
x_4	-0.34	-0.69	-0.60	-0.62	-0.68
x_5	-0.34	-0.40	-0.43	-0.43	-0.47
x_6	-0.34	-0.17	-0.24	-0.26	-0.28
x_7	-0.34	-0.17	-0.09	-0.08	-0.09
x_8	0.34	0.17	0.09	0.08	0.09
x_9	0.34	0.17	0.24	0.26	0.28
x_{10}	0.34	0.40	0.43	0.43	0.47
x_{11}	0.34	0.69	0.60	0.62	0.68
x_{12}	0.90	0.69	0.82	0.84	0.90
x_{13}	1.16	1.10	1.05	1.10	1.15
x_{14}	1.16	1.33	1.40	1.42	1.42
x_{15}	2.02	1.96	1.91	1.85	1.76

FIG. 15a

PAM-16 constellations optimized for parallel decoding capacity at different

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	9.00	12.25	15.42	18.72	22.13
x_0	-1.72	-1.98	-1.89	-1.84	-1.75
x_1	-1.72	-1.29	-1.36	-1.42	-1.42
x_2	-0.81	1.94	1.89	1.84	1.75
x_3	-0.81	-1.17	-1.14	-1.11	-1.15
x_4	1.72	-0.38	-0.35	-0.40	-0.47
x_5	1.72	-0.65	-0.70	-0.65	-0.68
x_6	-0.62	-0.38	-0.34	-0.29	-0.28
x_7	-0.62	-0.68	-0.76	-0.83	-0.90
x_8	0.62	1.09	1.13	1.11	1.15
x_9	0.62	0.76	0.76	0.84	0.90
x_{10}	0.02	1.26	1.35	1.42	1.42
x_{11}	0.02	0.76	0.70	0.65	0.68
x_{12}	0.81	0.06	0.00	0.05	0.09
x_{13}	0.81	0.29	0.34	0.29	0.28
x_{14}	-0.02	0.06	0.00	-0.05	-0.09
x_{15}	-0.02	0.29	0.35	0.40	0.47

FIG. 15b

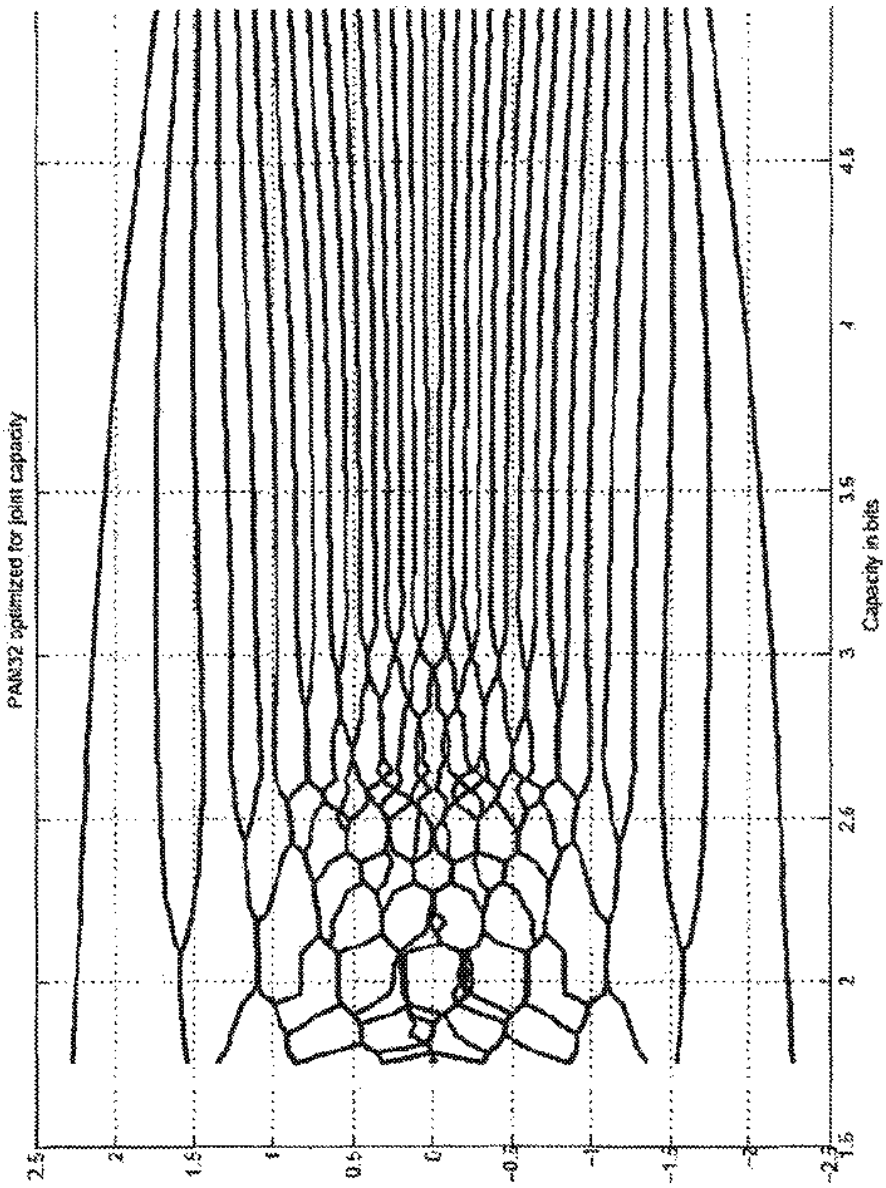

PAM32 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.3
1.5
2
2.5
3
3.5
4
4.5

FIG. 16a

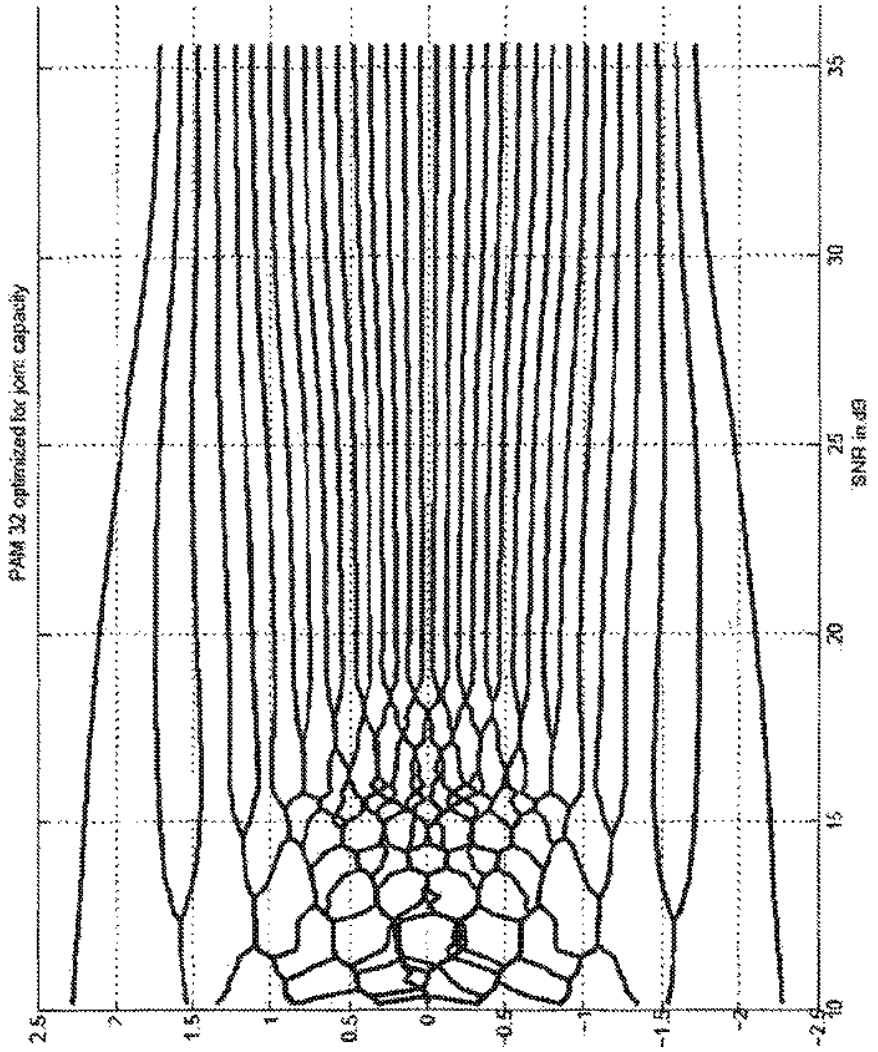

PAM 32 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

FIG. 16b

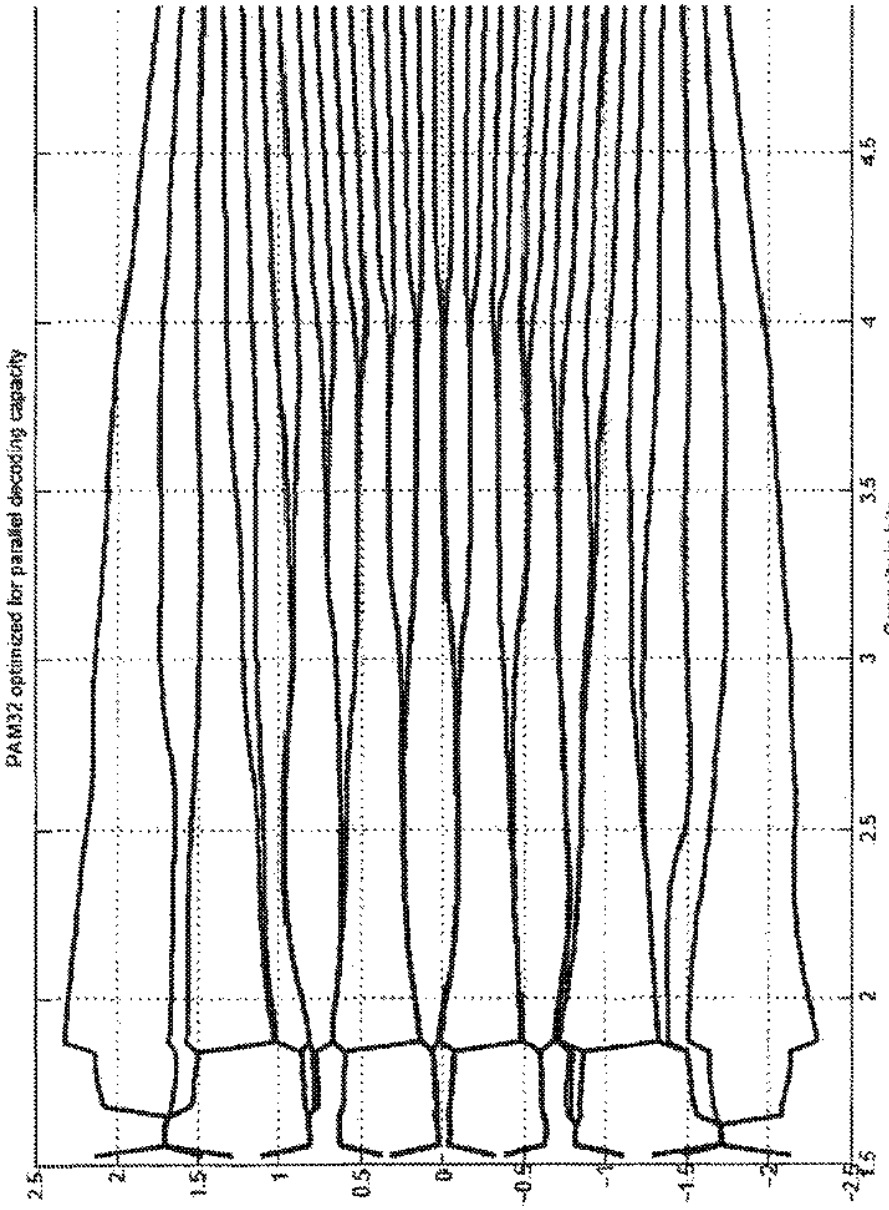

PAM32 optimized for parallel decoding capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16c

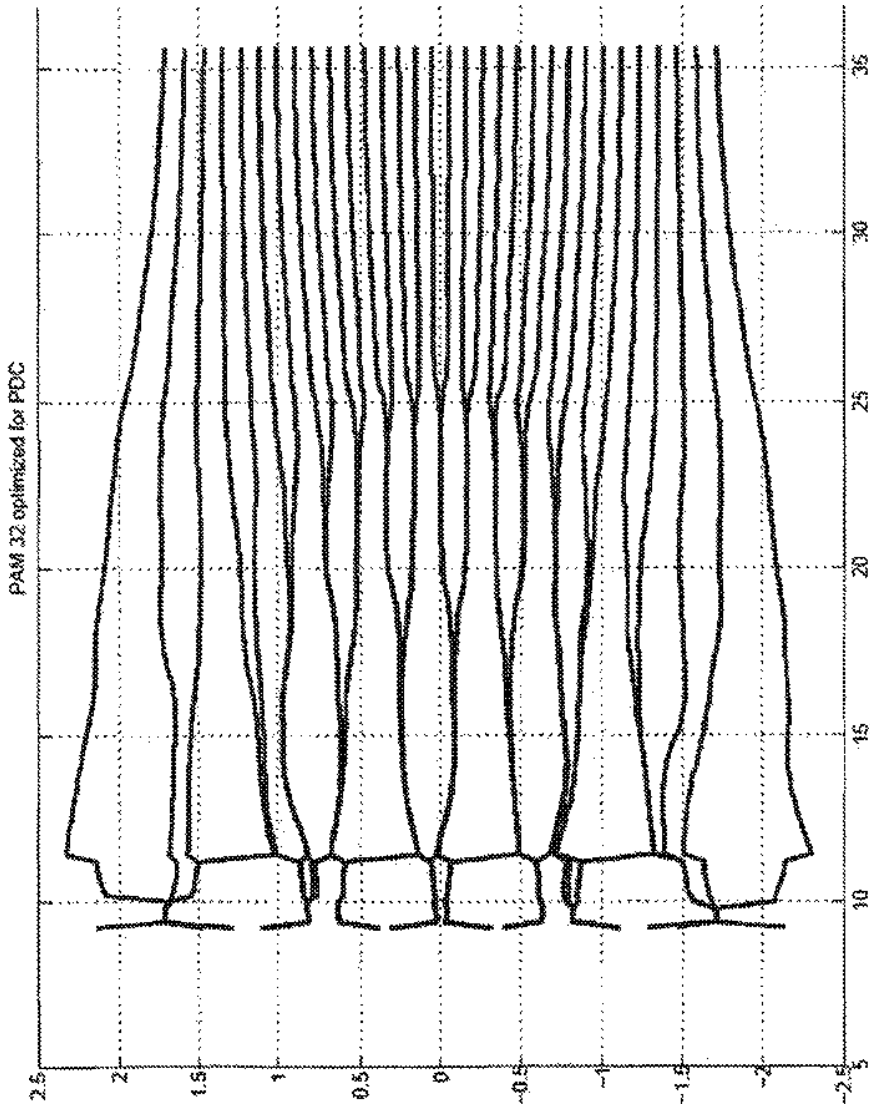

PAM 32 optimized for PDC
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

FIG. 16d

PAM-32 constellations optimized for joint capacity at different rates

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	11.83	13.05	18.23	21.42	24.69	28.19
x_0	-2.25	-2.19	-2.14	-2.07	-1.97	-1.85
x_1	-1.58	-1.71	-1.74	-1.74	-1.72	-1.66
x_2	-1.58	-1.46	-1.46	-1.49	-1.51	-1.50
x_3	-1.10	-1.23	-1.27	-1.29	-1.33	-1.35
x_4	-1.10	-1.13	-1.11	-1.13	-1.17	-1.21
x_5	-1.10	-0.90	-0.98	-0.99	-1.02	-1.08
x_6	-0.83	-0.90	-0.85	-0.87	-0.90	-0.95
x_7	-0.60	-0.75	-0.75	-0.76	-0.78	-0.84
x_8	-0.60	-0.58	-0.63	-0.65	-0.67	-0.73
x_9	-0.60	-0.58	-0.57	-0.56	-0.57	-0.62
x_{10}	-0.60	-0.49	-0.42	-0.46	-0.48	-0.52
x_{11}	-0.24	-0.29	-0.40	-0.38	-0.39	-0.42
x_{12}	-0.21	-0.28	-0.24	-0.29	-0.30	-0.32
x_{13}	-0.20	-0.28	-0.24	-0.21	-0.21	-0.23
x_{14}	-0.20	-0.09	-0.09	-0.12	-0.13	-0.14
x_{15}	-0.16	-0.00	-0.07	-0.04	-0.04	-0.05
x_{16}	0.16	0.00	0.07	0.04	0.04	0.05
x_{17}	0.19	0.09	0.09	0.12	0.13	0.14
x_{18}	0.21	0.28	0.24	0.21	0.21	0.23
x_{19}	0.22	0.28	0.24	0.29	0.30	0.32
x_{20}	0.23	0.28	0.41	0.38	0.39	0.42
x_{21}	0.60	0.49	0.42	0.46	0.48	0.52
x_{22}	0.60	0.58	0.57	0.56	0.57	0.62
x_{23}	0.60	0.58	0.62	0.65	0.67	0.73
x_{24}	0.60	0.75	0.75	0.76	0.78	0.84
x_{25}	0.83	0.90	0.85	0.87	0.90	0.95
x_{26}	1.10	0.90	0.98	0.99	1.02	1.08
x_{27}	1.10	1.13	1.11	1.13	1.17	1.21
x_{28}	1.10	1.23	1.27	1.29	1.33	1.35
x_{29}	1.58	1.46	1.46	1.49	1.51	1.50
x_{30}	1.58	1.71	1.74	1.74	1.72	1.66
x_{31}	2.25	2.19	2.14	2.07	1.97	1.85

FIG. 17a

PAM-32 constellations optimized for parallel decoding capacity at differen

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	12.21	15.27	18.42	21.52	24.79	28.20
x_0	-2.25	-2.16	-2.14	-2.05	-1.97	-1.85
x_1	-1.52	-1.64	-1.75	-1.74	-1.72	-1.66
x_2	2.30	2.19	-1.31	2.05	1.97	-1.35
x_3	-1.39	-1.48	-1.43	-1.49	-1.51	-1.49
x_4	1.56	1.54	2.14	-0.96	-1.03	1.85
x_5	-1.31	-1.23	1.75	-1.15	-1.17	1.66
x_6	1.67	1.65	-1.07	-0.91	-0.90	-1.21
x_7	-1.31	-1.24	-1.04	-1.28	-1.33	-1.08
x_8	-0.48	-0.43	-0.36	-0.17	-0.17	-0.42
x_9	-0.72	-0.76	-0.36	-0.34	-0.31	-0.52
x_{10}	-0.48	-0.43	-0.62	-0.17	-0.15	-0.73
x_{11}	-0.73	-0.76	-0.62	-0.34	-0.35	-0.62
x_{12}	-0.48	-0.42	-0.29	-0.71	-0.67	-0.33
x_{13}	-0.76	-0.86	-0.29	-0.52	-0.55	-0.23
x_{14}	-0.48	-0.42	-0.77	-0.72	-0.77	-0.84
x_{15}	-0.76	-0.86	-0.77	-0.52	-0.48	-0.96
x_{16}	0.87	0.98	1.07	1.49	1.51	1.21
x_{17}	0.66	0.63	1.04	1.28	1.33	1.08
x_{18}	0.87	0.98	0.77	1.74	1.72	0.84
x_{19}	0.66	0.63	0.77	1.15	1.17	0.96
x_{20}	1.07	1.13	1.31	0.72	0.77	1.35
x_{21}	0.66	0.59	1.43	0.91	0.90	1.49
x_{22}	1.05	1.10	0.62	0.71	0.67	0.73
x_{23}	0.66	0.60	0.62	0.96	1.03	0.62
x_{24}	-0.01	-0.08	0.02	0.00	0.01	0.05
x_{25}	0.17	0.25	0.02	0.17	0.15	0.14
x_{26}	-0.01	-0.08	0.29	0.00	-0.01	0.33
x_{27}	0.17	0.25	0.29	0.17	0.17	0.23
x_{28}	-0.01	-0.08	-0.02	0.52	0.48	-0.05
x_{29}	0.17	0.25	-0.02	0.34	0.35	-0.14
x_{30}	-0.01	-0.08	0.36	0.52	0.55	0.42
x_{31}	0.17	0.25	0.36	0.34	0.31	0.52

FIG. 17b

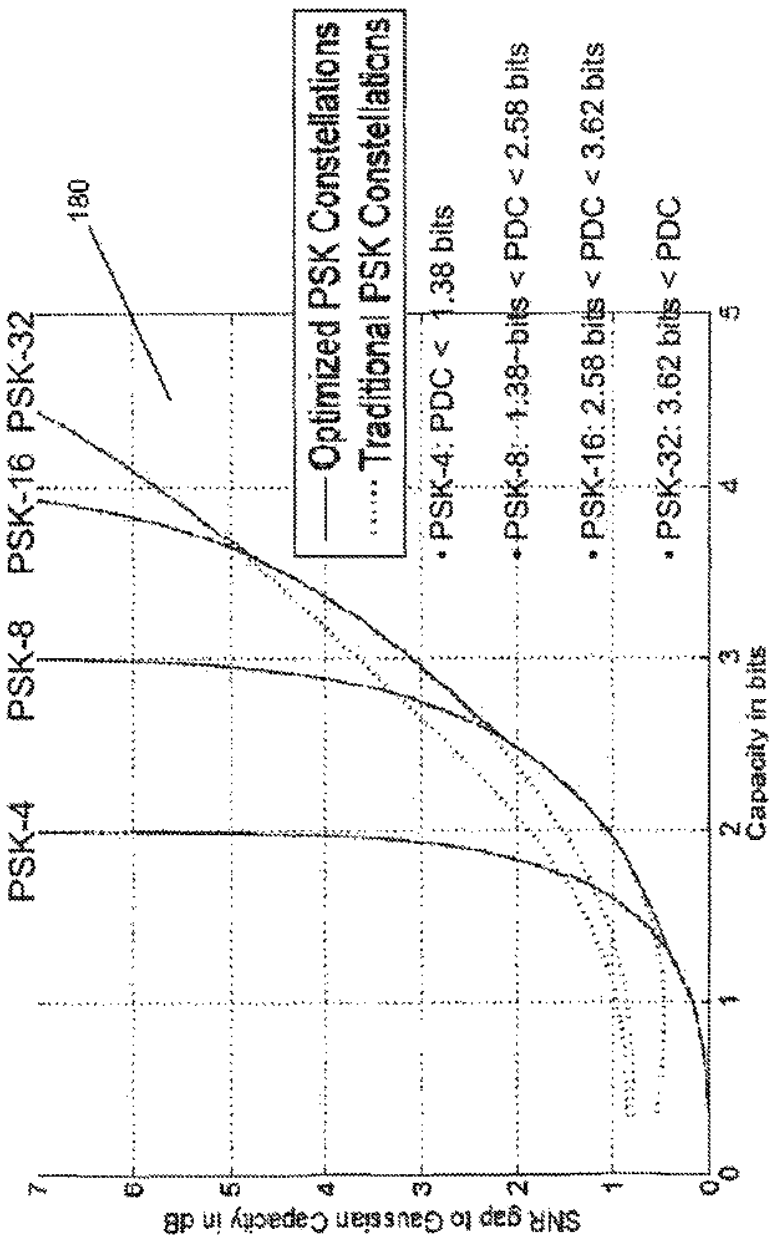

180
PSK-4
PSK-8
PSK-16
PSK-32
Optimized PSK Constellations
Traditional PSK Constellations
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38 bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
SNR gap to Gaussian Capacity in dB
Capacity in bits
0
1
2
3
4
5
6
7

FIG. 18

Points location in degrees
0
50
100
150
200
250
300
350
SNR in dB
0
2
4
6
8
10
12
14
16
18
20

FIG. 19

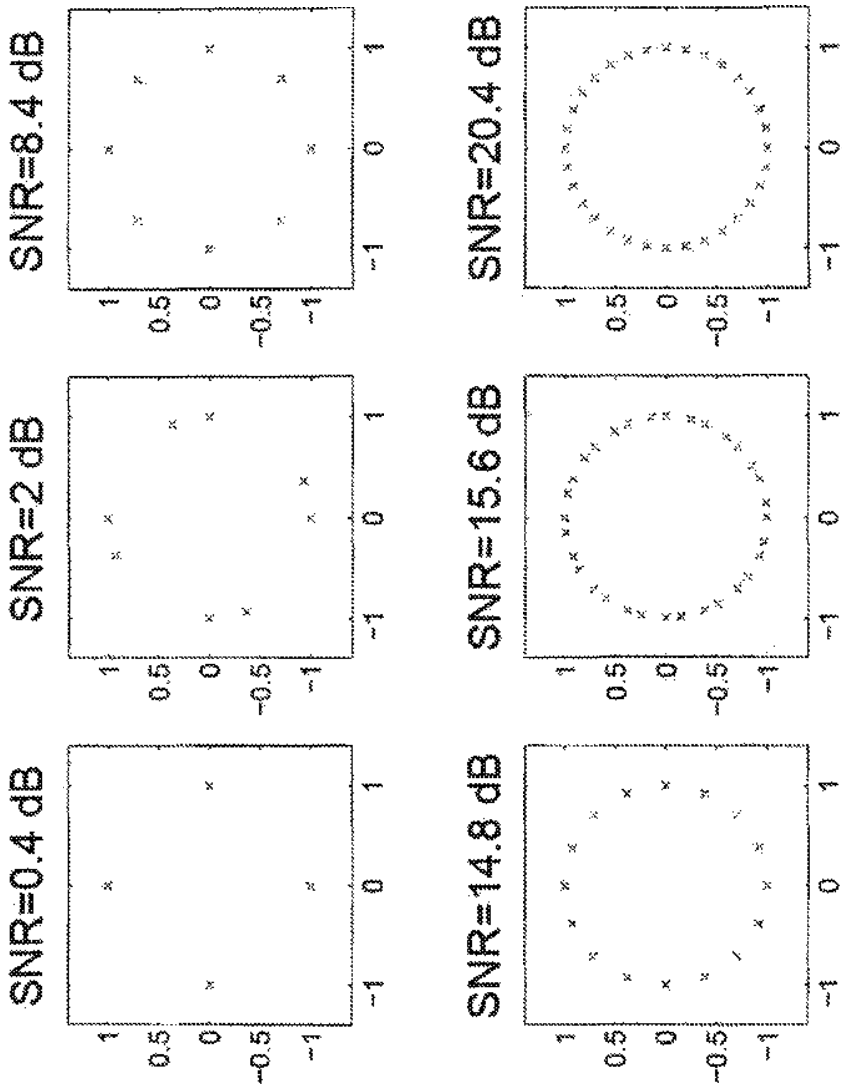


FIG. 20

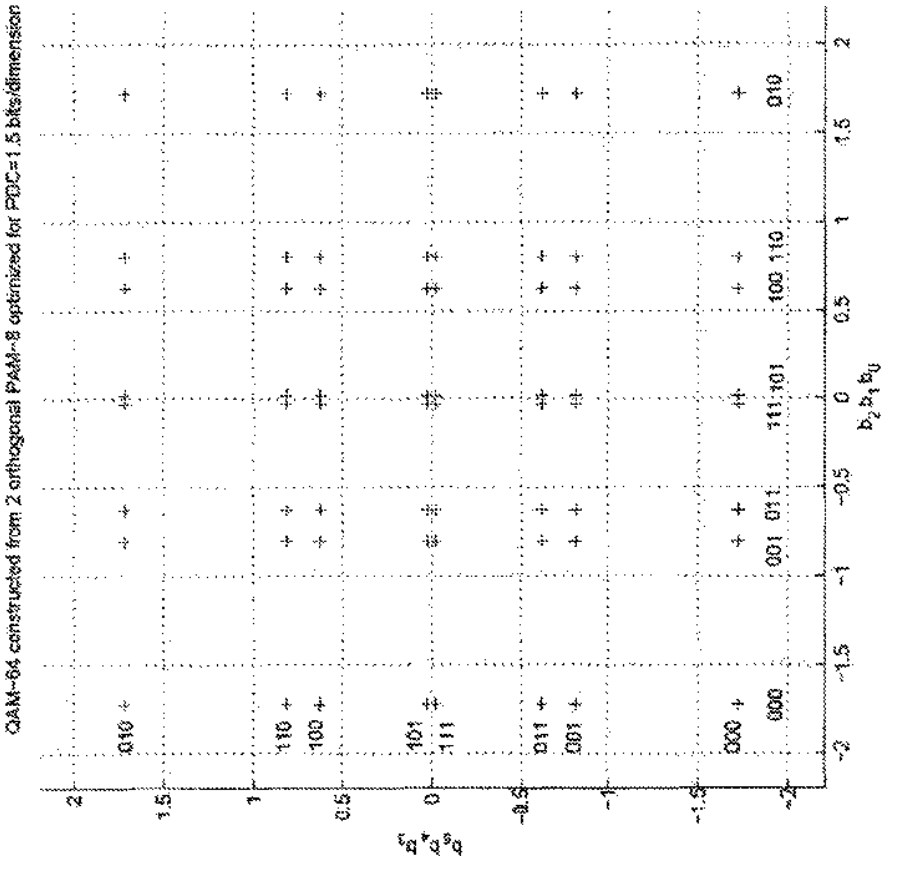

QAM-64 constructed from 2 orthogonal PAM-8 optimized for PDC=1.5 bits/dimension
$b_2 b_1 b_0$
$b_5 b_4 b_3$
010
110
100
101
111
011
001
000
000
001
011
111
101
100
110
010

FIG. 21

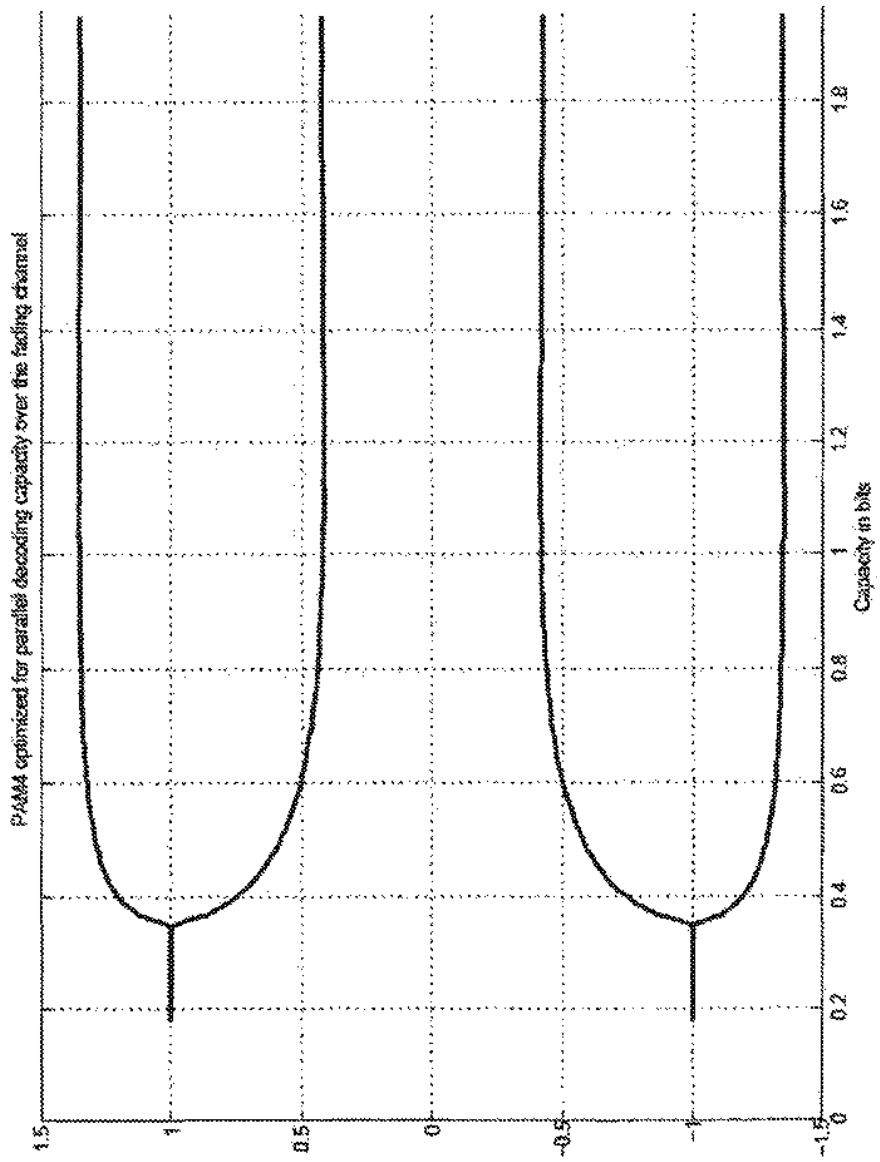

PAM4 optimized for parallel decoding capacity over the fading channel
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 22a

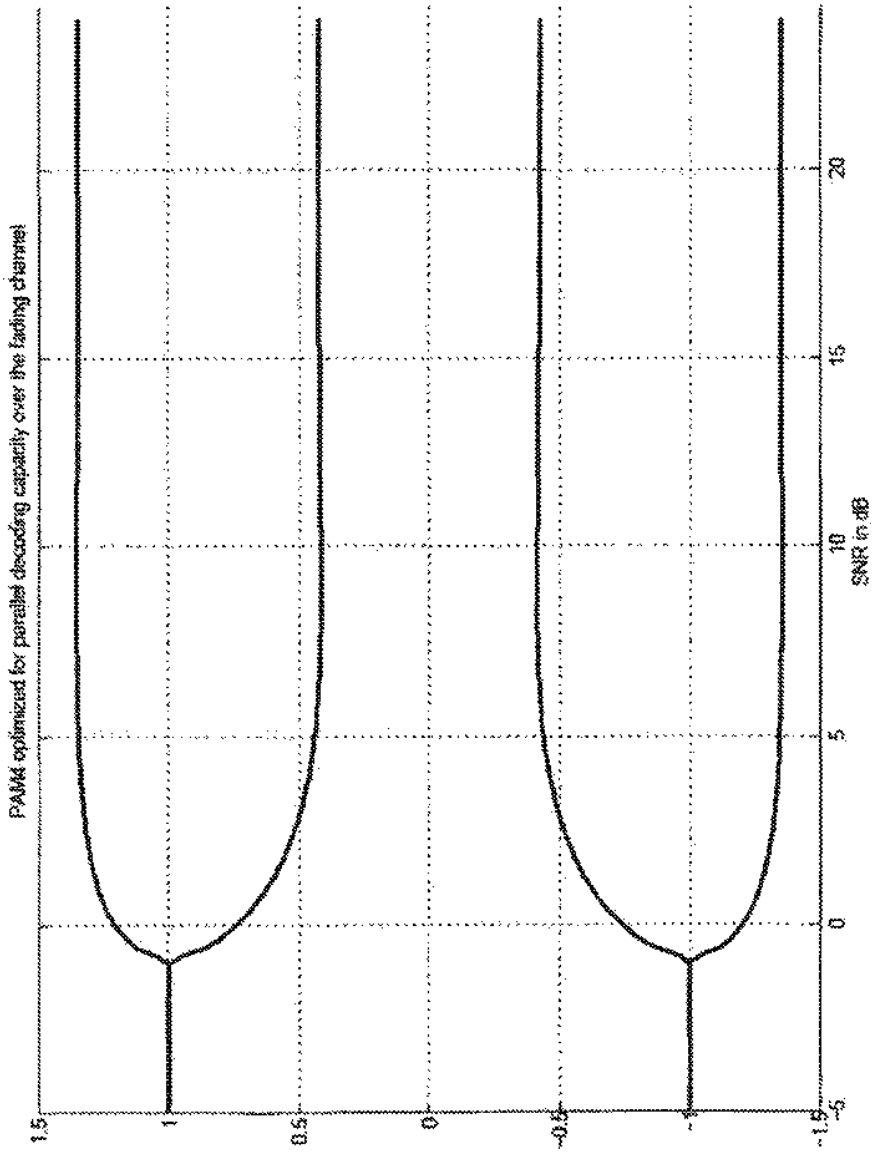

PAM4 optimized for parallel decoding capacity over the fading channel
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20

FIG. 22b

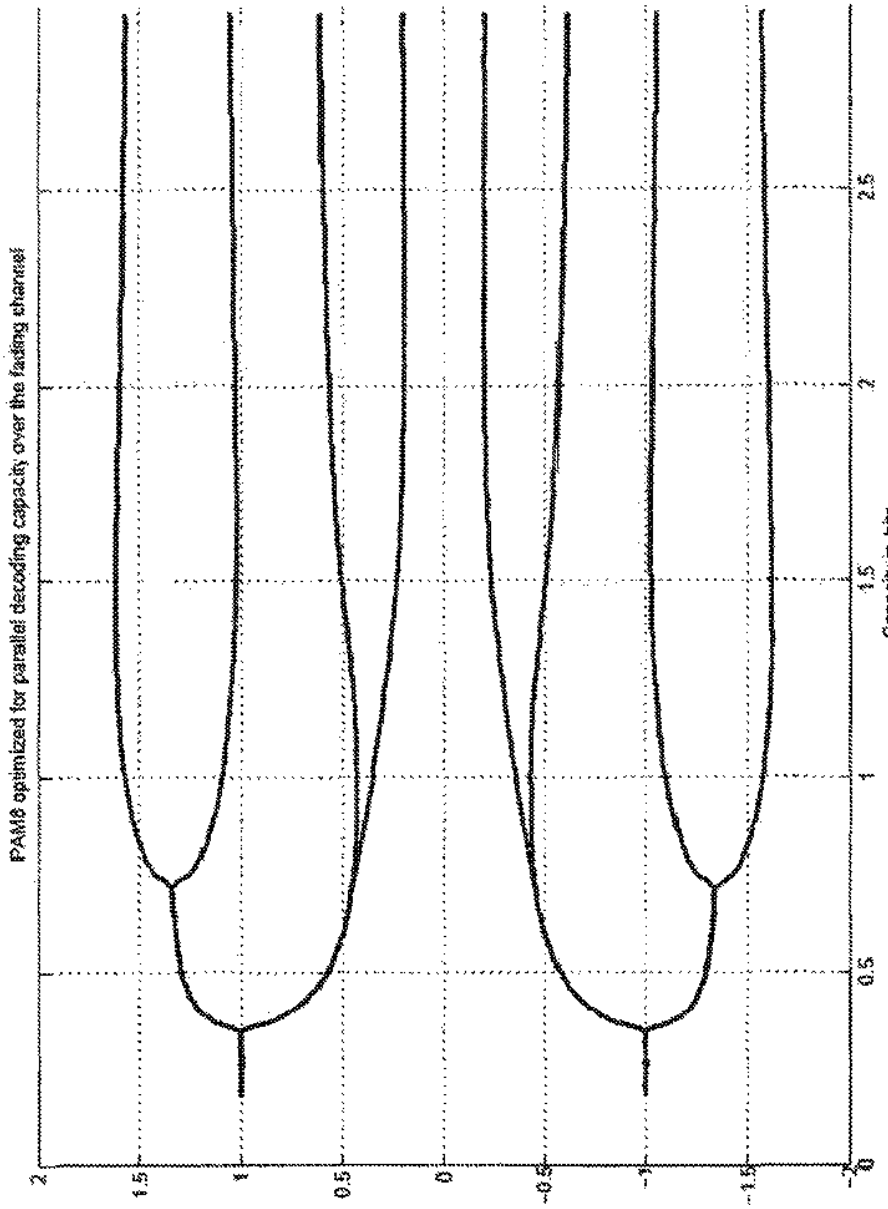

PAM8 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 23a

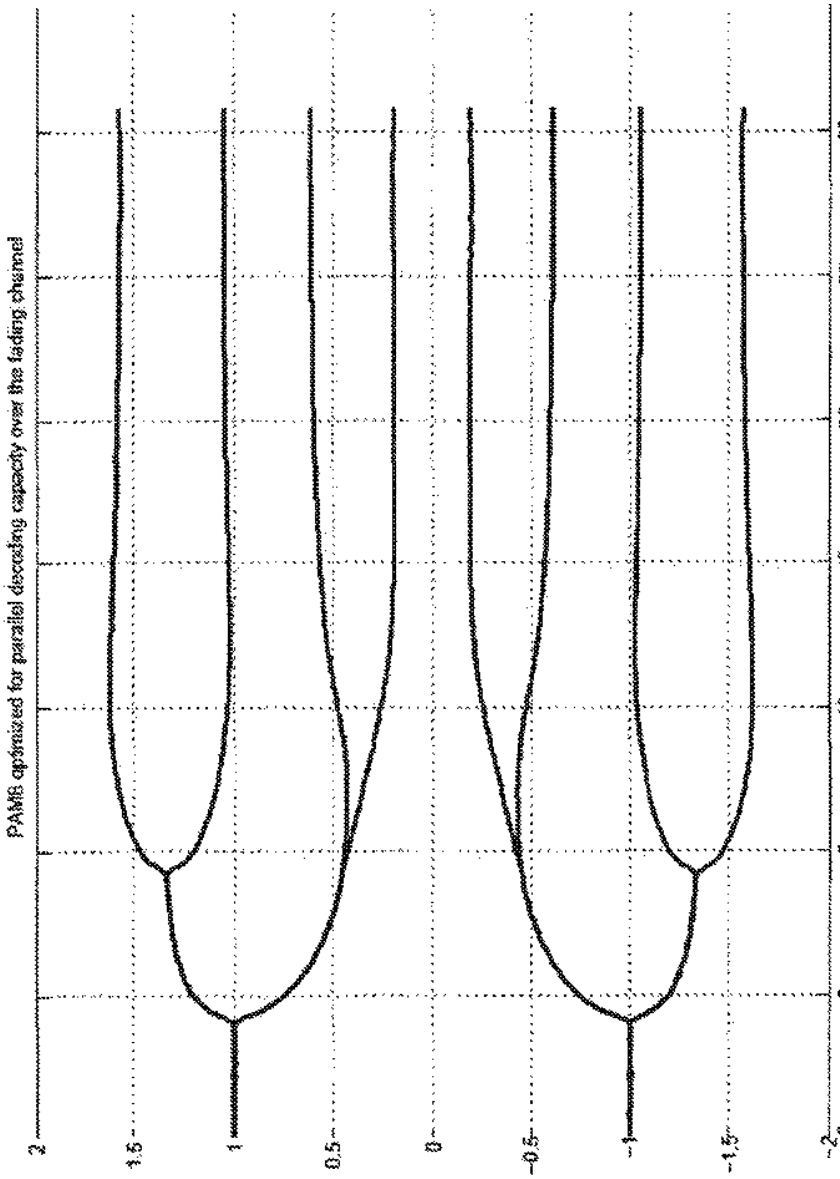

PAMS optimized for parallel decoding capacity over the fading channel
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30
SNR in dB

FIG. 23b

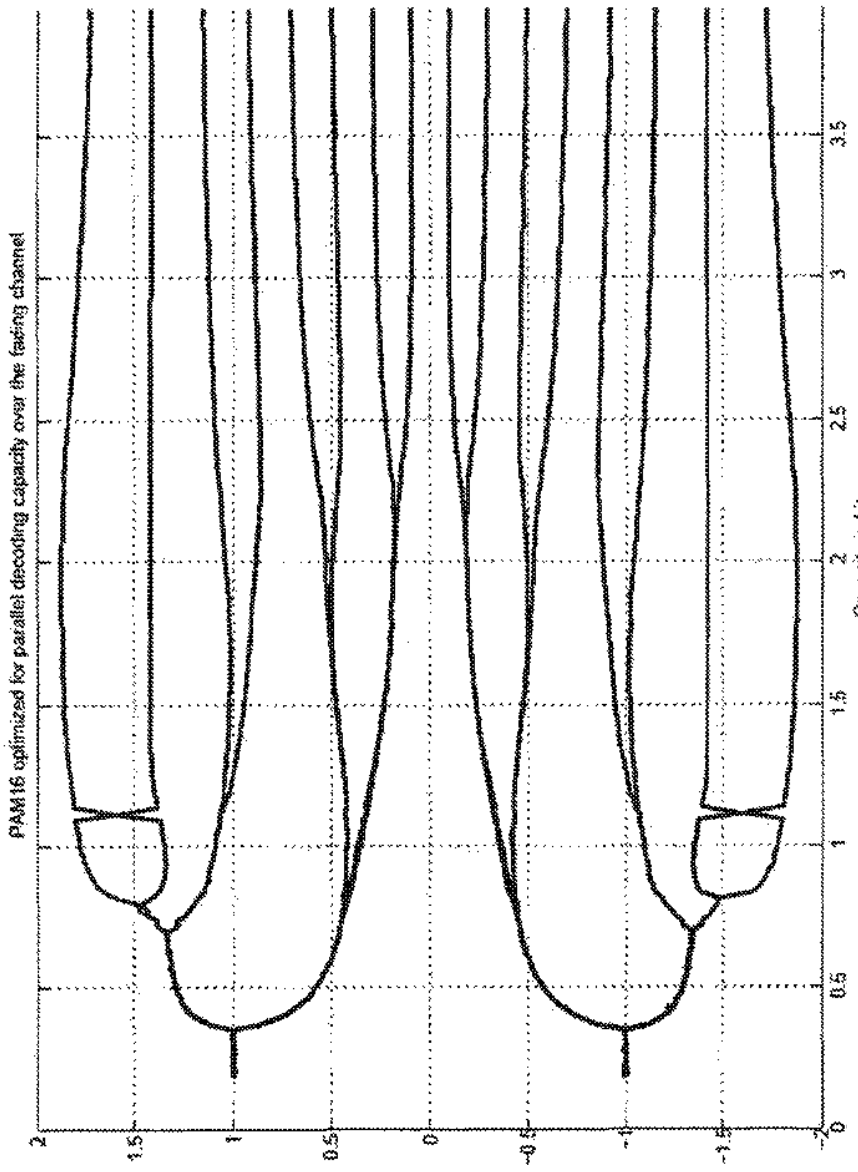

PAM16 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

FIG. 24a

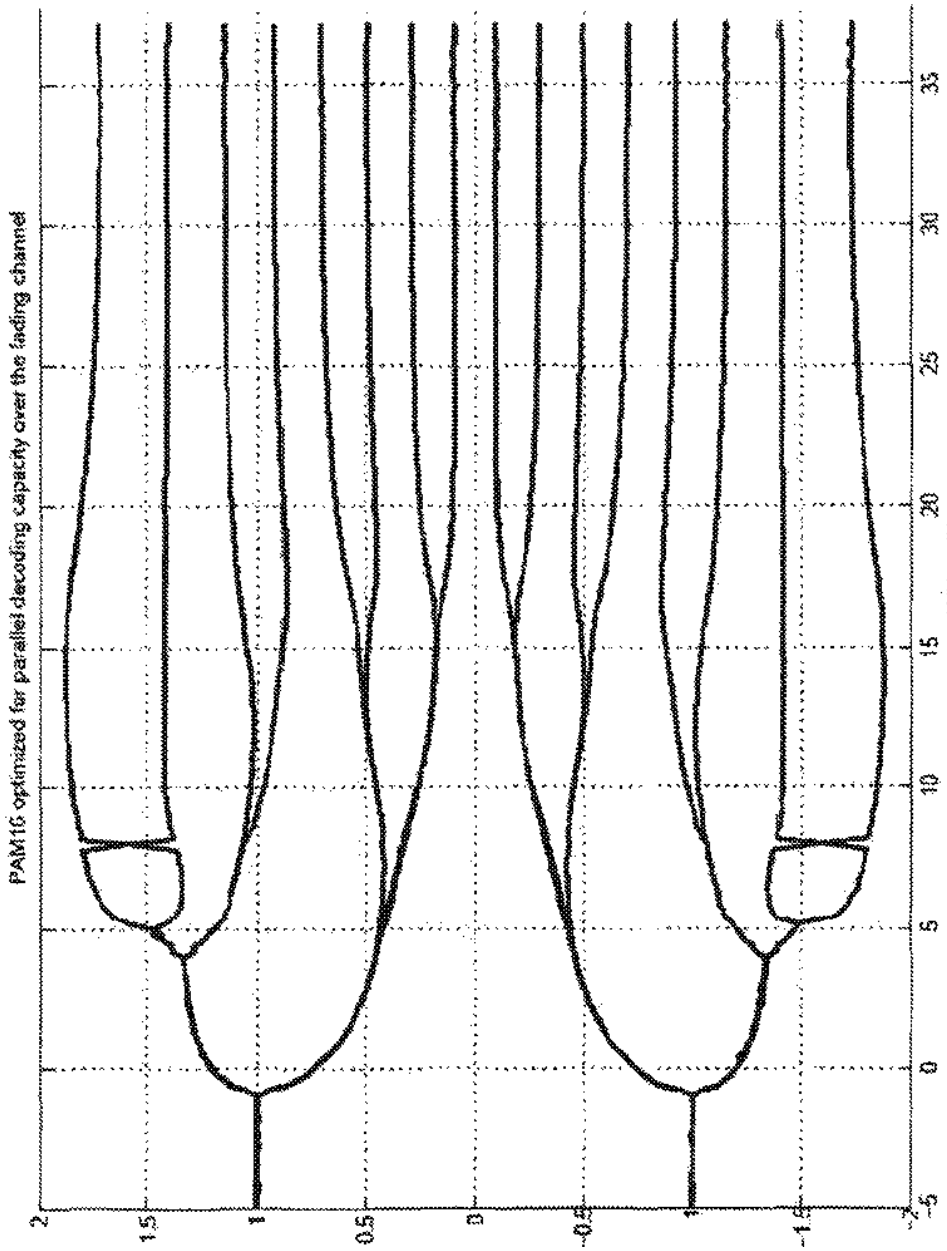

PAM16 optimized for parallel decoding capacity over the fading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
35
30
25
20
15
10
5
0
-5

FIG. 24b

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US2008/055994

A. CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H04J 11/00 (2008.04)
USPC - 370/207
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC(8) - H04J 11/00 (2008.04)
USPC - 370/207

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
PatBase, Google Patents

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X --- Y	US 2004/0022179 A1 (GIANNAKIS et al) 05 February 2004 (05.02.2004) entire document	9-11 --- 1-8, 12
Y	US 2007/0116161 A1 (TOKORO et al) 24 May 2007 (24.05.2007) entire document	1-8, 12
Y	US 6,603,801 B1 (ANDREN et al) 05 August 2003 (05.08.2003) entire document	4
Y	US 2005/0156643 A1 (GLASS et al) 13 July 2005 (13.07.2005) entire document	3

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
03 October 2008	22 OCT 2008
Name and mailing address of the ISA/US Mail Stop PCT, Attn: ISA/US, Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Authorized officer: Blaine R. Copenheaver PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/210 (second sheet) (April 2005)

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 3 518 485 A1

(12) EUROPEAN PATENT APPLICATION

(43) Date of publication:
31.07.2019 Bulletin 2019/31

(51) Int Cl:
H04L 27/34 (2006.01) *H04L 1/00* (2006.01)

(21) Application number: 18212572.4

(22) Date of filing: 05.06.2008

(84) Designated Contracting States:
AT BE BG CH CY CZ DE DK EE ES FI FR GB GR HR HU IE IS IT LI LT LU LV MC MT NL NO PL PT RO SE SI SK TR

(30) Priority: 05.06.2007 US 93331907 P

(62) Document number(s) of the earlier application(s) in accordance with Art. 76 EPC:
08795885.6 / 2 153 581

(71) Applicant: **Constellation Designs, Inc.**
Pacific Palisades, CA 90272 (US)

(72) Inventors:
- **BARSOUM, Maged F.**
Pacific Palisades, CA 90272 (US)
- **JONES, Christopher Richard**
Pacific Palisades, CA 90272 (US)

(74) Representative: **Derry, Paul Stefan et al**
Venner Shipley LLP
200 Aldersgate
London EC1A 4HD (GB)

Remarks:
•This application was filed on 14-12-2018 as a divisional application to the application mentioned under INID code 62.
•Claims filed after the date of filing of the application (Rule 68(4) EPC).
•Amended claims in accordance with Rule 137(2) EPC.

(54) METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS

(57) Communication systems are described that use geometrically shaped constellations that have increased capacity compared to conventional constellations operating within a similar SNR band. In several embodiments, the geometrically shaped is optimized based upon a capacity measure such as parallel decoding capacity or joint capacity. In many embodiments, a capacity optimized geometrically shaped constellation can be used to replace a conventional constellation as part of a firmware upgrade to transmitters and receivers within a communication system. In a number of embodiments, the geometrically shaped constellation is optimized for an Additive White Gaussian Noise channel or a fading channel. In numerous embodiments, the communication uses adaptive rate encoding and the location of points within the geometrically shaped constellation changes as the code rate changes. One embodiment of the invention includes a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter includes a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate, a mapper configured to map encoded bits to symbols in a symbol constellation, a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper. In addition, the receiver includes a demodulator configured to demodulate the received signal via the communication channel, a demapper configured to estimate likelihoods from the demodulated signal, a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper. Furthermore, the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

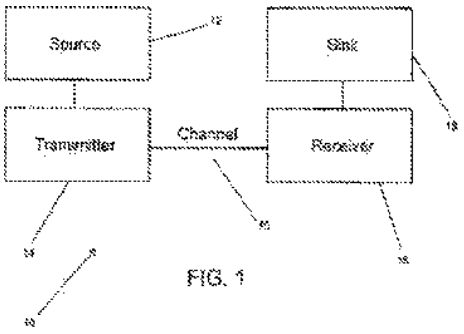


FIG. 1

EP 3 518 485 A1

Printed by Jouve, 75001 PARIS (FR)

Description

STATEMENT OF FEDERALLY SPONSORED RESEARCH

[0001] This invention was made with Government support under contract NAS7-03001 awarded by NASA. The Government has certain rights in this invention.

BACKGROUND

[0002] The present invention generally relates to bandwidth and/or power efficient digital transmission systems and more specifically to the use of unequally spaced constellations having increased capacity.

[0003] The term "constellation" is used to describe the possible symbols that can be transmitted by a typical digital communication system. A receiver attempts to detect the symbols that were transmitted by mapping a received signal to the constellation. The minimum distance (d_{min}) between constellation points is indicative of the capacity of a constellation at high signal-to-noise ratios (SNRs). Therefore, constellations used in many communication systems are designed to maximize d_{min}. Increasing the dimensionality of a constellation allows larger minimum distance for constant constellation energy per dimension. Therefore, a number of multi-dimensional constellations with good minimum distance properties have been designed.

[0004] Communication systems have a theoretical maximum capacity, which is known as the Shannon limit. Many communication systems attempt to use codes to increase the capacity of a communication channel. Significant coding gains have been achieved using coding techniques such as turbo codes and LDPC codes. The coding gains achievable using any coding technique are limited by the constellation of the communication system. The Shannon limit can be thought of as being based upon a theoretical constellation known as a Gaussian distribution, which is an infinite constellation where symbols at the center of the constellation are transmitted more frequently than symbols at the edge of the constellation. Practical constellations are finite and transmit symbols with equal likelihoods, and therefore have capacities that are less than the Gaussian capacity. The capacity of a constellation is thought to represent a limit on the gains that can be achieved using coding when using that constellation.

[0005] Prior attempts have been made to develop unequally spaced constellations. For example, a system has been proposed that uses unequally spaced constellations that are optimized to minimize the error rate of an uncoded system. Another proposed system uses a constellation with equiprobable but unequally spaced symbols in an attempts to mimic a Gaussian distribution.

[0006] Other approaches increases the dimensionality of a constellation or select a new symbol to be transmitted taking into consideration previously transmitted symbols. However, these constellation were still designed based on a minimum distance criteria.

SUMMARY OF THE INVENTION

[0007] Systems and methods are described for constructing a modulation such that the constrained capacity between a transmitter and a receiver approaches the Gaussian channel capacity limit first described by Shannon [ref Shannon 1948]. Traditional communications systems employ modulations that leave a significant gap to Shannon Gaussian capacity. The modulations of the present invention reduce, and in some cases, nearly eliminate this gap. The invention does not require specially designed coding mechanisms that tend to transmit some points of a modulation more frequently than others but rather provides a method for locating points (in a one or multiple dimensional space) in order to maximize capacity between the input and output of a bit or symbol mapper and demapper respectively. Practical application of the method allows systems to transmit data at a given rate for less power or to transmit data at a higher rate for the same amount of power.

[0008] One embodiment of the invention includes a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter includes a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate, a mapper configured to map encoded bits to symbols in a symbol constellation, a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper. In addition, the receiver includes a demodulator configured to demodulate the received signal via the communication channel, a demapper configured to estimate likelihoods from the demodulated signal, a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper. Furthermore, the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

[0009] A further embodiment of the invention includes encoding the bits of user information using a coding scheme, mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}, modulating the symbols in accordance with a modulation scheme, transmitting the modulated signal via the communication channel, receiving a modulated signal, demodulating the modulated signal in accordance with the modulation scheme, demapping the demodulated signal using the geometrically shaped signal constellation to produce likelihoods, and decoding the likelihoods to obtain an estimate of the decoded bits.

[0010] Another embodiment of the invention includes

selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0011] A still further embodiment of the invention includes selecting an appropriate constellation size and a desired capacity per dimension, estimating an initial SNR at which the system is likely to operate, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

[0012] Still another embodiment of the invention includes selecting an appropriate constellation size and a desired SNR, and optimizing the location of the points of the constellation to maximize a capacity measure of the constellation.

[0013] A yet further embodiment of the invention includes obtaining a geometrically shaped PAM constellation with a constellation size that is the square root of said given constellation size, where the geometrically shaped PAM constellation has a capacity greater than that of a PAM constellation that maximizes d_{min}, creating an orthogonalized PAM constellation using the geometrically shaped PAM constellation, and combining the geometrically shaped PAM constellation and the orthogonalized PAM constellation to produce a geometrically shaped QAM constellation.

[0014] Another further embodiment of the invention includes transmitting information over a channel using a geometrically shaped symbol constellation, and modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate.

BRIEF DESCRIPTION OF DRAWINGS

[0015]

FIG. 1 is a conceptual illustration of a communication system in accordance with an embodiment of the invention.

FIG. 2 is a conceptual illustration of a transmitter in accordance with an embodiment of the invention.

FIG. 3 is a conceptual illustration of a receiver in accordance with an embodiment of the invention.

FIG. 4a is a conceptual illustration of the joint capacity of a channel.

FIG. 4b is a conceptual illustration of the parallel decoding capacity of a channel.

FIG. 5 is a flow chart showing a process for obtaining a constellation optimized for capacity for use in a communication system having a fixed code rate and modulation scheme in accordance with an embodiment of the invention.

FIG. 6a is a chart showing a comparison of Gaussian capacity and PD capacity for traditional PAM-2,4,8,16,32.

FIG. 6b is a chart showing a comparison between Gaussian capacity and joint capacity for traditional PAM-2,4,8,16,32.

FIG. 7 is a chart showing the SNR gap to Gaussian capacity for the PD capacity and joint capacity of traditional PAM-2,4,8,16,32 constellations.

FIG. 8a is a chart comparing the SNR gap to Gaussian capacity of the PD capacity for tradtional and optimized PAM-2,4,8,16,32 constellations.

FIG. 8b is a chart comparing the SNR gap to Gaussian capacity of the joint capacity for tradtional and optimized PAM-2,4,8,16,32 constellations

FIG. 9 is a chart showing Frame Error Rate performance of traditional and PD capacity optimized PAM-32 constellations in simulations involving several different length LDPC codes.

FIGS. 10a - 10d are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 11a and 11b are design tables of PD capacity and joint capacity optimized PAM-4 constellations in accordance with embodiments of the invention.

FIGS. 12a - 12d are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 13a and 13b are design tables of PD capacity and joint capacity optimized PAM-8 constellations in accordance with embodiments of the invention.

FIGS. 14a - 14d are locus plots showing the location of constellation points of a PAM-16 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 15a and 15b are design tables of PD capacity and joint capacity optimized PAM-16 constellations in accordance with embodiments of the invention.

FIGS. 16a - 16d are locus plots showing the location of constellation points of a PAM-32 constellation optimized for PD capacity and joint capacity versus user bit rate per dimension and versus SNR.

FIGS. 17a and 17b are design tables of PD capacity and joint capacity optimized PAM-32 constellations in accordance with embodiments of the invention.

FIG. 18 is a chart showing the SNR gap to Gaussian capacity for traditional and capacity optimized PSK constellations.

FIG. 19 is a chart showing the location of constellation points of PD capacity optimized PSK-32 constellations.

FIG. 20 is a series of PSK-32 constellations optimized for PD capacity at different SNRs in accordance with embodiments of the invention.

FIG. 21 illustrates a QAM-64 constructed from or-

thogonal Cartesian product of two PD optimized PAM-8 constellations in accordance with an embodiment of the invention.

FIGS. 22a and 22b are locus plots showing the location of constellation points of a PAM-4 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

FIGS. 23a and 23b are locus plots showing the location of constellation points of a PAM-8 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

FIGS. 24a and 24b are locus plots showing the location of constellation points of a PAM-16 constellation optimized for PD capacity over a fading channel versus user bit rate per dimension and versus SNR.

DETAILED DESCRIPTION OF THE INVENTION

[0016] Turning now to the drawings, communication systems in accordance with embodiments of the invention are described that use signal constellations, which have unequally spaced (i.e. 'geometrically' shaped) points. In several embodiments, the locations of geometrically shaped points are designed to provide a given capacity measure at a reduced signal-to-noise ratio (SNR) compared to the SNR required by a constellation that maximizes d_{min^2}. In many embodiments, the constellations are selected to provide increased capacity at a predetermined range of channel signal-to-noise ratios (SNR). Capacity measures that can be used in the selection of the location of constellation points include, but are not limited to, parallel decode (PD) capacity and joint capacity.

[0017] In many embodiments, the communication systems utilize capacity approaching codes including, but not limited to, LDPC and Turbo codes. As is discussed further below, direct optimization of the constellation points of a communication system utilizing a capacity approaching channel code, can yield different constellations depending on the SNR for which they are optimized. Therefore, the same constellation is unlikely to achieve the same coding gains applied across all code rates; that is, the same constellation will not enable the best possible performance across all rates. In many instances, a constellation at one code rate can achieve gains that cannot be achieved at another code rate. Processes for selecting capacity optimized constellations to achieve increased coding gains based upon a specific coding rate in accordance with embodiments of the invention are described below. In a number of embodiments, the communication systems can adapt location of points in a constellation in response to channel conditions, changes in code rate and/or to change the target user data rate.

COMMUNICATION SYSTEMS

[0018] A communication system in accordance with an embodiment of the invention is shown in FIG. 1. The communication system 10 includes a source 12 that provides user bits to a transmitter 14. The transmitter transmits symbols over a channel to a receiver 16 using a predetermined modulation scheme. The receiver uses knowledge of the modulation scheme, to decode the signal received from the transmitter. The decoded bits are provided to a sink device that is connected to the receiver.

[0019] A transmitter in accordance with an embodiment of the invention is shown in FIG. 2. The transmitter 14 includes a coder 20 that receives user bits from a source and encodes the bits in accordance with a predetermined coding scheme. In a number of embodiments, a capacity approaching code such as a turbo code or a LDPC code is used. In other embodiments, other coding schemes can be used to providing a coding gain within the communication system. A mapper 22 is connected to the coder. The mapper maps the bits output by the coder to a symbol within a geometrically distributed signal constellation stored within the mapper. The mapper provides the symbols to a modulator 24, which modulates the symbols for transmission via the channel.

[0020] A receiver in accordance with an embodiment of the invention is illustrated in FIG. 3. The receiver 16 includes a demodulator 30 that demodulates a signal received via the channel to obtain symbol or bit likelihoods. The demapper uses knowledge of the geometrically shaped symbol constellation used by the transmitter to determine these likelihoods. The demapper 32 provides the likelihoods to a decoder 34 that decodes the encoded bit stream to provide a sequence of received bits to a sink.

GEOMETRICALLY SHAPED CONSTELLATIONS

[0021] Transmitters and receivers in accordance with embodiments of the invention utilize geometrically shaped symbol constellations. In several embodiments, a geometrically shaped symbol constellation is used that optimizes the capacity of the constellation. Various geometrically shaped symbol constellations that can be used in accordance with embodiments of the invention, techniques for deriving geometrically shaped symbol constellations are described below.

SELECTION OF A GEOMETRICALLY SHAPED CONSTELLATIONS

[0022] Selection of a geometrically shaped constellation for use in a communication system in accordance with an embodiment of the invention can depend upon a variety of factors including whether the code rate is fixed. In many embodiments, a geometrically shaped constellation is used to replace a conventional constellation (i.e. a constellation maximized for d_{min^2}) in the mapper of transmitters and the demapper of receivers within a communication system. Upgrading a communication system involves selection of a constellation and in many instances the upgrade can be achieved via a simple firmware up-

grade. In other embodiments, a geometrically shaped constellation is selected in conjunction with a code rate to meet specific performance requirements, which can for example include such factors as a specified bit rate, a maximum transmit power. Processes for selecting a geometric constellation when upgrading existing communication systems and when designing new communication systems are discussed further below.

UPGRADING EXISTING COMMUNICATION SYSTEMS

[0023] A geometrically shaped constellation that provides a capacity, which is greater than the capacity of a constellation maximized for d_{min}, can be used in place of a conventional constellation in a communication system in accordance with embodiments of the invention. In many instances, the substitution of the geometrically shaped constellation can be achieved by a firmware or software upgrade of the transmitters and receivers within the communication system. Not all geometrically shaped constellations have greater capacity than that of a constellation maximized for d_{min}. One approach to selecting a geometrically shaped constellation having a greater capacity than that of a constellation maximized for d_{min} is to optimize the shape of the constellation with respect to a measure of the capacity of the constellation for a given SNR. Capacity measures that can be used in the optimization process can include, but are not limited to, joint capacity or parallel decoding capacity.

JOINT CAPACITY AND PARALLEL DECODING CAPACITY

[0024] A constellation can be parameterized by the total number of constellation points, M, and the number of real dimensions, N_{dim}. In systems where there are no belief propagation iterations between the decoder and the constellation demapper, the constellation demapper can be thought of as part of the channel. A diagram conceptually illustrating the portions of a communication system that can be considered part of the channel for the purpose of determining PD capacity is shown in FIG. 4a. The portions of the communication system that are considered part of the channel are indicated by the ghost line 40. The capacity of the channel defined as such is the parallel decoding (PD) capacity, given by:

$$C_{PD} = \sum_{i=0}^{i=1} I(X_i; Y)$$

where X_i is the ith bit of the n-bits transmitted symbol, and Y is the received symbol, and $I(A;B)$ denotes the mutual information between random variables A and B.

[0025] Expressed another way, the PD capacity of a channel can be viewed in terms of the mutual information between the output bits of the encoder (such as an LDPC encoder) at the transmitter and the likelihoods computed by the demapper at the receiver. The PD capacity is influenced by both the placement of points within the constellation and by the labeling assignments.

[0026] With belief propagation iterations between the demapper and the decoder, the demapper can no longer be viewed as part of the channel, and the joint capacity of the constellation becomes the tightest known bound on the system performance. A diagram conceptually illustrating the portions of a communication system that are considered part of the channel for the purpose of determining the joint capacity of a constellation is shown in FIG. 4b. The portions of the communication system that are considered part of the channel are indicated by the ghost line 42. The joint capacity of the channel is given by:

$$C_{JOINT} = I(X;Y)$$

[0027] Joint capacity is a description of the achievable capacity between the input of the mapper on the transmit side of the link and the output of the channel (including, for example AWGN and Fading channels). Practical systems must often 'demap' channel observations prior to decoding. In general, the step causes some loss of capacity. In fact it can be proven that $C_C \geq C_{JOINT} \geq C_{PD}$. That is, C_{JOINT} upper bounds the capacity achievable by C_{PD}. The methods of the present invention are motivated by considering the fact that practical limits to a given communication system capacity are limited by C_{JOINT} and C_{PD}. In several embodiments of the invention, geometrically shaped constellations are selected that maximize these measures.

SELECTING A CONSTELLATION HAVING AN OPTIMAL CAPACITY

[0028] Geometrically shaped constellations in accordance with embodiments of the invention can be designed to optimize capacity measures including, but not limited to PD capacity or joint capacity. A process for selecting the points, and potentially the labeling, of a geometrically shaped constellation for use in a communication system having a fixed code rate in accordance with an embodiment of the invention is shown in FIG. 5. The process 50 commences with the selection (52) of an appropriate constellation size M and a desired capacity per dimension η. In the illustrated embodiment, the process involves a check (52) to ensure that the constellation size can support the desired capacity. In the event that the constellation size could support the desired capacity, then the process iteratively optimizes the M-ary constellation for the specified capacity. Optimizing a constellation for a specified capacity often involves an iterative process, because the optimal constellation depends upon the SNR

at which the communication system operates. The SNR for the optimal constellation to give a required capacity is not known *a priori*. Throughout the description of the present invention SNR is defined as the ratio of the average constellation energy per dimension to the average noise energy per dimension. In most cases the capacity can be set to equal the target user bit rate per symbol per dimension. In some cases adding some implementation margin on top of the target user bit rate could result in a practical system that can provide the required user rate at a lower rate. The margin is code dependent. The following procedure could be used to determine the target capacity that includes some margin on top of the user rate. First, the code (e.g. LDPC or Turbo) can be simulated in conjunction with a conventional equally spaced constellation. Second, from the simulation results the actual SNR of operation at the required error rate can be found. Third, the capacity of the conventional constellation at that SNR can be computed. Finally, a geometrically shaped constellation can be optimized for that capacity.

[0029] In the illustrated embodiment, the iterative optimization loop involves selecting an initial estimate of the SNR at which the system is likely to operate (i.e. SNR_{in}). In several embodiments the initial estimate is the SNR required using a conventional constellation. In other embodiments, other techniques can be used for selecting the initial SNR. An *M*-ary constellation is then obtained by optimizing (56) the constellation to maximize a selected capacity measure at the initial SNR_{in} estimate. Various techniques for obtaining an optimized constellation for a given SNR estimate are discussed below.

[0030] The SNR at which the optimized *M*-ary constellation provides the desired capacity per dimension γ (SNR_{out}) is determined (57). A determination (58) is made as to whether the SNR_{out} and SNR_{in} have converged. In the illustrated embodiment convergence is indicated by SNR_{out} equaling SNR_{in}. In a number of embodiments, convergence can be determined based upon the difference between SNR_{out} and SNR_{in} being less than a predetermined threshold. When SNR_{out} and SNR_{in} have not converged, the process performs another iteration selecting SNR_{out} as the new SNR_{in} (55). When SNR_{out} and SNR_{in} have converged, the capacity measure of the constellation has been optimized. As is explained in more detail below, capacity optimized constellation at low SNRs are geometrically shaped constellations that can achieve significantly higher performance gains (measured as reduction in minimum required SNR) than constellations that maximize d_{min}.

[0031] The process illustrated in FIG. 6 can maximize PD capacity or joint capacity of an *M*-ary constellation for a given SNR. Although the process illustrated in FIG. 5 shows selecting an *M*-ary constellation optimized for capacity, a similar process could be used that terminates upon generation of an *M*-ary constellation where the SNR gap to Gaussian capacity at a given capacity is a predetermined margin lower than the SNR gap of a conventional constellation, for example 0.5 db. Alternatively, other processes that identify *M*-ary constellations having capacity greater than the capacity of a conventional constellation can be used in accordance with embodiments of the invention. A geometrically shaped constellation in accordance with embodiments of the invention can achieve greater capacity than the capacity of a constellation that maximizes d_{min}, without having the optimal capacity for the SNR range within which the communication system operates.

[0032] We note that constellations designed to maximize joint capacity may also be particularly well suited to codes with symbols over GF(q), or with multi-stage decoding.. Conversely constellations optimized for PD capacity could be better suited to the more common case of codes with symbols over GF(2)

OPTIMIZING THE CAPACITY OF AN M-ARY CONSTELLATION AT A GIVEN SNR

[0033] Processes for obtaining a capacity optimized constellation often involve determining the optimum location for the points of an *M*-ary constellation at a given SNR. An optimization process, such as the optimization process 56 shown in FIG. 5, typically involves unconstrained or constrained non-linear optimization. Possible objective functions to be maximized are the Joint or PD capacity functions. These functions may be targeted to channels including but not limited to Additive White Gaussian Noise (AWGN) or Rayleigh fading channels. The optimization process gives the location of each constellation point identified by its symbol labeling. In the case where the objective is joint capacity, point bit labelings are irrelevant meaning that changing the bit labelings doesn't change the joint capacity as long as the set of point locations remains unchanged..

[0034] The optimization process typically finds the constellation that gives the largest PD capacity or joint capacity at a given SNR. The optimization process itself often involves an iterative numerical process that among other things considers several constellations and selects the constellation that gives the highest capacity at a given SNR. In other embodiments, the constellation that requires the least SNR to give a required PD capacity or joint capacity can also be found. This requires running the optimization process iteratively as shown in FIG. 5.

[0035] Optimization constraints on the constellation point locations may include, but are not limited to, lower and upper bounds on point location, peak to average power of the resulting constellation, and zero mean in the resulting constellation. It can be easily shown that a globally optimal constellation will have zero mean (no DC component). Explicit inclusion of a zero mean constraint helps the optimization routine to converge more rapidly. Except for cases where exhaustive search of all combinations of point locations and labelings is possible it will not necessarily always be the case that solutions are provably globally optimal. In cases where exhaustive

search is possible, the solution provided by the non-linear optimizer is in fact globally optimal.

[0036] The processes described above provide examples of the manner in which a geometrically shaped constellation having an increased capacity relative to a conventional capacity can be obtained for use in a communication system having a fixed code rate and modulation scheme. The actual gains achievable using constellations that are optimized for capacity compared to conventional constellations that maximize d_{min} are considered below.

GAINS ACHIEVED BY OPTIMIZED GEOMETRICALLY SPACED CONSTELLATIONS

[0037] The ultimate theoretical capacity achievable by any communication method is thought to be the Gaussian capacity, C_G which is defined as:

$$C_G = \frac{1}{2}log_2(1+SNR)$$

[0038] Where signal-to-noise (SNR) is the ratio of expected signal power to expected noise power. The gap that remains between the capacity of a constellation and C_G can be considered a measure of the quality of a given constellation design.

[0039] The gap in capacity between a conventional modulation scheme in combination with a theoretically optimal coder can be observed with reference to FIGS. 6a and 6b. FIG. 6a includes a chart 60 showing a comparison between Gaussian capacity and the PD capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}. Gaps 62 exist between the plot of Gaussian capacity and the PD capacity of the various PAM constellations. FIG. 6b includes a chart 64 showing a comparison between Gaussian capacity and the joint capacity of conventional PAM-2, 4, 8, 16, and 32 constellations that maximize d_{min}. Gaps 66 exist between the plot of Gaussian capacity and the joint capacity of the various PAM constellations. These gaps in capacity represent the extent to which conventional PAM constellations fall short of obtaining the ultimate theoretical capacity i.e. the Gaussian capacity..

[0040] In order to gain a better view of the differences between the curves shown in FIGS. 6a and 6b at points close to the Gaussian capacity, the SNR gap to Gaussian capacity for different values of capacity for each constellation are plotted in FIG. 7. It is interesting to note from the chart 70 in FIG. 7 that (unlike the joint capacity) at the same SNR, the PD capacity does not necessarily increase with the number of constellation points. As is discussed further below, this is not the case with PAM constellations optimized for PD capacity.

[0041] FIGS. 8a and 8b summarize performance of constellations for PAM-4, 8, 16, and 32 optimized for PD capacity and joint capacity (it should be noted that BPSK is the optimal PAM-2 constellation at all code rates). The constellations are optimized for PD capacity and joint capacity for different target user bits per dimension (i.e. code rates). The optimized constellations are different depending on the target user bits per dimension, and also depending on whether they have been designed to maximize the PD capacity or the joint capacity. All the PD optimized PAM constellations are labeled using a gray labeling which is not always the binary reflective gray labeling. It should be noted that not all gray labels achieve the maximum possible PD capacity even given the freedom to place the constellation points anywhere on the real line. FIG. 8a shows the SNR gap for each constellation optimized for PD capacity. FIG. 8b shows the SNR gap to Gaussian capacity for each constellation optimized for joint capacity. Again, it should be emphasized that each '+' on the plot represents a different constellation.

[0042] Referring to FIG. 8a, the coding gain achieved using a constellation optimized for PD capacity can be appreciated by comparing the SNR gap at a user bit rate per dimension of 2.5 bits for PAM-32. A user bit rate per dimension of 2.5 bits for a system transmitting 5 bits per symbol constitutes a code rate of 1/2. At that code rate, the constellation optimized for PD capacity provides an additional coding gain of approximately 1.5 dB when compared to the conventional PAM-32 constellation.

[0043] The SNR gains that can be achieved using constellations that are optimized for PD capacity can be verified through simulation. The results of a simulation conducted using a rate 1/2 LDPC code in conjunction with a conventional PAM-32 constellation and in conjunction with a PAM-32 constellation optimized for PD capacity are illustrated in FIG. 9. A chart 90 includes plots of Frame Error Rate performance of the different constellations with respect to SNR and using different length codes (i.e. k=4,096 and k=16,384). Irrespective of the code that is used, the constellation optimized for PD capacity achieves a gain of approximately 1.5 dB, which closely approaches the gain predicted from FIG. 8a.

CAPACITY OPTIMIZED PAM CONSTELLATIONS

[0044] Using the processes outlined above, locus plots of PAM constellations optimized for capacity can be generated that show the location of points within PAM constellations versus SNR. Locus plots of PAM-4, 8, 16, and 32 constellations optimized for PD capacity and joint capacity and corresponding design tables at various typical user bit rates per dimension are illustrated in FIGS. 10a - 17b. The locus plots and design tables show PAM-4,8,16,32 constellation point locations and labelings from low to high SNR corresponding to a range of low to high spectral efficiency.

[0045] In FIG. 10a, a locus plot 100 shows the location of the points of PAM-4 constellations optimized for joint capacity plotted against achieved capacity. A similar locus plot 105 showing the location of the points of Joint

capacity optimized PAM-4 constellations plotted against SNR is included in FIG. 10b. In FIG. 10c, the location of points for PAM-4 optimized for PD capacity is plotted against achievable capacity and in FIG. 10d the location of points for PAM-4 for PD capacity is plotted against SNR. At low SNRs, the PD capacity optimized PAM-4 constellations have only 2 unique points, while the Joint optimized constellations have 3. As SNR is increased, each optimization eventually provides 4 unique points. This phenomenon is explicitly described in FIG. 11a and FIG. 11b where vertical slices of FIGS. 10ab and 10cd are captured in tables describing some PAM-4 constellations designs of interest. The SNR slices selected represent designs that achieve capacities = {0.5, 0.75, 1.0, 1.25, 1.5} bits per symbol (bps). Given that PAM-4 can provide at most $\log_2(4) = 2$ bps, these design points represent systems with information code rates R = {1/4, 3/8, 1/2, 5/8, 3/4} respectively.

[0046] FIGS. 12ab and 12cd present locus plots of PD capacity and joint capacity optimized PAM-8 constellation points versus achievable capacity and SNR. FIGS. 13a and 13b provide slices from these plots at SNRs corresponding to achievable capacities η = {0.5, 1.0, 1.5, 2.0, 2.5} bps. Each of these slices correspond to systems with code rate $R = \eta$ bps / $\log_2(8)$, resulting in R ={ 1/6, 1/3, 1/2, 2/3, 5/6}. As an example of the relative performance of the constellations in these tables, consider FIG. 13b which shows a PD capacity optimized PAM-8 constellation optimized for SNR = **9.00 dB**, or 1.5 bps. We next examine the plot provided in FIG. **6a** and see that the gap of the optimized constellation to the ultimate, Gaussian, capacity (CG) is approximately 0.5 dB. At the same spectral efficiency, the gap of the traditional PAM-8 constellation is approximately 1.0 dB. The advantage of the optimized constellation is 0.5dB for the same rate (in this case R=1/2). This gain can be obtained by only changing the mapper and demapper in the communication system and leaving all other blocks the same.

[0047] Similar information is presented in FIGS. 14abcd, and 15ab which provide loci plots and *design* tables for PAM-16 PD capacity and joint capacity optimized constellations. Likewise FIGS. 16abcd, 17ab provide loci plots and design tables for PAM-32 PD capacity and joint capacity optimized constellations.

CAPACITY OPTIMIZED PSK CONSTELLATIONS

[0048] Traditional phase shift keyed (PSK) constellations are already quite optimal. This can be seen in the chart 180 comparing the SNR gaps of tradition PSK with capacity optimized PSK constellations shown in FIG. 18 where the gap between PD capacity and Gaussian capacity is plotted for traditional PSK-4,8,16,32 and for PD capacity optimized PSK-4,8,16,32.

[0049] The locus plot of PD optimized PSK-32 points across SNR is shown in FIG. 19, which actually characterizes all PSKs with spectral efficiency $\eta \leq 5$. This can be seen in FIG. 20. Note that at low SNR (0.4 dB) the optimal PSK-32 design is the same as traditional PSK-4, at SNR = 8.4 dB optimal PSK-32 is the same as traditional PSK-8, at SNR = 14.8 dB optimal PSK-32 is the same as traditional PSK-16, and finally at SNRs greater than 20.4 dB optimized PSK-32 is the same as traditional PSK-32. There are SNRs between these discrete points (for instance SNR = 2 and 15. dB) for which optimized PSK-32 provides superior PD capacity when compared to traditional PSK constellations.

[0050] We note now that the locus of points for PD optimized PSK-32 in FIG. 19 in conjunction with the gap to Gaussian capacity curve for optimized PSK-32 in FIG. 18 implies a potential design methodology. Specifically, the designer could achieve performance equivalent or better than that enabled by traditional PSK-4,8,16 by using only the optimized PSK-32 in conjunction with a single tuning parameter that controlled where the constellation points should be selected from on the locus of FIG. 19. Such an approach would couple a highly rate adaptive channel code that could vary its rate, for instance, rate 4/5 to achieve and overall (code plus optimized PSK-32 modulation) spectral efficiency of 4 bits per symbol, down to 1/5 to achieve an overall spectral efficiency of 1 bit per symbol. Such an adaptive modulation and coding system could essentially perform on the optimal continuum represented by the rightmost contour of FIG. 18.

ADAPTIVE RATE DESIGN

[0051] In the previous example spectrally adaptive use of PSK-32 was described. Techniques similar to this can be applied for other capacity optimized constellations across the link between a transmitter and receiver. For instance, in the case where a system implements quality of service it is possible to instruct a transmitter to increase or decrease spectral efficiency on demand. In the context of the current invention a capacity optimized constellation designed precisely for the target spectral efficiency can be loaded into the transmit mapper in conjunction with a code rate selection that meets the end user rate goal. When such a modulation / code rate change occurred a message could propagated to the receiver so that the receiver, in anticipation of the change, could select a demapper / decoder configuration in order to match the new transmit-side configuration.

[0052] Conversely, the receiver could implement a quality of performance based optimized constellation / code rate pair control mechanism. Such an approach would include some form of receiver quality measure. This could be the receiver's estimate of SNR or bit error rate. Take for example the case where bit error rate was above some acceptable threshold. In this case, via a backchannel, the receiver could request that the transmitter lower the spectral efficiency of the link by swapping to an alternate capacity optimized constellation / code rate pair in the coder and mapper modules and then signaling the receiver to swap in the complementary pairing in the demapper / decoder modules.

GEOMETRICALLY SHAPED QAM CONSTELLATIONS

[0053] Quadrature amplitude modulation (QAM) constellations can be constructed by orthogonalizing PAM constellations into QAM inphase and quadrature components. Constellations constructed in this way can be attractive in many applications because they have low-complexity demappers.

[0054] In FIG. 21 we provide an example of a Quadrature Amplitude Modulation constellation constructed from a Pulse Amplitude Modulation constellation. The illustrated embodiment was constructed using a PAM-8 constellation optimized for PD capacity at user bit rate per dimension of 1.5 bits (corresponds to an SNR of 9.0 dB) (see FIG. 13b). The label-point pairs in this PAM-8 constellation are {(000, -1.72), (001, -0.81), (010, 1.72), (011,-0.62), (100, 0.62), (101, 0.02), (110, 0.81), (111, -0.02)}. Examination of FIG. 21 shows that the QAM constellation construction is achieved by replicating a complete set of PAM-8 points in the quadrature dimension for each of the 8 PAM-8 points in the in-phase dimension. Labeling is achieved by assigning the PAM-8 labels to the LSB range on the in-phase dimension and to the MSB range on the quadrature dimension. The resulting 8x8 outer product forms a highly structured QAM-64 for which very low-complexity de-mappers can be constructed. Due to the orthogonality of the in-phase and quadrature components the capacity characteristics of the resulting QAM-64 constellation are identical to that of the PAM-8 constellation on a per-dimension basis.

N-DIMENSIONAL CONSTELLATION OPTIMIZATION

[0055] Rather than designing constellations in 1-D (PAM for instance) and then extending to 2-D (QAM), it is possible to take direct advantage in the optimization step of the additional degree of freedom presented by an extra spatial dimension. In general it is possible to design N-dimensional constellations and associated labellings. The complexity of the optimization step grows exponentially in the number of dimensions as does the complexity of the resulting receiver de-mapper. Such constructions constitute embodiments of the invention and simply require more 'run-time' to produce.

CAPACITY OPTIMIZED CONSTELALTIONS FOR FADING CHANNELS

[0056] Similar processes to those outlined above can be used to design capacity optimized constellations for fading channels in accordance with embodiments of the invention. The processes are essentially the same with the exception that the manner in which capacity is calculated is modified to account for the fading channel. A fading channel can be described using the following equation:

$$Y = a(t)X+N$$

where X is the transmitted signal, N is an additive white Gaussian noise signal and $a(t)$ is the fading distribution, which is a function of time.

[0057] In the case of a fading channel, the instantaneous SNR at the receiver changes according to a fading distribution. The fading distribution is Rayleigh and has the property that the average SNR of the system remains the same as in the case of the AWGN channel, $E[X^2]/E[N^2]$. Therefore, the capacity of the fading channel can be computed by taking the expectation of AWGN capacity, at a given average SNR, over the Rayleigh fading distribution of a that drives the distribution of the instantaneous SNR.

[0058] Many fading channels follow a Rayleigh distribution. FIGS. 22a - 24b are locus plots of PAM-4, 8, and 16 constellations that have been optimized for PD capacity on a Rayleigh fading channel. Locus plots versus user bit rate per dimension and versus SNR are provided. Similar processes can be used to obtain capacity optimized constellations that are optimized using other capacity measures, such as joint capacity, and/or using different modulation schemes.

[0059] The following clauses form part of the description.

Clause 1. A digital communication system, comprising:

a transmitter configured to transmit signals to a receiver via a communication channel, wherein the transmitter, comprises:

a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate;
a mapper configured to map encoded bits to symbols in a symbol constellation;
a modulator configured to generate a signal for transmission via the

communication channel using symbols generated by the mapper;
wherein the receiver, comprises:

a demodulator configured to demodulate the received signal via the communication channel;
a demapper configured to estimate likelihoods from the demodulated signal;
a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper; and

wherein the symbol constellation is a capacity

optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}.

Clause 2. The communication system of clause 1, wherein the geometrically spaced symbol constellation is capacity optimized subject to additional constraints.

Clause 3. The communication system of clause 1, wherein the geometrically shaped symbol constellation is optimized for capacity using parallel decode capacity.

Clause 4. The communication system of clause 1, wherein the geometrically shaped symbol constellation is optimized for capacity using joint capacity.

Clause 5. The communication system of clause 1, wherein the channel is an AWGN channel.

Clause 6. The communication system of clause 1, wherein the channel is a fading channel.

Clause 7. The communication system of clause 1, wherein the constellations points of the geometrically spaced symbol constellation are variable depending on the user rate.

Clause 8. A method of transmitting bits of user information over a communication channel, comprising:

encoding the bits of user information using a coding scheme;
mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min};
modulating the symbols in accordance with a modulation scheme;
transmitting the modulated signal via the communication channel;
receiving a modulated signal;
demodulating the modulated signal in accordance with the modulation scheme;
demapping the demodulated signal using the geometrically shaped signal constellation to produce likelihoods;
decoding the likelihoods to obtain an estimate of the decoded bits.

Clause 9. A method of selecting the points of a geometrically shaped symbol constellation used in a communication system, comprising:

selecting an appropriate constellation size and a desired capacity per dimension;
estimating an initial SNR at which the system is likely to operate; and
iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

Clause 10. A geometrically shaped symbol constellation used in a communication system, obtained by a process of:

selecting an appropriate constellation size and a desired capacity per dimension;
estimating an initial SNR at which the system is likely to operate; and
iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved.

Clause 11. A geometrically shaped symbol constellation used in a communication system, obtained by a process of:

selecting an appropriate constellation size and a desired SNR; and
optimizing the location of the points of the constellation to maximize a capacity measure of the constellation.

Clause 12. A method of transmitting information, comprising:

transmitting information over a channel using a geometrically shaped symbol constellation; and
modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate.

Claims

1. A communication system (10), comprising:

a transmitter (14) that transmits signals via a communication channel (15) having a channel signal-to-noise ratio, where the transmitter (14), comprises:

a coder (20) configured to receive user bits

and output encoded bits at a data rate;
a mapper (22) configured to map encoded bits to symbols in a symbol constellation; and
a modulator (24) configured to generate a signal for transmission via the communication channel using symbols generated by the mapper;

wherein the transmitter is **characterized in that** the symbol constellation comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

2. The communication system (10) of claim 1, further comprising:
a receiver (16) that receives signals via the communication channel (15), wherein the receiver comprises:

a demodulator (30) configured to demodulate a signal received via the communication channel;
a demapper (32) configured to estimate likelihoods from the demodulated signal using the symbol constellation; and
a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32).

3. A communication system (10), comprising:

a receiver (16) that receives signals via a communication channel (15) having a channel signal-to-noise ratio, wherein the receiver comprises:

a demodulator (30) configured to demodulate a signal received via the communication channel;
a demapper (32) configured to estimate likelihoods from the demodulated signal using a symbol constellation; and
a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32) based upon a code having a data rate;

wherein the receiver is **characterized in that** the symbol constellation comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

4. The communication system (10) of any preceding claim, wherein the labels and unequal spacing of the constellation points target the data rate at the channel signal-to-noise ratio.

5. The communication system (10) of any preceding claim, wherein the labels and spacing of the constellation points of the symbol constellation correspond to the labels and spacing of an optimized geometrically shaped symbol constellation obtained by optimizing the locations and the labeling of constellation points to increase parallel decode capacity at a given signal-to noise ratio in such a way that the obtained constellation points have a greater parallel decode capacity than the capacity of a uniform constellation that maximizes the minimum distance, d_{min}, between constellation points.

6. The communication system (10) of claim 5, wherein the locations and the labeling of the constellation points of the optimized geometrically shaped symbol constellation were optimized subject to optimization constraints including at least one of:

lower and upper bounds on point location;
peak to average power of the resulting constellation; and
the mean of the constellation points being zero.

7. The communication system (10) of any of claims 1 - 4, wherein the labels and spacing of the constellation points of the symbol constellation are obtainable by optimizing the locations and the labeling of constellation points to increase parallel decode capacity at a given signal-to noise ratio in such a way that the obtained geometrically shaped constellation points have a greater parallel decode capacity than the capacity of a uniform constellation that maximizes the minimum distance, d_{min}, between constellation points.

8. The communication system (10) of claim 7, wherein the labels and spacing of the constellation points of the symbol constellation are obtainable by optimizing the locations and the labeling of constellation points to increase parallel decode capacity at a given signal-to noise ratio subject to optimization constraints including at least one of:

lower and upper bounds on point location;
peak to average power of the resulting constellation; and
the mean of the constellation points being zero.

9. The communication system (10) of any preceding claim, wherein the symbol constellation is a multidimensional symbol constellation, where the labels and spacing of the constellation points of the multidimensional symbol constellation provide greater parallel decode capacity at the channel signal-to-noise ratio than a uniform symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring and that maximizes the minimum distance, d_{min}, between the constellation points of the uniform symbol constellation.

10. The communication system (10) of any of claims 1, 2, and 4-9, wherein a coder or a decoder is configured to use a Low Density Parity Check (LDPC) code.

11. The communication system (10) of any preceding claim, wherein the symbol constellation is selected in conjunction with the code rate to meet a maximum transmit power requirement.

12. The communication system (10) of any preceding claim wherein the channel is an Additive White Gaussian Noise (AWGN) channel.

13. The communication system (10) of any preceding claim wherein spacing of constellation points within the symbol constellation varies depending on the user rate.

14. The communication system (10) of any preceding claim, wherein each symbol in the symbol constellation is equiprobable to be transmitted and the symbol constellation is symmetrical.

15. The communication system (10) of any preceding claim, wherein the symbol constellation provides a given parallel decode capacity at a signal-to-noise ratio that is 0.5 dB lower compared to a symbol constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

16. The communication system of claim 1, wherein the symbol constellation comprises 16 constellation points and the channel signal-to-noise ratio is greater than 8.4 dB and less than 14.8 dB.

17. The communication system of claim 1, wherein the symbol constellation comprises 16 constellation points and the channel signal-to-noise ratio is 8.87dB.

18. The communication system of claim 1, wherein the ring of constellation points comprises a ring selected from:

a ring of 4 constellation points;
a ring of 8 constellation points;
a ring of 16 constellation points; and
a ring of 32 constellation points.

19. The communication system (10) of any preceding claim, wherein:

the communication system (10) is configured to operate at a first data rate using a first symbol constellation and at a second data rate using a second symbol constellation;
the first symbol constellation comprises a specific number of constellation points that are unequally spaced around a first ring; and
the second symbol constellation comprises the specific number of constellation points that are equally spaced around a second ring.

20. The communication system (10) of any of claims 1 - 18, wherein:

the communication system (10) is configured to operate at a first data rate using a first symbol constellation and at a second data rate using a second symbol constellation;
the first symbol constellation comprises a specific number of constellation points that are unequally spaced around a first ring;
the second symbol constellation comprises the specific number of constellation points that are unequally spaced around a second ring; and
the spacing of the specific number of constellation points around the first ring in the first symbol constellation differs from the spacing of the specific number of constellation points around the second ring in the second symbol constellation.

21. A signal transmitted via a communication channel (15) at a channel signal-to-noise ratio, comprising:

modulated symbols selected from a symbol constellation;
wherein the symbol constellation is **characterized in that** it comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points evenly equally spaced around the ring.

22. A method of transmitting information, comprising transmitting information over a communication channel (15) at a channel signal-to-noise ratio using a geometrically shaped symbol constellation that comprises a ring of constellation points that are unequally

spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

23. The method of claim 22 further comprising modifying the location of points within the geometrically shaped symbol constellation in response to a change in a target data rate.

24. The method of claim 22 further comprising:

changing a target data rate; and
selecting a new symbol constellation in response to a change in a target data rate, where the new symbol constellation comprises a ring of constellation points that are in different locations to the geometrically shaped symbol constellation.

Amended claims in accordance with Rule 137(2) EPC.

1. A communication system (10), comprising:

a transmitter (14) that transmits signals via a communication channel (15) having a channel signal-to-noise ratio, where the transmitter (14), comprises:

a coder (20) configured to receive user bits and output encoded bits according to a code rate;
a mapper (22) configured to map encoded bits to symbols in a symbol constellation; and
a modulator (24) configured to generate a signal for transmission via the communication channel using symbols generated by the mapper;

wherein the transmitter is **characterized in that** the symbol constellation comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

2. The communication system (10) of claim 1, further comprising:
a receiver (16) that receives signals via the communication channel (15), wherein the receiver comprises:

a demodulator (30) configured to demodulate a signal received via the communication channel;
a demapper (32) configured to estimate likelihoods from the demodulated signal using the symbol constellation; and
a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32).

3. A communication system (10), comprising:

a receiver (16) that receives signals via a communication channel (15) having a channel signal-to-noise ratio, wherein the receiver comprises:

a demodulator (30) configured to demodulate a signal received via the communication channel;
a demapper (32) configured to estimate likelihoods from the demodulated signal using a symbol constellation; and
a decoder (34) that is configured to estimate decoded bits from the likelihoods generated by the demapper (32) based upon a code having a data rate;

wherein the receiver is **characterized in that** the symbol constellation comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

4. The communication system (10) of any preceding claim, wherein the labels and unequal spacing of the constellation points target the data rate at the channel signal-to-noise ratio.

5. The communication system (10) of any preceding claim, wherein the labels and spacing of the constellation points of the symbol constellation correspond to the labels and spacing of an optimized geometrically shaped symbol constellation obtained by optimizing the locations and the labeling of constellation points to maximise parallel decode capacity at a given signal-to noise ratio in such a way that the obtained constellation points have a greater parallel decode capacity than the capacity of a uniform constellation that maximizes the minimum distance, d_{min}, between constellation points.

6. The communication system (10) of claim 5, wherein

the locations and the labeling of the constellation points of the optimized geometrically shaped symbol constellation were optimized subject to optimization constraints including at least one of:

lower and upper bounds on point location;
peak to average power of the resulting constellation; and
the mean of the constellation points being zero.

7. The communication system (10) of any of claims 1 - 4, wherein the labels and spacing of the constellation points of the symbol constellation are obtainable by optimizing the locations and the labeling of constellation points to maximise parallel decode capacity at a given signal-to noise ratio in such a way that the obtained geometrically shaped constellation points have a greater parallel decode capacity than the capacity of a uniform constellation that maximizes the minimum distance, d_{min}, between constellation points.

8. The communication system (10) of claim 7, wherein the labels and spacing of the constellation points of the symbol constellation are obtainable by optimizing the locations and the labeling of constellation points to increase parallel decode capacity at a given signal-to noise ratio subject to optimization constraints including at least one of:

lower and upper bounds on point location;
peak to average power of the resulting constellation; and
the mean of the constellation points being zero.

9. The communication system (10) of any preceding claim, wherein the symbol constellation is a multidimensional symbol constellation, where the labels and spacing of the constellation points of the multidimensional symbol constellation provide greater parallel decode capacity at the channel signal-to noise ratio than a uniform symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring and that maximizes the minimum distance, d_{min}, between the constellation points of the uniform symbol constellation.

10. The communication system (10) of any of claims 1, 2, and 4-9, wherein a coder or a decoder is configured to use a Low Density Parity Check code.

11. The communication system (10) of any preceding claim, wherein the symbol constellation is selected in conjunction with the code rate to meet a maximum transmit power requirement.

12. The communication system (10) of any preceding claim wherein the channel is an Additive White Gaussian Noise channel.

13. The communication system (10) of any preceding claim wherein spacing of constellation points within the symbol constellation varies depending on the user rate.

14. The communication system (10) of any preceding claim, wherein the symbol constellation has a zero mean.

15. The communication system (10) of any preceding claim, wherein the symbol constellation provides a given parallel decode capacity at a signal-to-noise ratio that is 0.5 dB lower compared to a symbol constellation that maximizes the minimum distance, d_{min}, between the constellation points of the symbol constellation.

16. The communication system of claim 1, wherein the symbol constellation comprises 16 constellation points and the channel signal-to-noise ratio is greater than 8.4 dB and less than 14.8 dB.

17. The communication system of claim 1, wherein the symbol constellation comprises 16 constellation points and the channel signal-to-noise ratio is 8.67dB.

18. The communication system of claim 1, wherein the ring of constellation points comprises a ring selected from:

a ring of 4 constellation points;
a ring of 8 constellation points;
a ring of 16 constellation points; and
a ring of 32 constellation points.

19. The communication system (10) of any preceding claim, wherein:

the communication system (10) is configured to operate at a first data rate using a first symbol constellation and at a second data rate using a second symbol constellation;
the first symbol constellation comprises a specific number of constellation points that are unequally spaced around a first ring; and
the second symbol constellation comprises the specific number of constellation points that are equally spaced around a second ring.

20. The communication system (10) of any of claims 1 - 19, wherein:

the communication system (10) is configured to operate at a first data rate using a first symbol constellation and at a second data rate using a

second symbol constellation;
the first symbol constellation comprises a specific number of constellation points that are unequally spaced around a first ring;
the second symbol constellation comprises the specific number of constellation points that are unequally spaced around a second ring; and
the spacing of the specific number of constellation points around the first ring in the first symbol constellation differs from the spacing of the specific number of constellation points around the second ring in the second symbol constellation.

21. A signal transmitted via a communication channel (15) at a channel signal-to-noise ratio, comprising:

modulated symbols selected from a symbol constellation;
wherein the symbol constellation is characterized in that it comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points evenly equally spaced around the ring.

22. A method of transmitting information, comprising transmitting information over a communication channel (15) at a channel signal-to-noise ratio using a geometrically shaped symbol constellation that comprises a ring of constellation points that are unequally spaced around the ring, where the labels and spacing of the constellation points provide greater parallel decode capacity at the channel signal-to-noise ratio than a symbol constellation comprising a ring of the same number of constellation points equally spaced around the ring.

23. The method of claim 22 further comprising modifying the location of points within the geometrically shaped symbol constellation in response to a change in a target data rate.

24. The method of claim 22 further comprising:

changing a target data rate; and
selecting a new symbol constellation in response to a change in a target data rate, where the new symbol constellation comprises a ring of constellation points that are in different locations to the geometrically shaped symbol constellation.

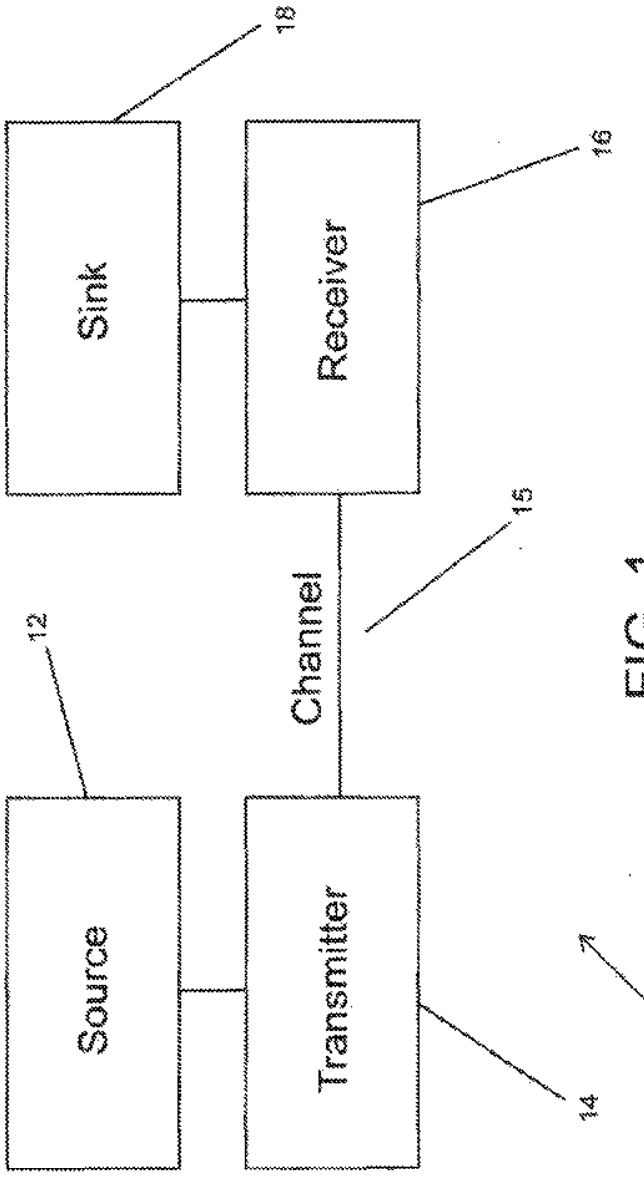


FIG. 1

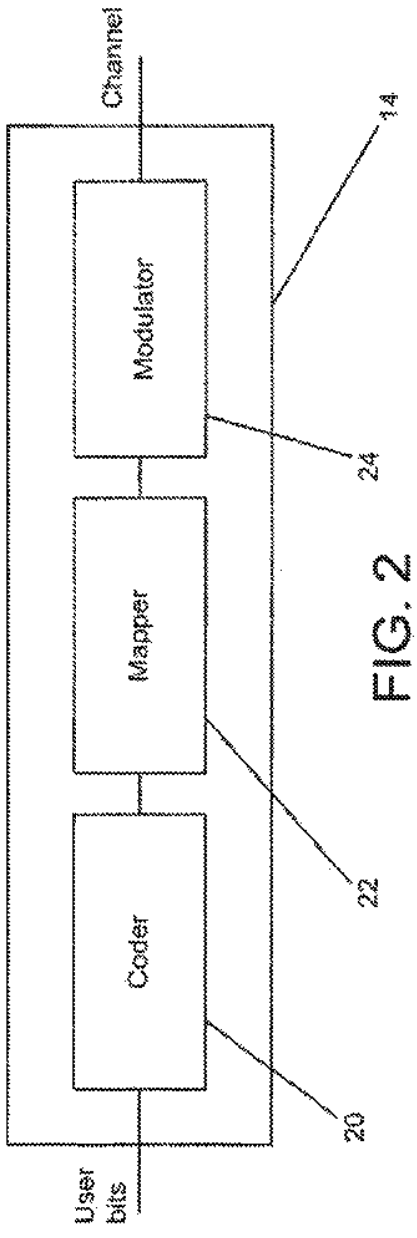

Channel
Modulator
Mapper
Coder
User bits
14
24
22
20

FIG. 2

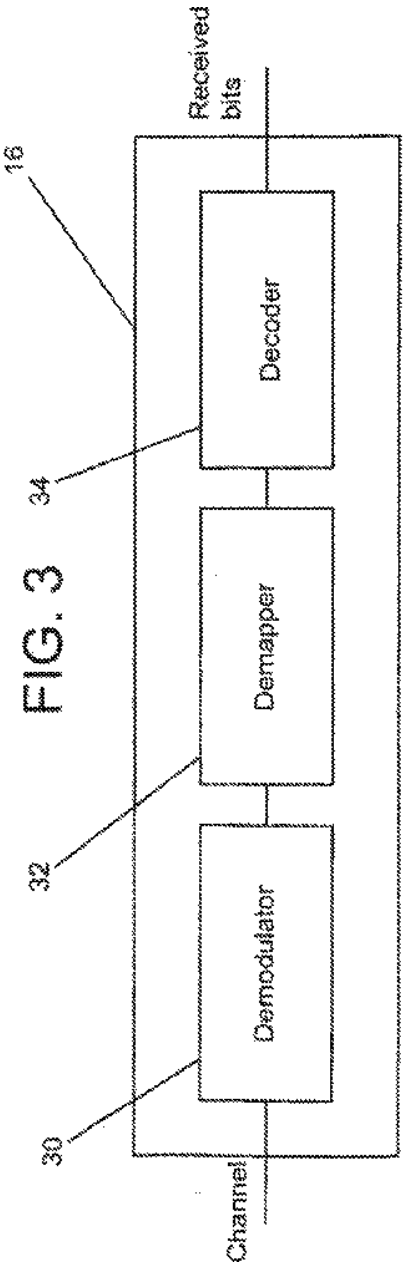

Received bits
Decoder
Demapper
Demodulator
Channel
16
34
32
30

FIG. 3

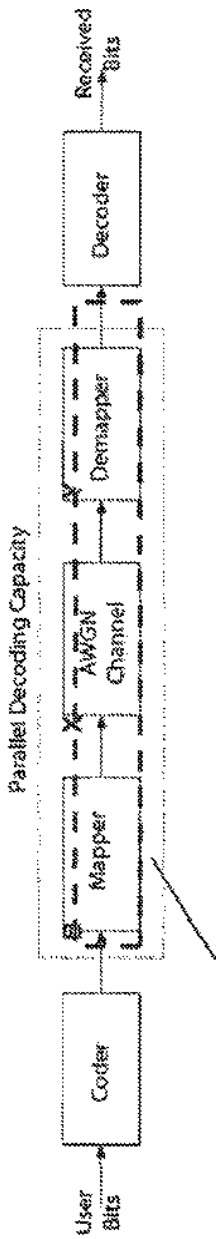

Parallel Decoding Capacity
User Bits
Coder
B
Mapper
X
AWGN Channel
Demapper
Decoder
Received Bits
40

FIG. 4a

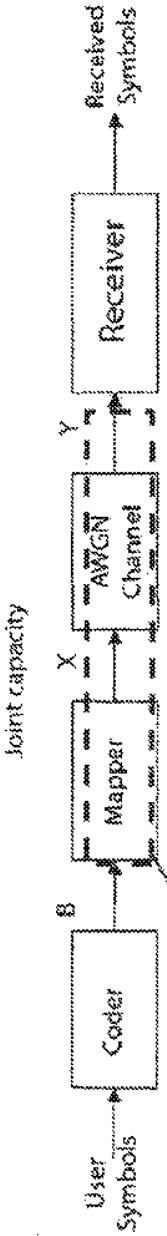

Joint capacity
User Symbols
Coder
B
Mapper
X
AWGN Channel
Y
Receiver
Received Symbols
42

FIG. 4b

50
Input M, η
51
log2(M) > η
52
No
Invalid Parameters
SNR_out = C^-1(η)
SNR_in = SNR_out
55
Optimize M-Ary Constellation for SNRin
56
SNR_out = C^-1(η)
57
SNR_in − SNR_out > ε
58
Yes
Output Constellation
59

FIG. 5

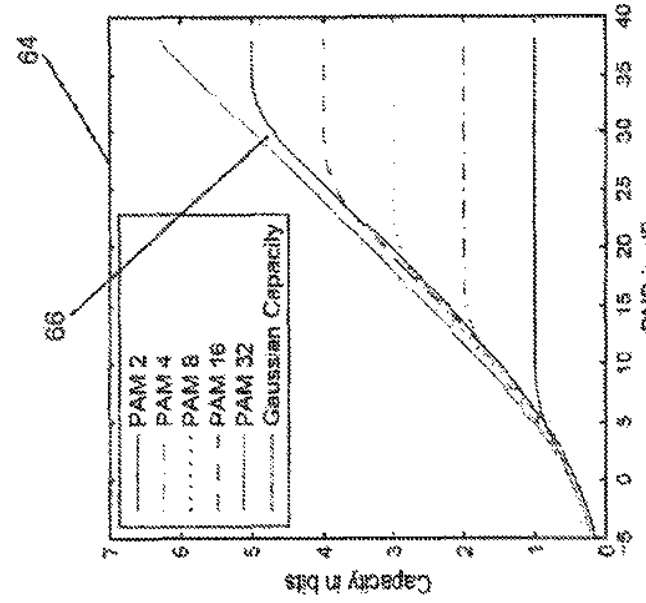

64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(b)

FIG. 6b

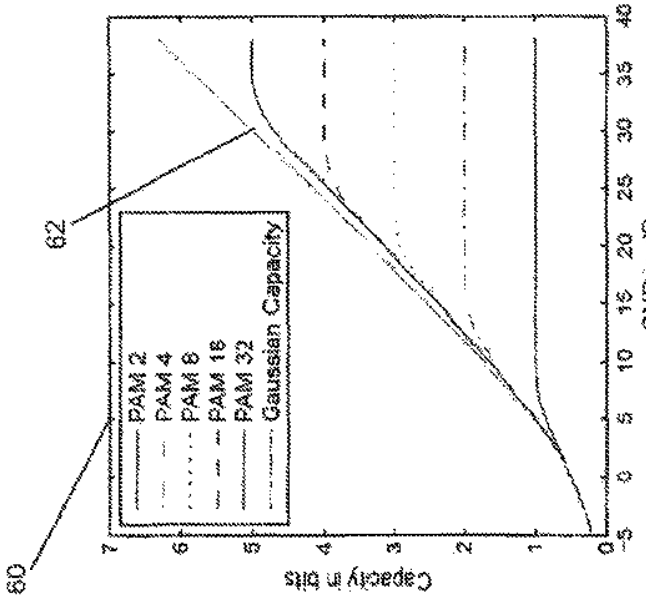

60
62
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(a)

FIG. 6a

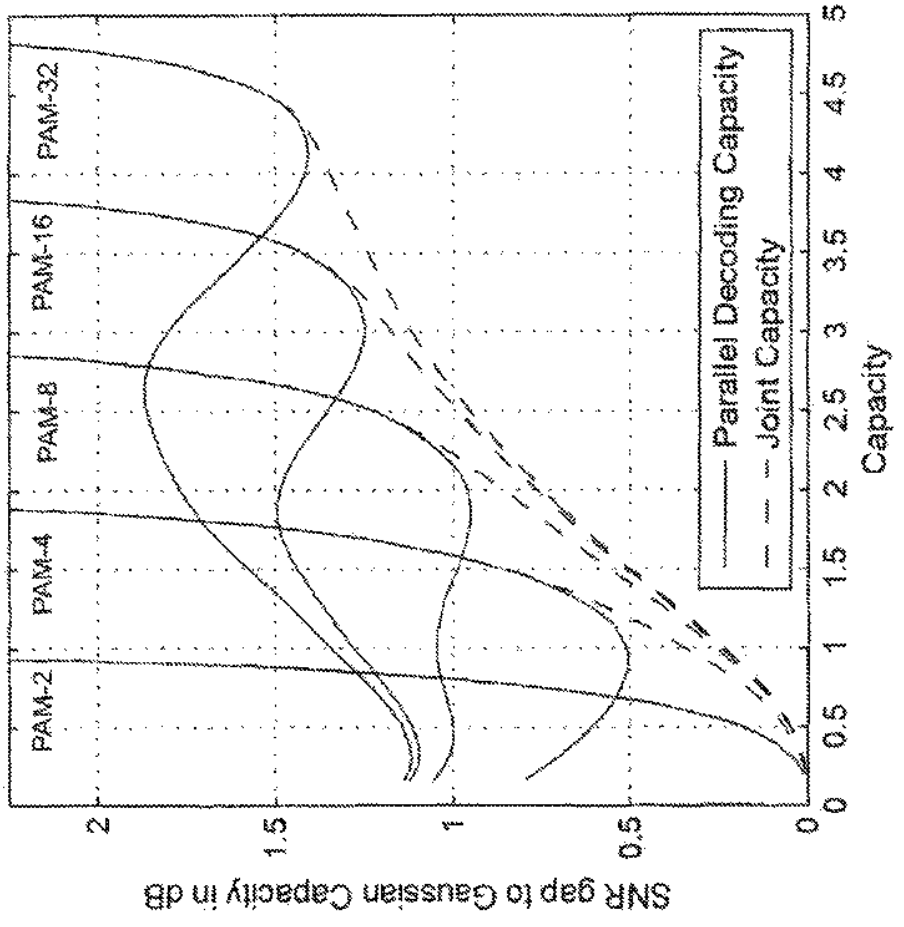

PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
Parallel Decoding Capacity
Joint Capacity
SNR gap to Gaussian Capacity in dB
Capacity
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 7

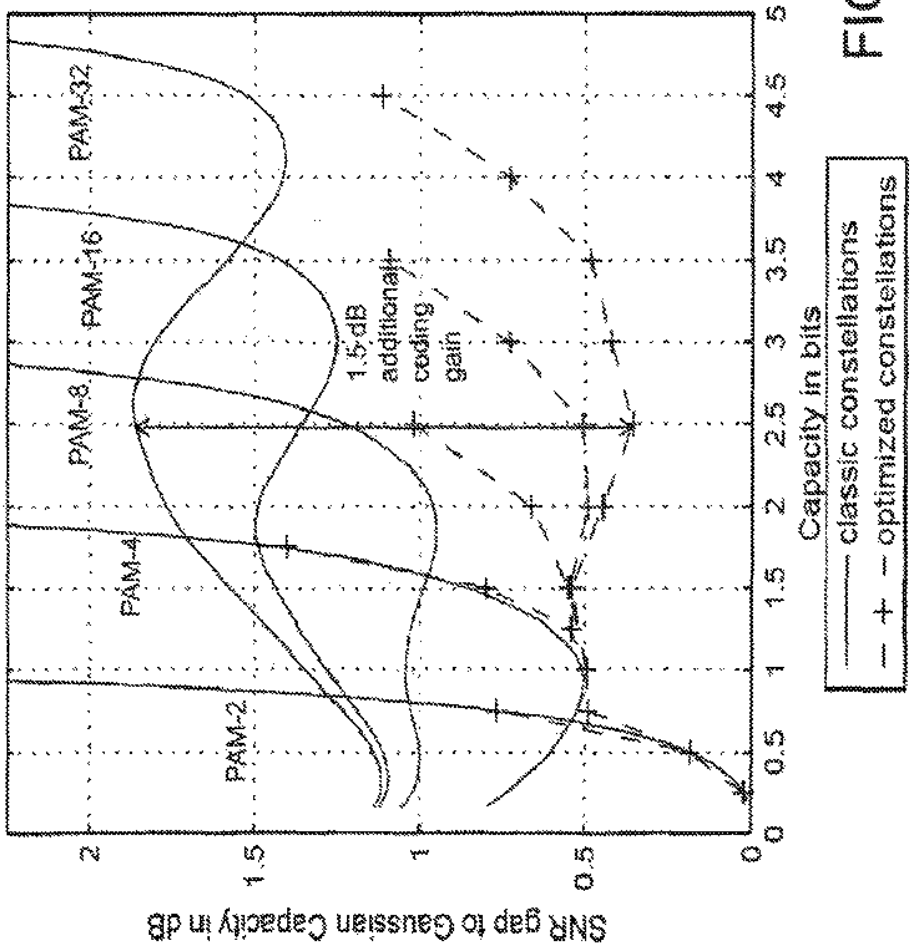

SNR gap to Gaussian Capacity in dB
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
1.5 dB additional coding gain
classic constellations
optimized constellations

FIG. 8a

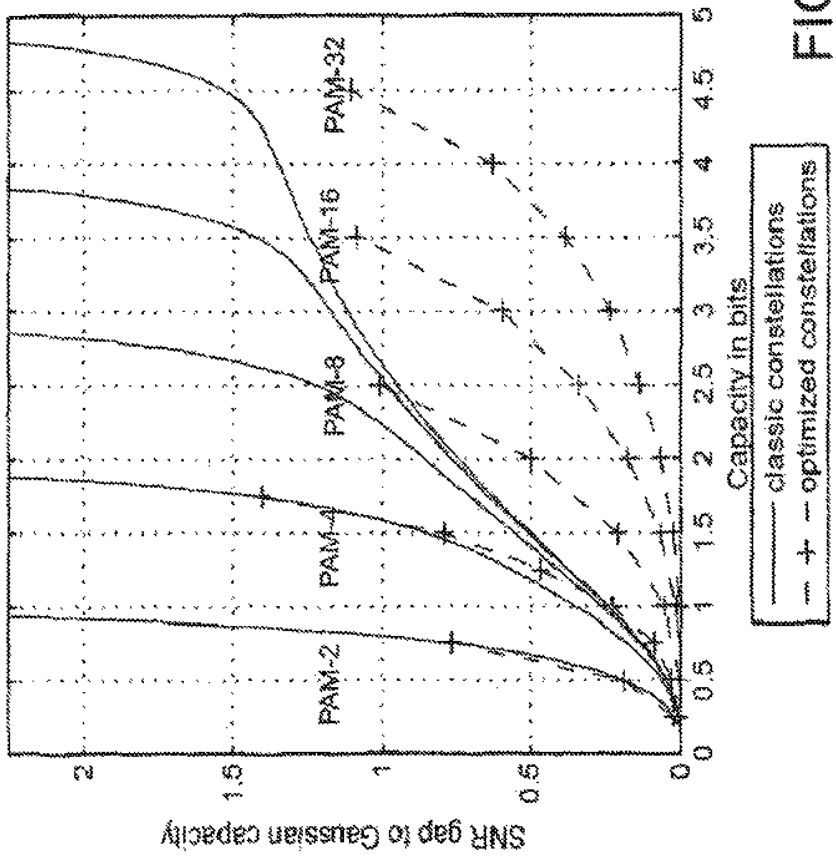

SNR gap to Gaussian capacity
Capacity in bits
classic constellations
optimized constellations
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8b

FER
SNR (dB)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20
$\Delta_{1/2}$ = 1.3 dB
$\Delta_{1/2}$ = 1.3 dB
PID Opt PAM32 w/ iter, k=4096
PAM32 w/ iter, k=4096
PID Opt PAM32, k=4096
PAM32, k=4096
PID Opt PAM32, k=16384
PAM32, k=16384
90

FIG. 9

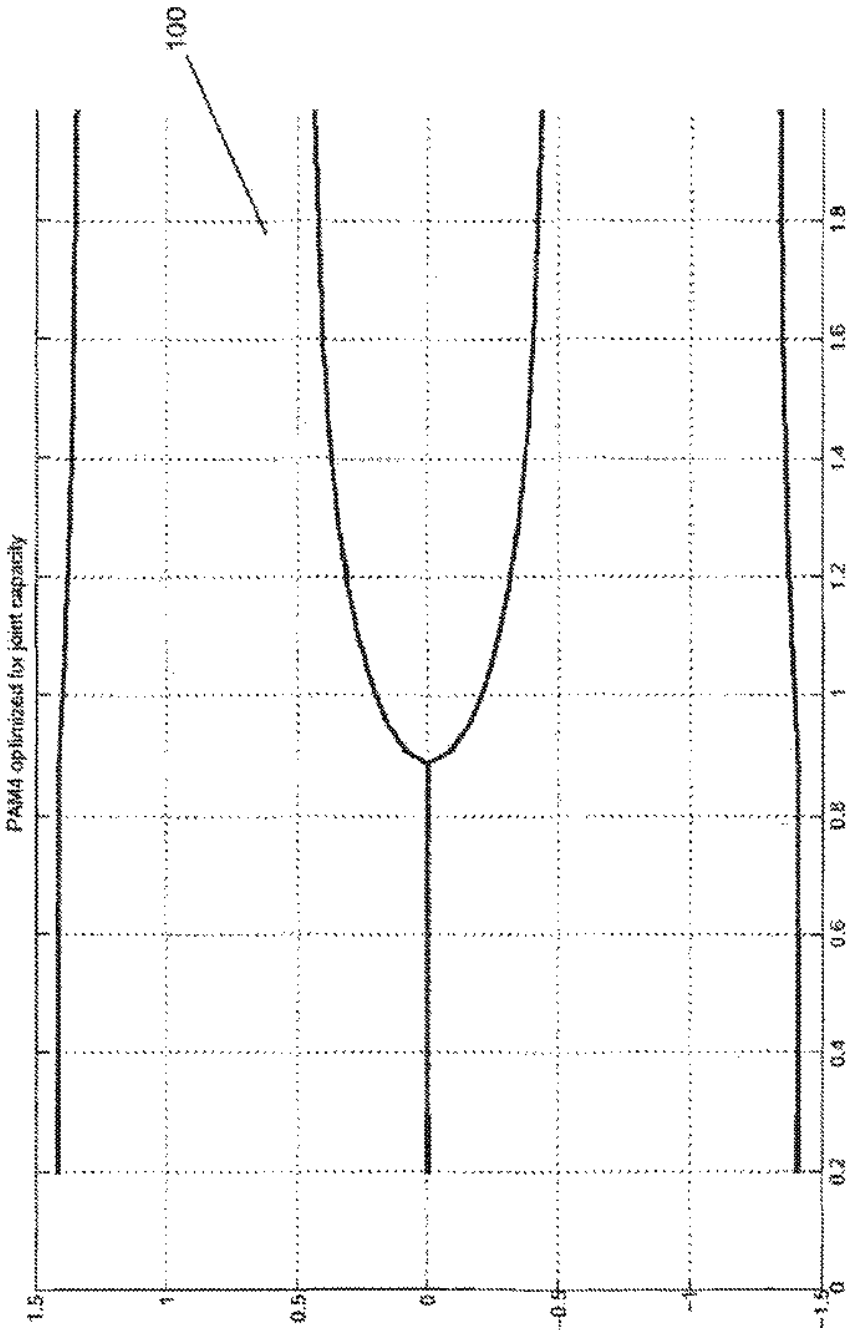

100
PAM4 optimized for joint capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10a

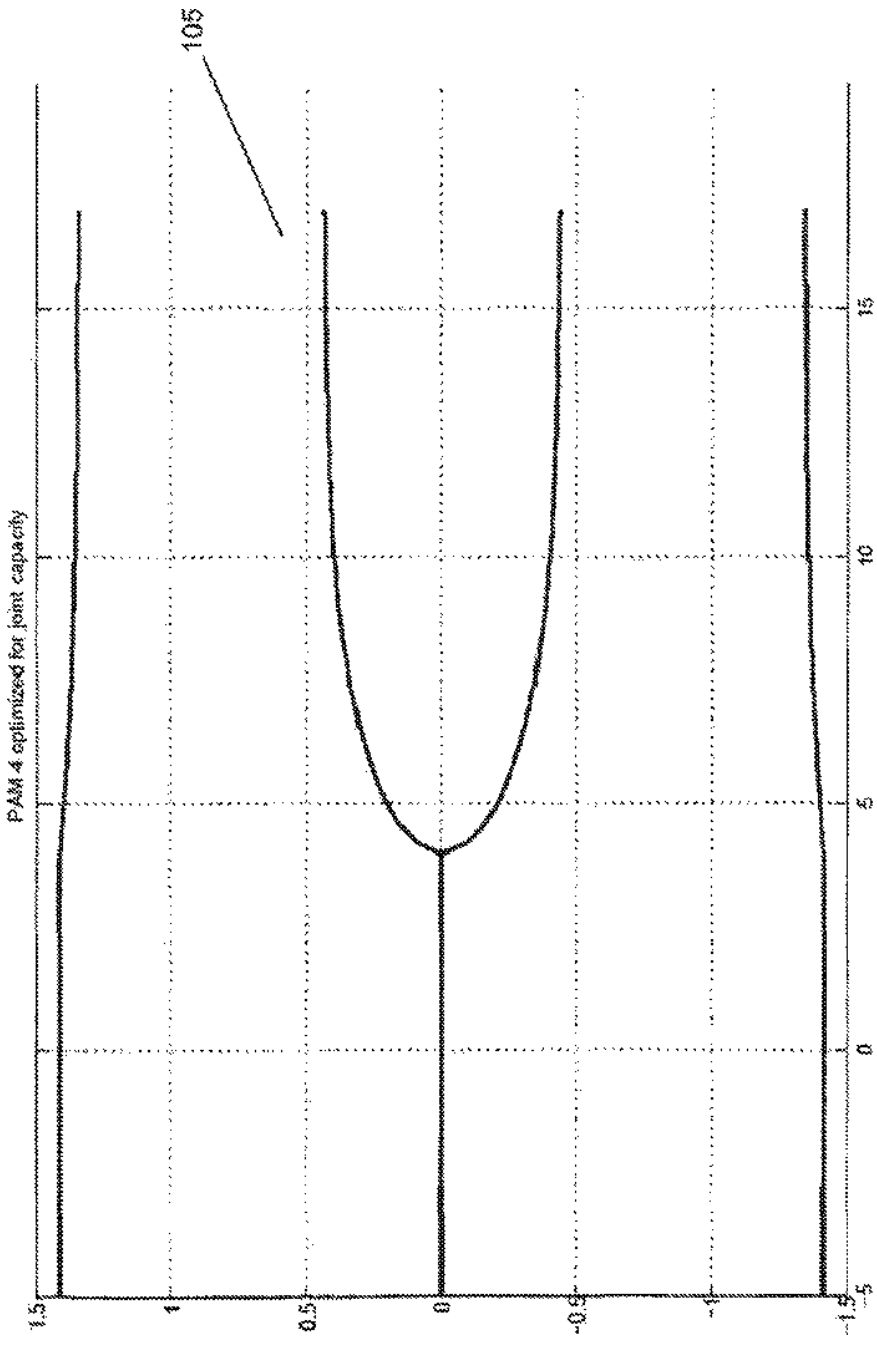

105
PAM 4 optimized for joint capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10b

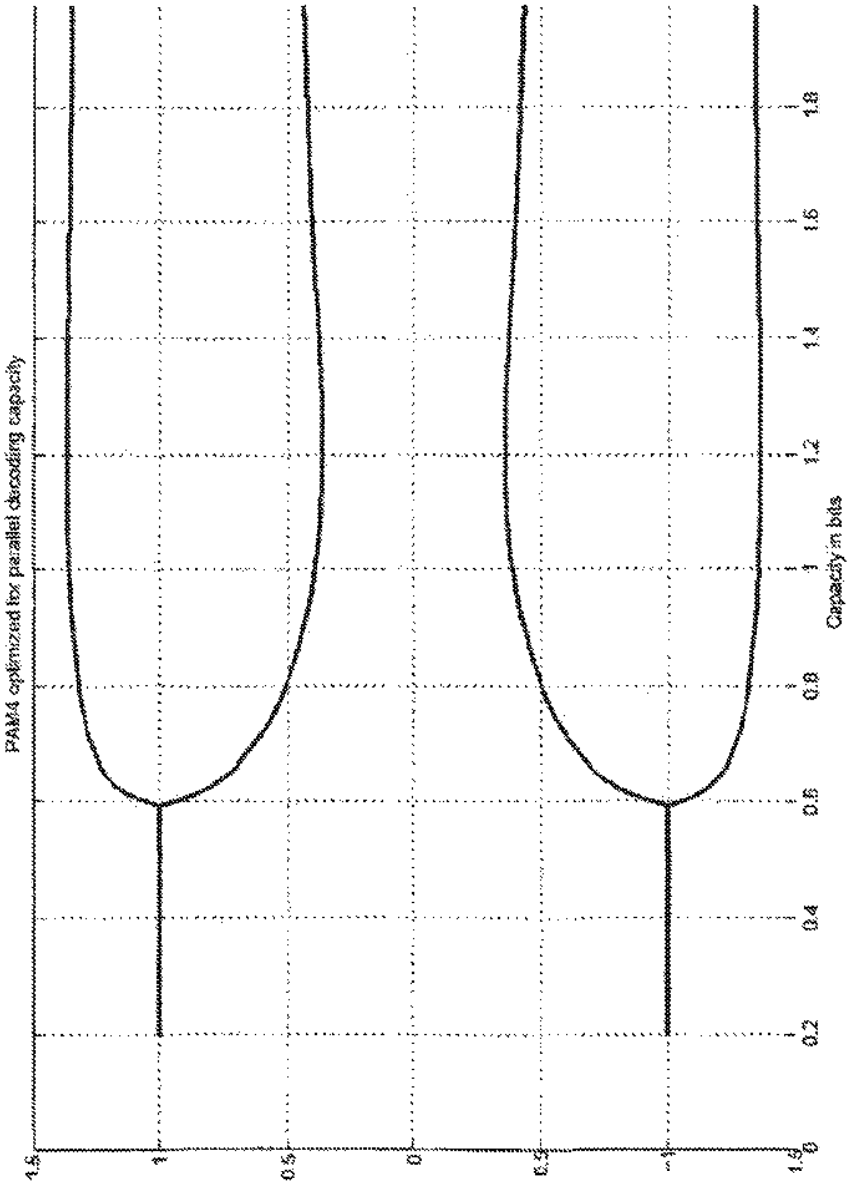

PAM4 optimized for parallel decoding capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10c

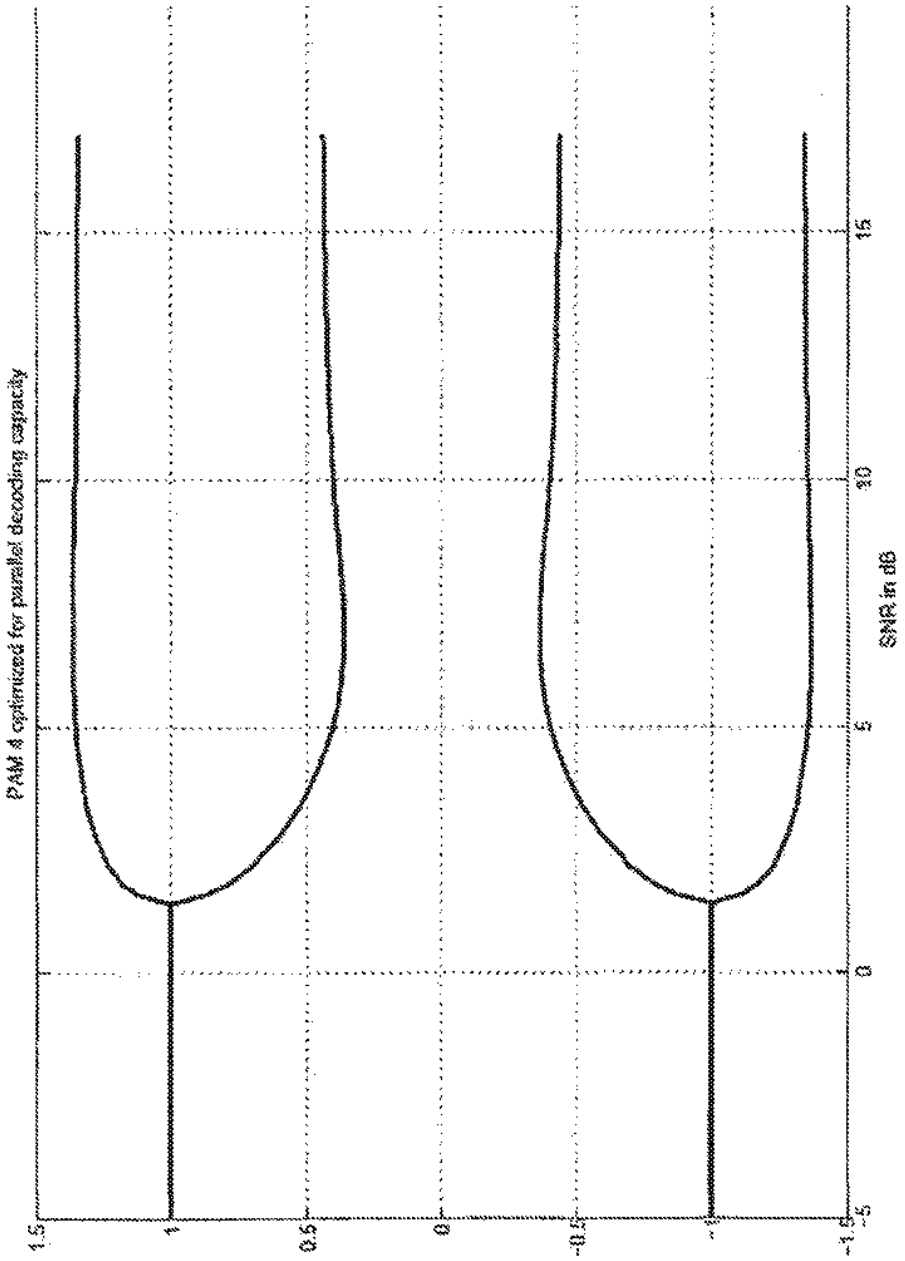

PAM 4 optimized for parallel decoding capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10d

PAM-4 constellations optimized for joint capacity at different rates

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.03	2.71	5.00	7.15	9.24
x_0	-1.41	-1.41	-1.40	-1.37	-1.36
x_1	0.00	0.00	-0.20	-0.33	-0.39
x_2	0.00	0.00	0.20	0.33	0.39
x_3	1.41	1.41	1.40	1.37	1.36

FIG. 11a

PAM-4 constellations optimized for parallel decoding capacity at different

(bps)	0.50	0.75	1.00	1.25	1.50
(SNR)	0.19	3.11	5.26	7.22	9.25
x_0	-1.00	-1.30	-1.36	-1.37	-1.36
x_1	-1.00	-0.56	-0.39	-0.33	-0.39
x_2	1.00	1.30	1.36	0.33	1.36
x_3	1.00	0.56	0.39	1.37	0.39

FIG. 11b

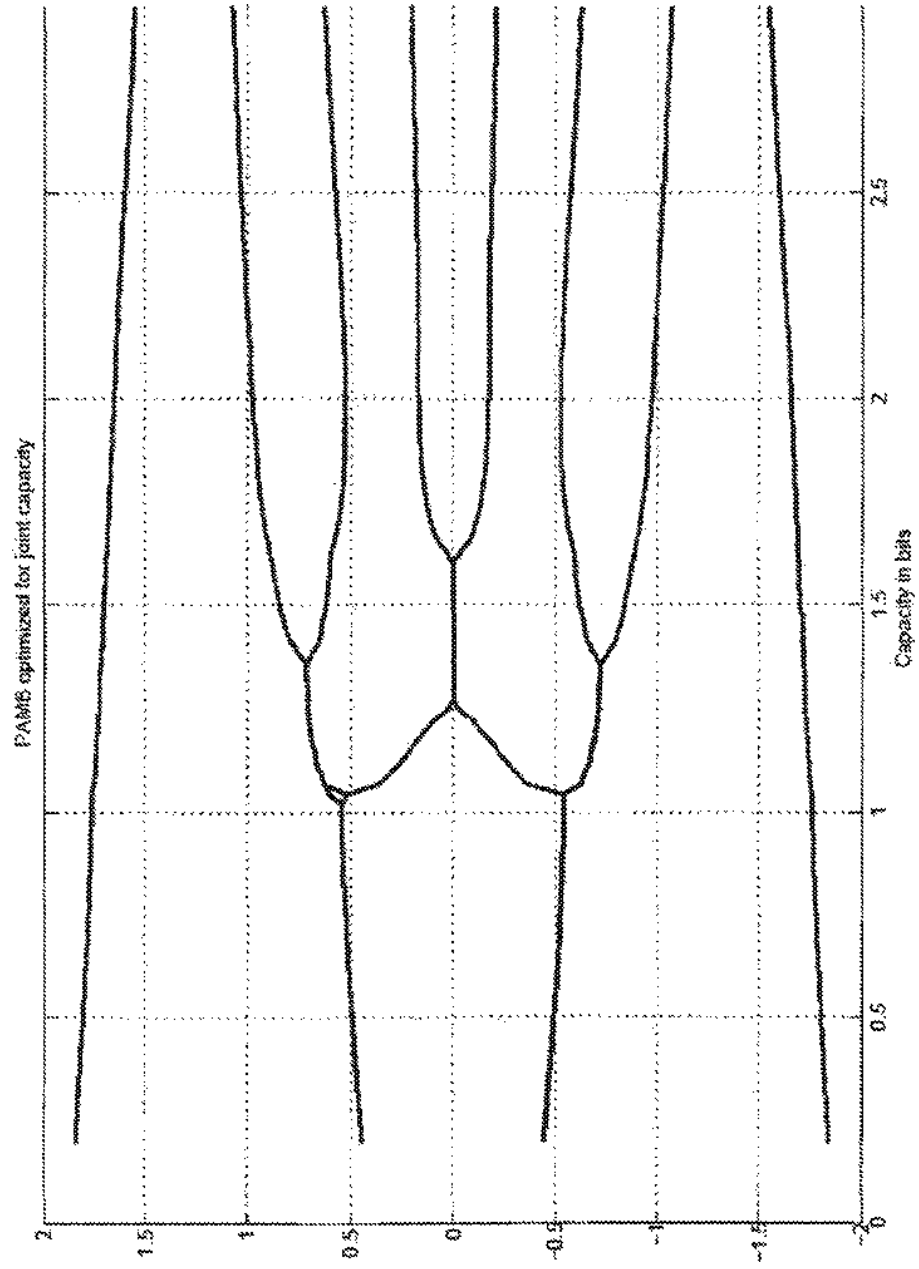

PAM6 optimized for joint capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12a

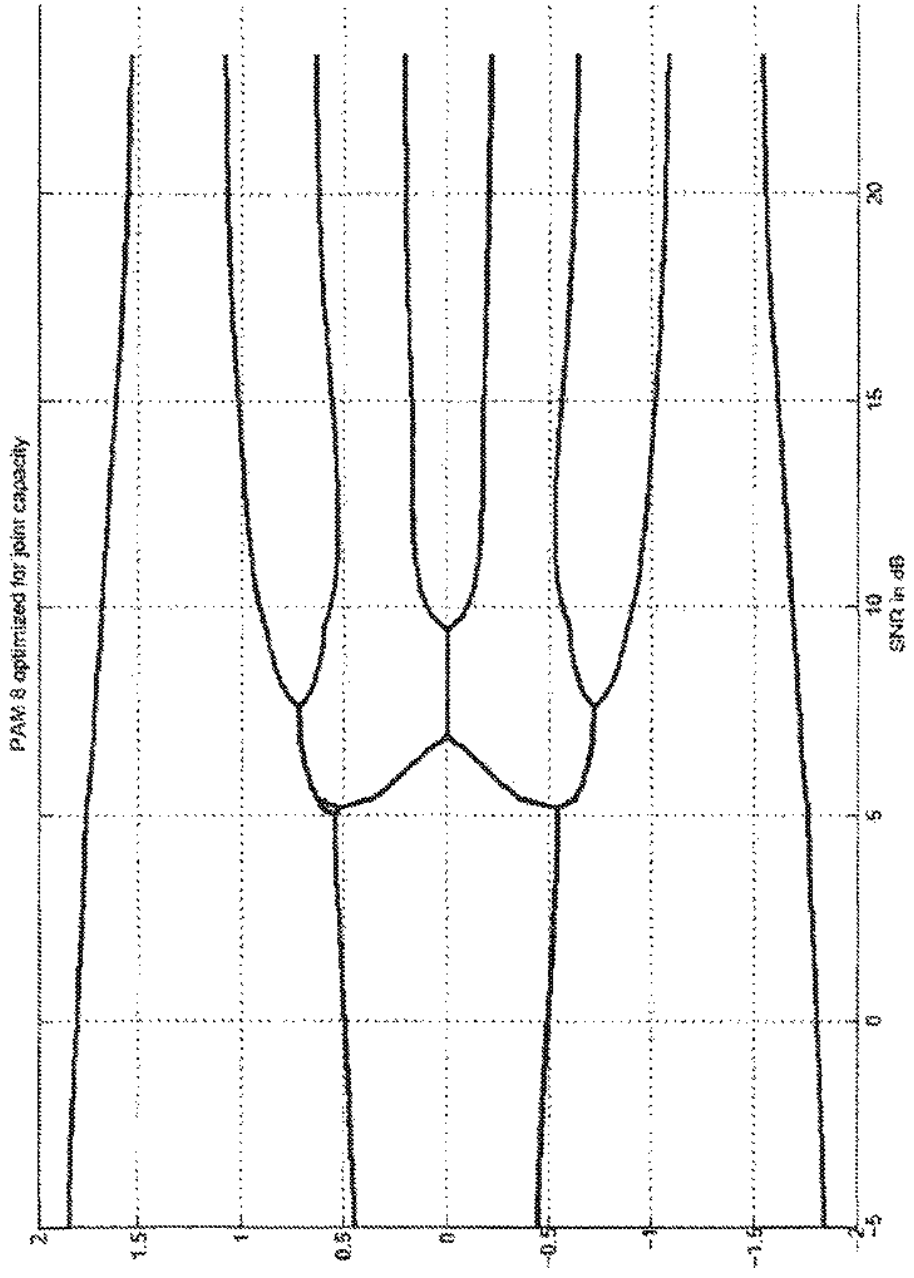

PAM-8 optimized for joint capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12b

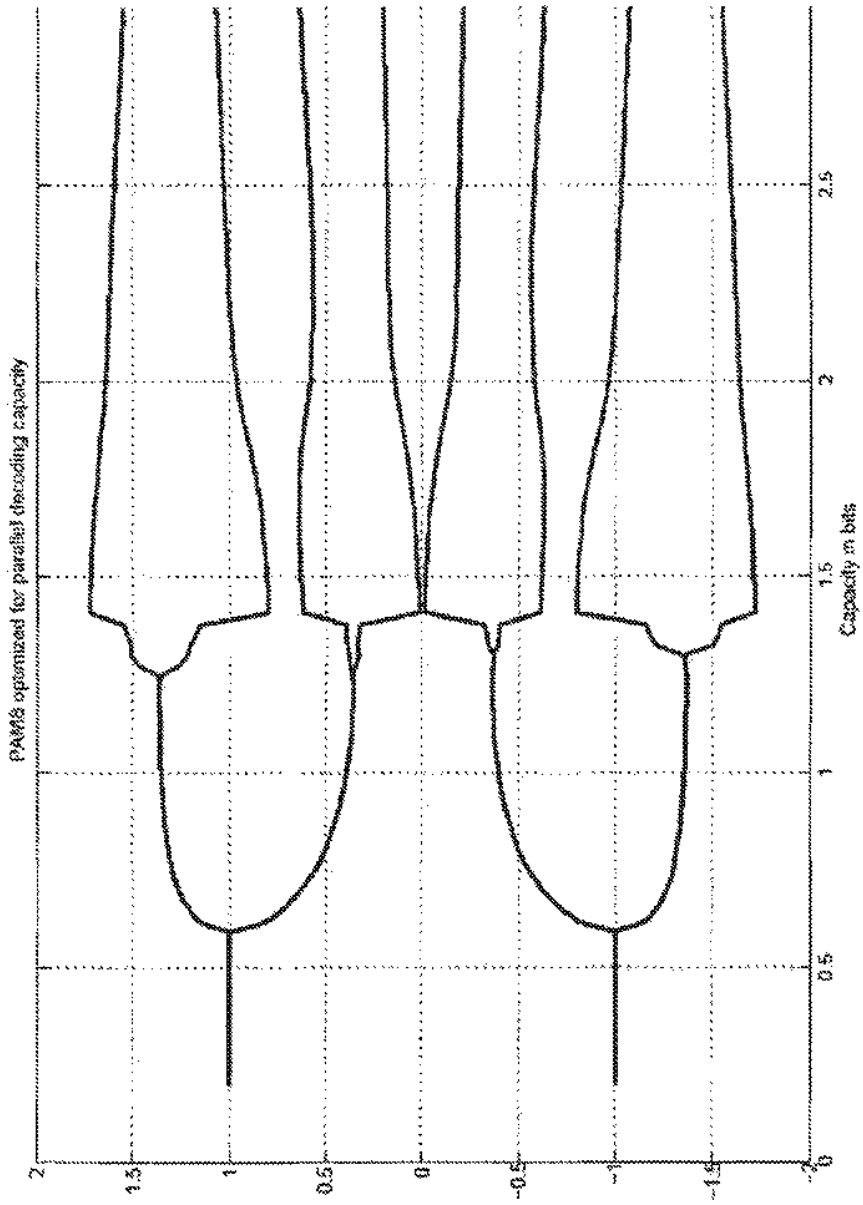

PAM8 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12c

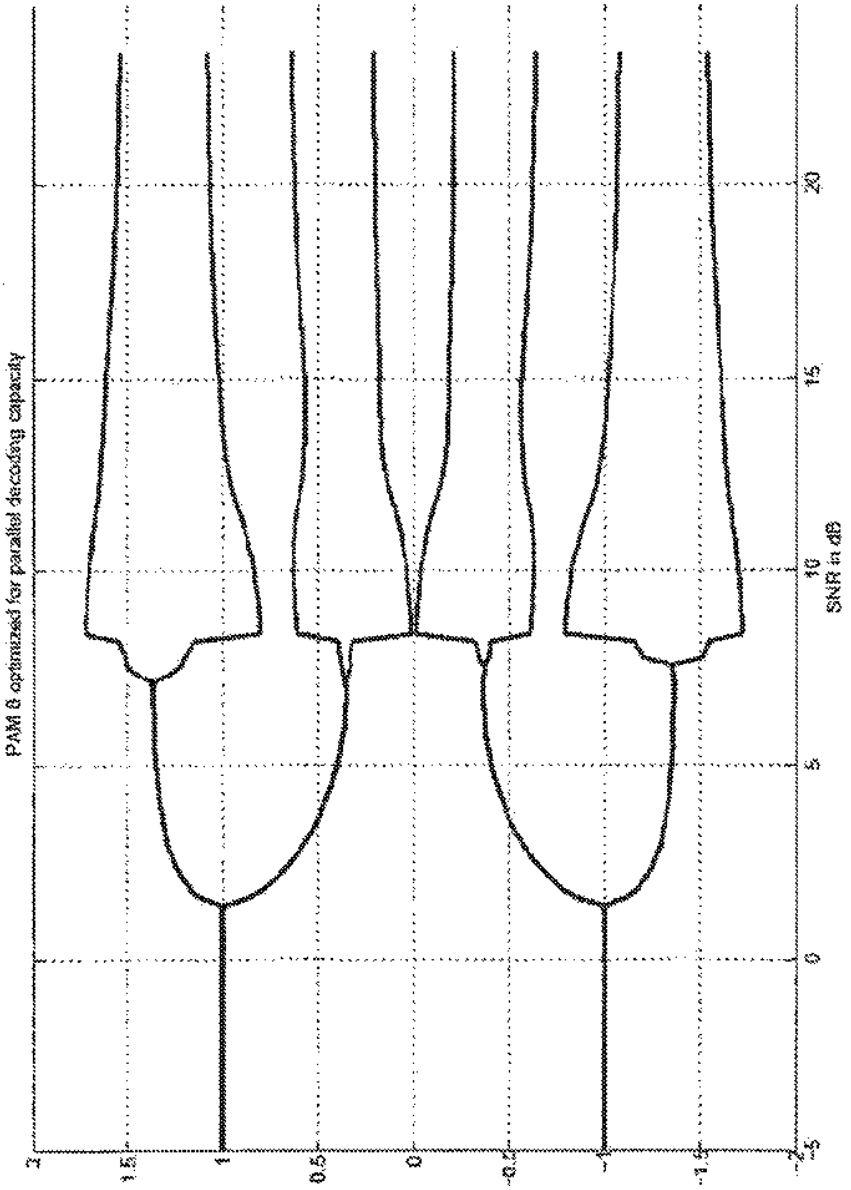

PAM 8 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12d

PAM-8 constellations optimized for joint capacity at different rates

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.00	4.82	8.66	12.26	15.93
x_0	-1.81	-1.76	-1.70	-1.66	-1.60
x_1	-0.50	-0.55	-0.84	-0.97	-1.03
x_2	-0.50	-0.55	-0.63	-0.53	-0.58
x_3	-0.50	-0.55	-0.00	-0.17	-0.19
x_4	0.50	0.55	0.00	0.17	0.19
x_5	0.50	0.55	0.63	0.53	0.58
x_6	0.50	0.55	0.84	0.97	1.03
x_7	1.81	1.76	1.70	1.66	1.60

FIG. 13a

PAM-8 constellations optimized for parallel decoding capacity at different

(bps)	0.5	1.0	1.5	2.0	2.5
(SNR)	0.19	5.27	9.00	12.42	15.93
x_0	-1.00	-1.36	-1.72	-1.64	-1.60
x_1	-1.00	-1.36	-0.81	-0.97	-1.03
x_2	-1.00	-0.39	1.72	1.64	-0.19
x_3	-1.00	-0.39	-0.62	-0.58	-0.58
x_4	1.00	1.36	0.62	0.58	1.60
x_5	1.00	1.36	0.02	0.15	1.03
x_6	1.00	0.39	0.81	0.97	0.19
x_7	1.00	0.39	-0.02	-0.15	0.58

FIG. 13b

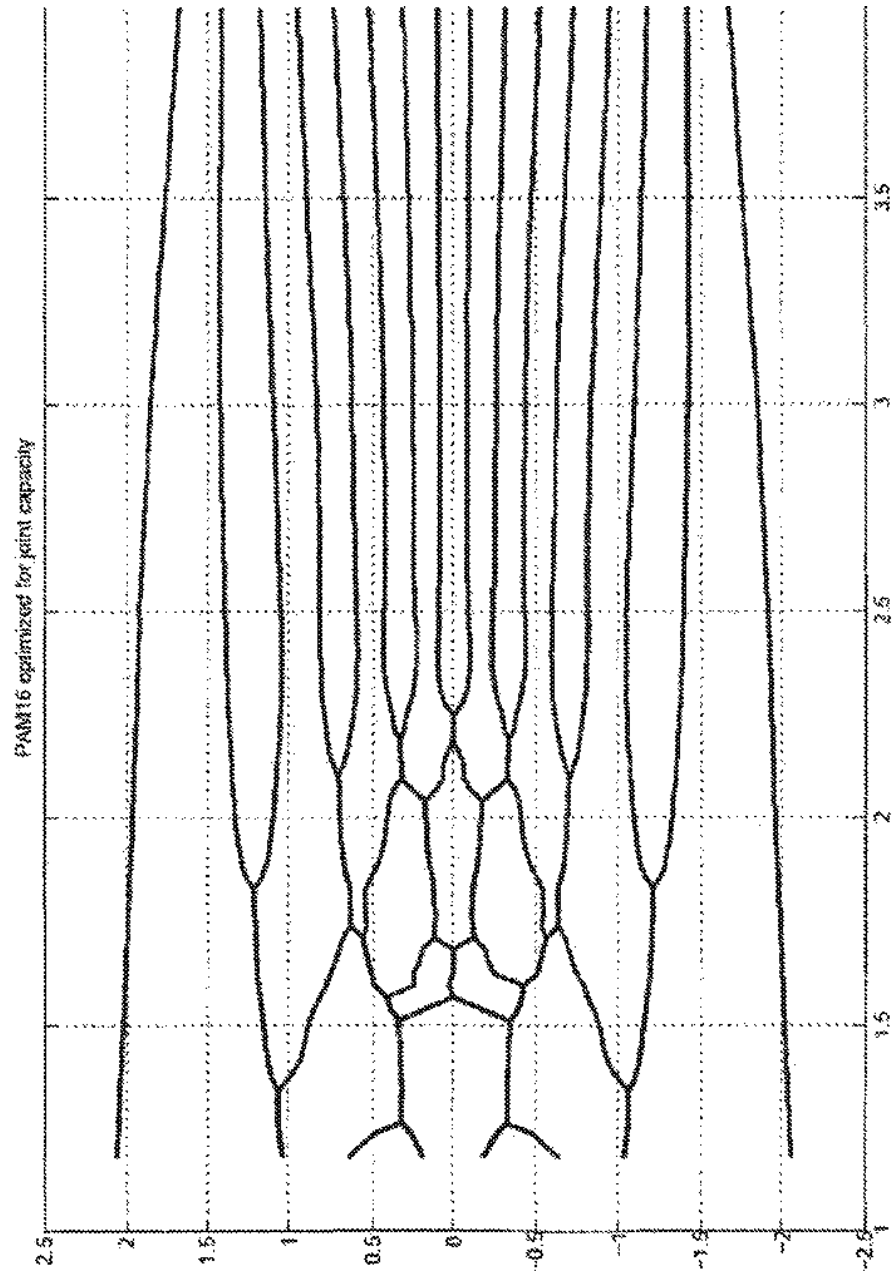

PAM16 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

FIG. 14a

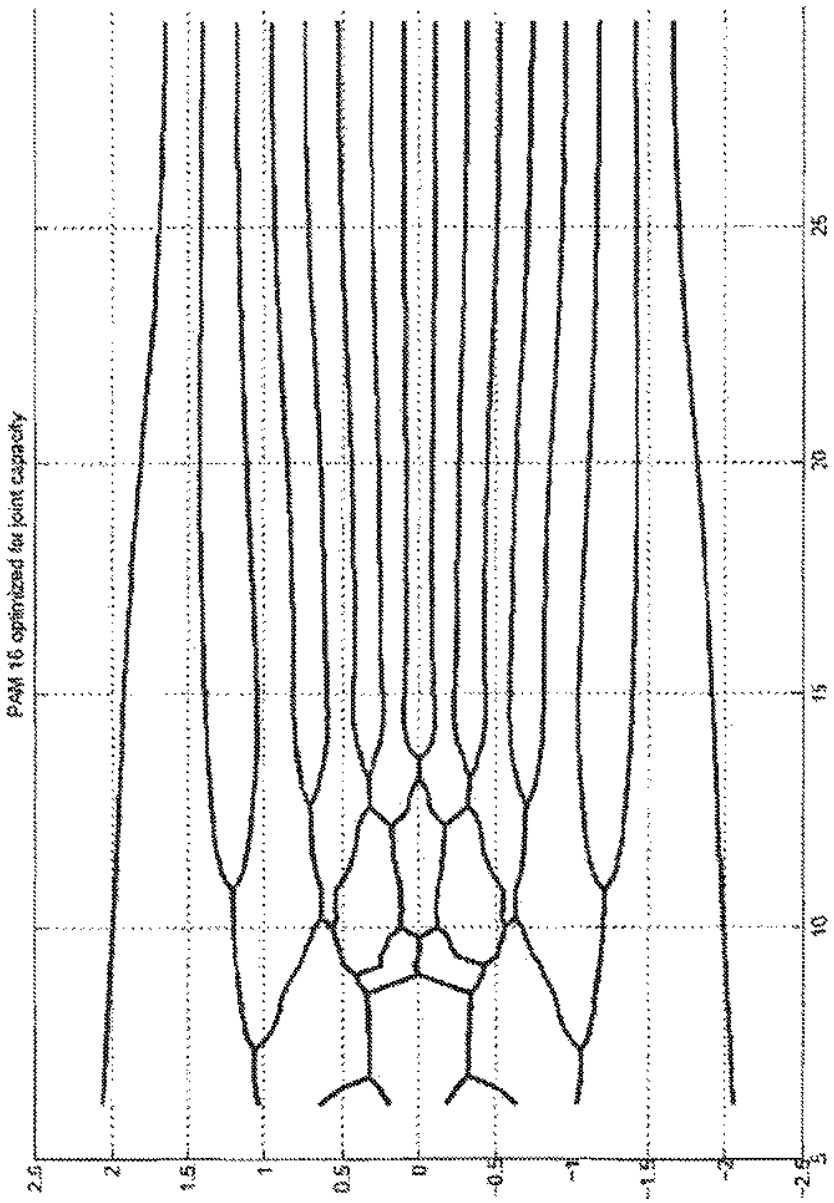

PAM 16 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14b

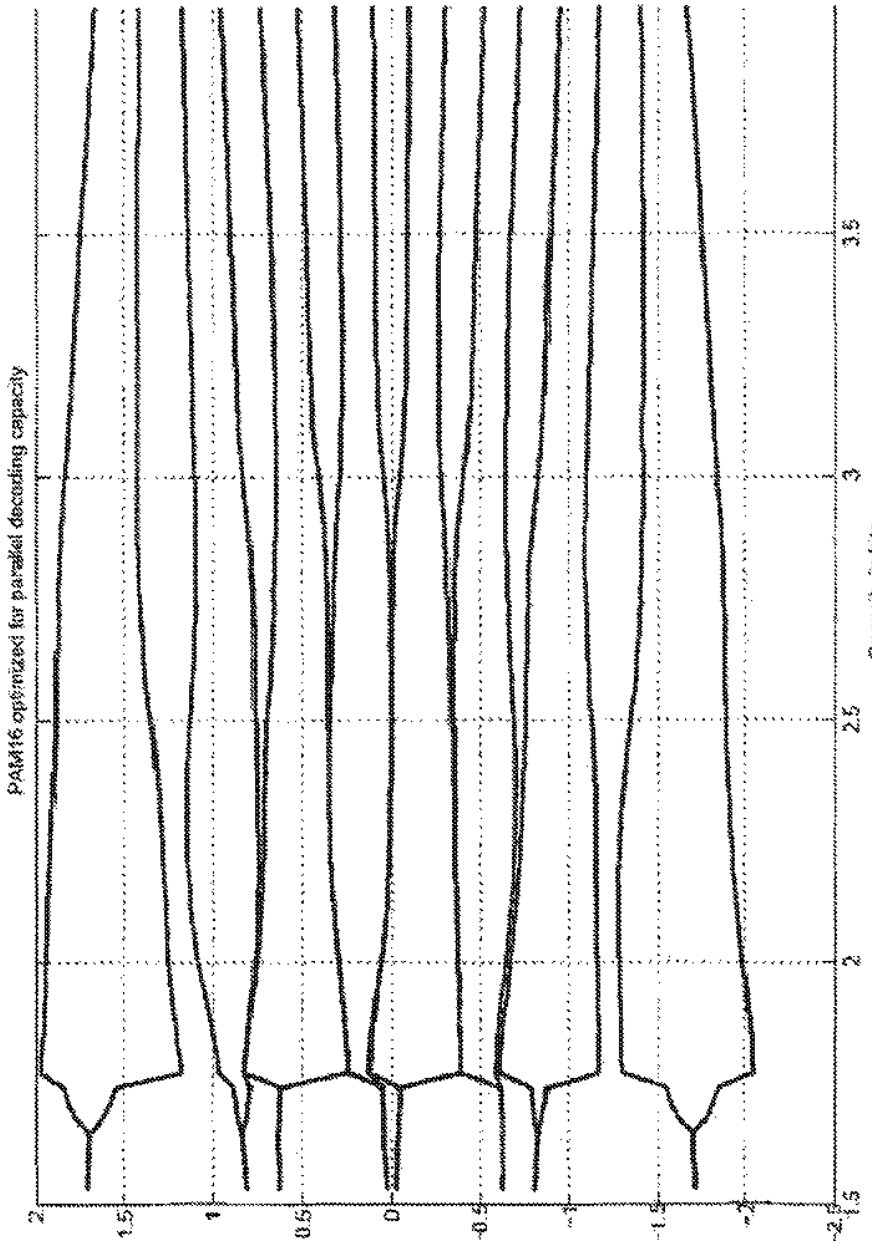

PAM16 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5

FIG. 14c

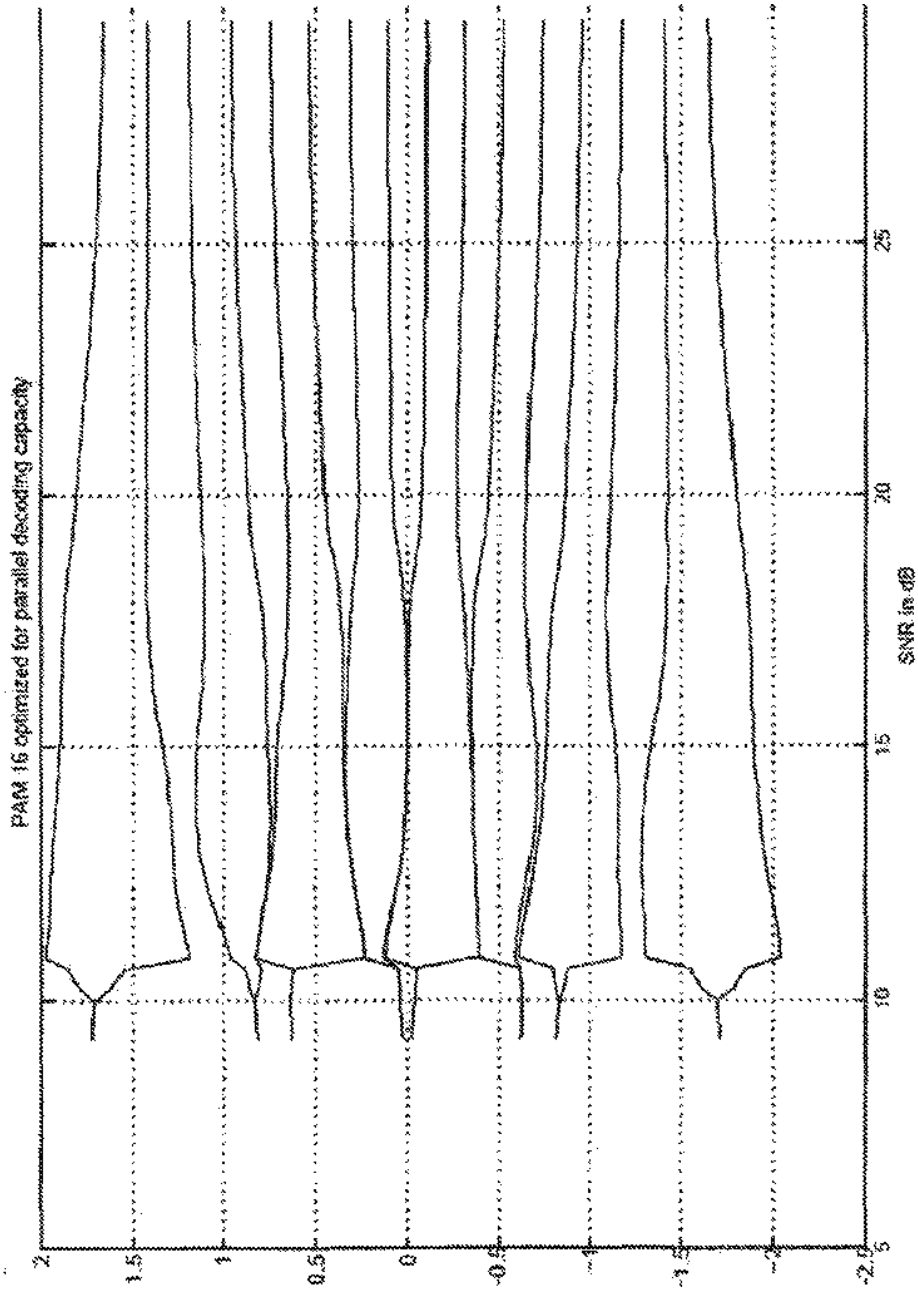

PAM 16 optimized for parallel decoding capacity
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
5
10
15
20
25
SNR in dB

FIG. 14d

PAM-16 constellations optimized for joint capacity at different rates

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	8.52	11.94	15.25	18.60	22.12
x_0	-2.02	-1.96	-1.91	-1.85	-1.76
x_1	-1.16	-1.33	-1.40	-1.42	-1.42
x_2	-1.16	-1.10	-1.05	-1.10	-1.15
x_3	-0.90	-0.69	-0.82	-0.84	-0.90
x_4	-0.34	-0.69	-0.60	-0.62	-0.68
x_5	-0.34	-0.40	-0.43	-0.43	-0.47
x_6	-0.34	-0.17	-0.24	-0.26	-0.28
x_7	-0.34	-0.17	-0.09	-0.08	-0.09
x_8	0.34	0.17	0.09	0.08	0.09
x_9	0.34	0.17	0.24	0.26	0.28
x_{10}	0.34	0.40	0.43	0.43	0.47
x_{11}	0.34	0.69	0.60	0.62	0.68
x_{12}	0.90	0.69	0.82	0.84	0.90
x_{13}	1.16	1.10	1.05	1.10	1.15
x_{14}	1.16	1.33	1.40	1.42	1.42
x_{15}	2.02	1.96	1.91	1.85	1.76

FIG. 15a

PAM-16 constellations optimized for parallel decoding capacity at different

(bps)	1.5	2.0	2.5	3.0	3.5
(SNR)	9.00	12.25	15.42	18.72	22.13
x_0	-1.72	-1.98	-1.89	-1.84	-1.75
x_1	-1.72	-1.29	-1.36	-1.42	-1.42
x_2	-0.81	1.94	1.89	1.84	1.75
x_3	-0.81	-1.17	-1.14	-1.11	-1.15
x_4	1.72	-0.38	-0.35	-0.40	-0.47
x_5	1.72	-0.65	-0.70	-0.65	-0.68
x_6	-0.62	-0.38	-0.34	-0.29	-0.28
x_7	-0.62	-0.68	-0.76	-0.83	-0.90
x_8	0.62	1.09	1.13	1.11	1.15
x_9	0.62	0.76	0.76	0.84	0.90
x_{10}	0.02	1.26	1.35	1.42	1.42
x_{11}	0.02	0.76	0.70	0.65	0.68
x_{12}	0.81	0.06	0.00	0.05	0.09
x_{13}	0.81	0.29	0.34	0.29	0.28
x_{14}	-0.02	0.06	0.00	-0.05	-0.09
x_{15}	-0.02	0.29	0.35	0.40	0.47

FIG. 15b

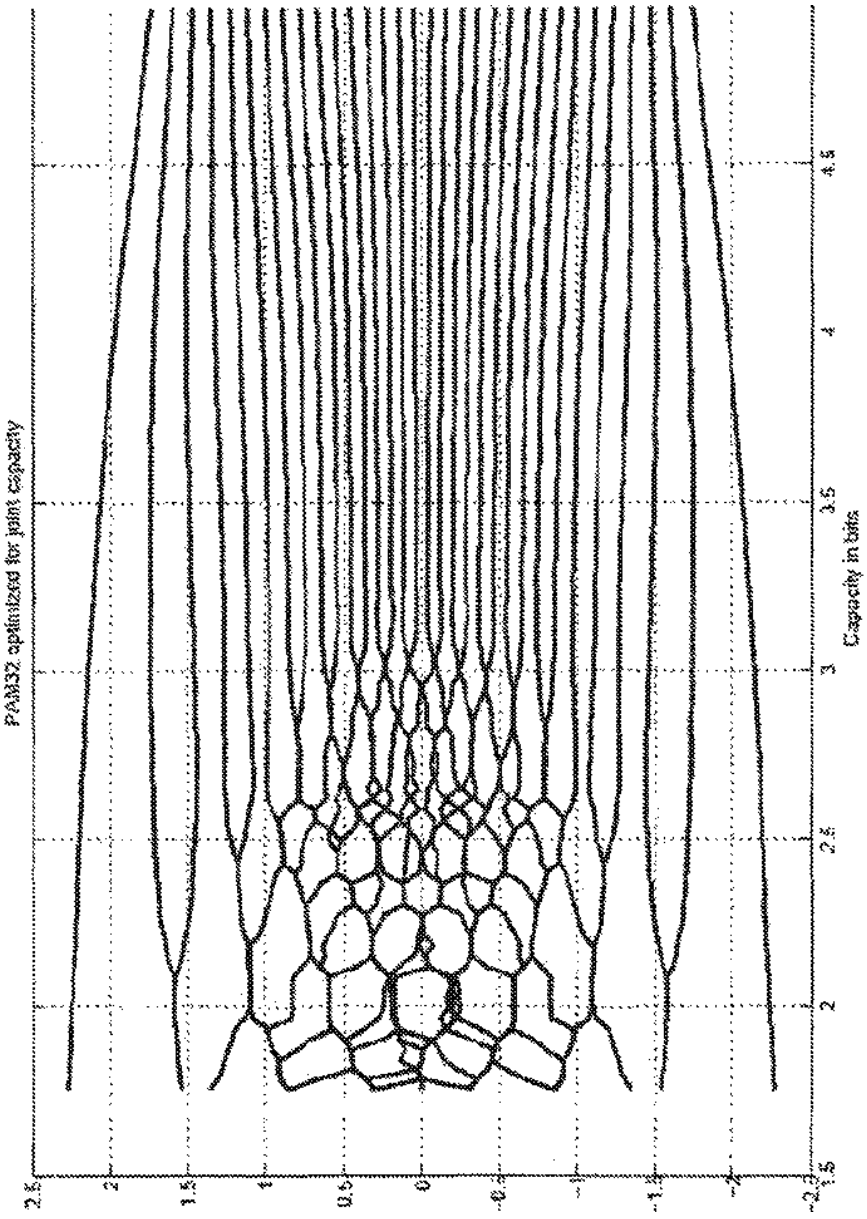

PAM32 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16a

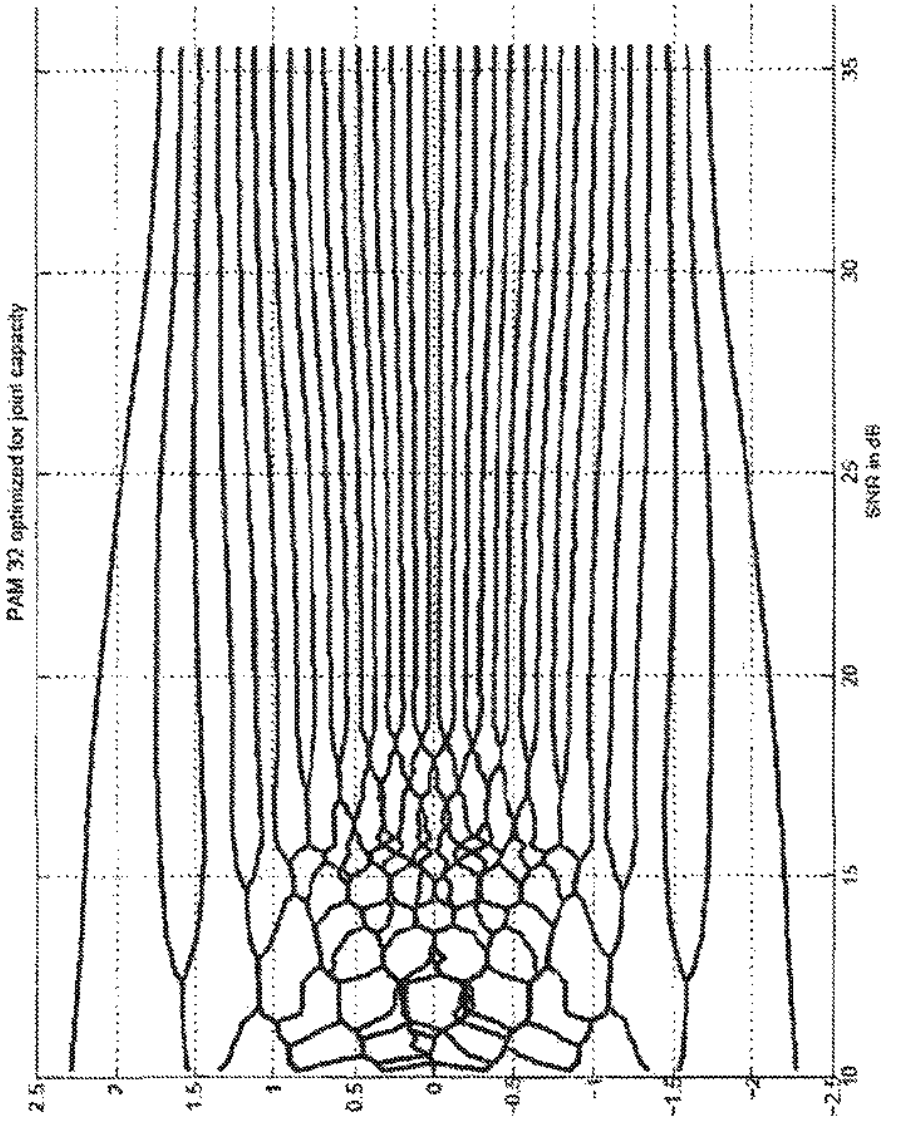

PAM 32 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

FIG. 16b

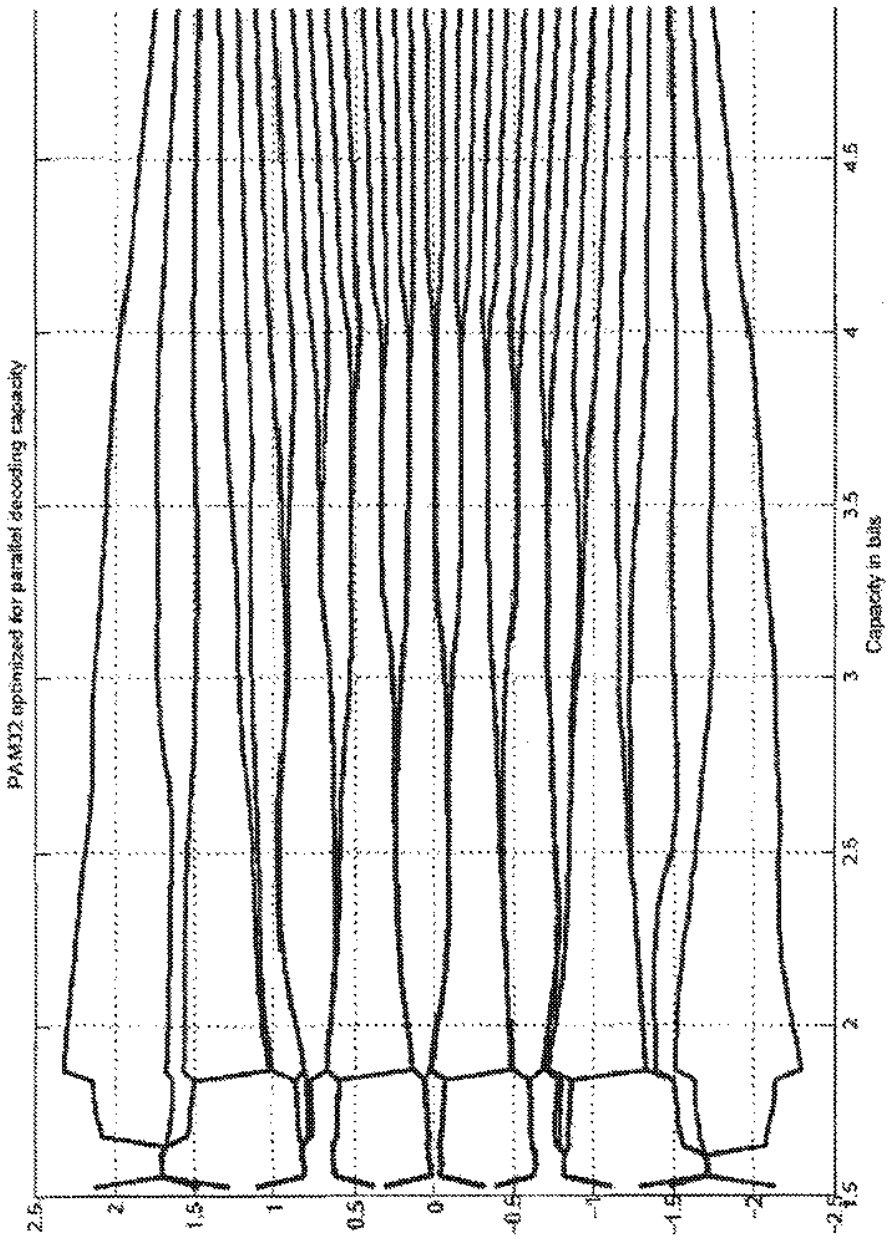

PAM32 optimized for parallel decoding capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16c

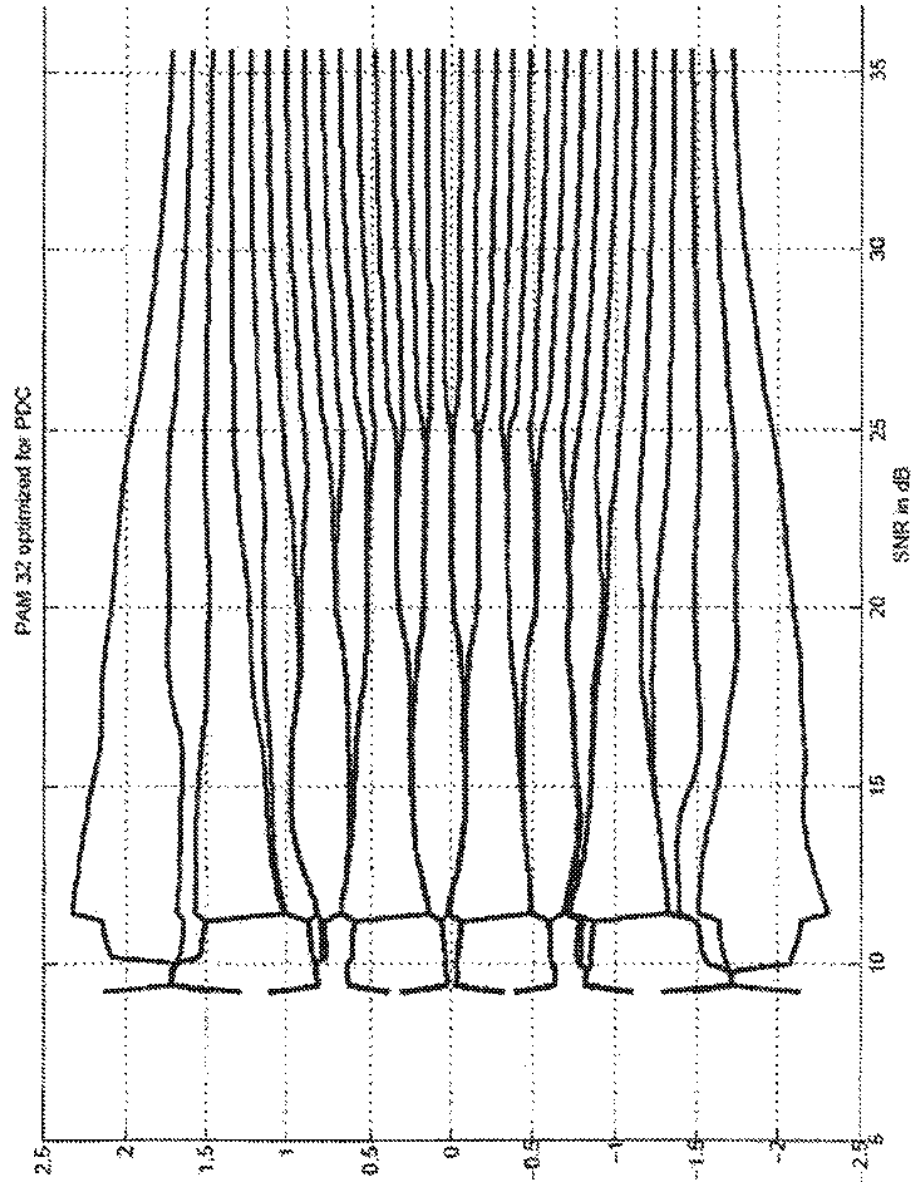

PAM 32 optimized for PDC
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

FIG. 16d

PAM-32 constellations optimized for joint capacity at different rates

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	11.83	15.05	18.23	21.42	24.69	28.19
x_0	-2.25	-2.19	-2.14	-2.07	-1.97	-1.85
x_1	-1.58	-1.71	-1.74	-1.74	-1.72	-1.66
x_2	-1.58	-1.46	-1.46	-1.49	-1.51	-1.50
x_3	-1.10	-1.23	-1.27	-1.29	-1.33	-1.35
x_4	-1.10	-1.13	-1.11	-1.13	-1.17	-1.21
x_5	-1.10	-0.90	-0.98	-0.99	-1.02	-1.08
x_6	-0.83	-0.90	-0.85	-0.87	-0.90	-0.95
x_7	-0.60	-0.75	-0.75	-0.76	-0.78	-0.84
x_8	-0.60	-0.58	-0.63	-0.65	-0.67	-0.73
x_9	-0.60	-0.58	-0.57	-0.56	-0.57	-0.62
x_{10}	-0.60	-0.49	-0.42	-0.46	-0.48	-0.52
x_{11}	-0.24	-0.29	-0.40	-0.38	-0.39	-0.42
x_{12}	-0.21	-0.28	-0.24	-0.29	-0.30	-0.32
x_{13}	-0.20	-0.28	-0.24	-0.21	-0.21	-0.23
x_{14}	-0.20	-0.09	-0.09	-0.12	-0.13	-0.14
x_{15}	-0.16	-0.00	-0.07	-0.04	-0.04	-0.05
x_{16}	0.16	0.00	0.07	0.04	0.04	0.05
x_{17}	0.19	0.09	0.09	0.12	0.13	0.14
x_{18}	0.21	0.28	0.24	0.21	0.21	0.23
x_{19}	0.22	0.28	0.24	0.29	0.30	0.32
x_{20}	0.23	0.28	0.41	0.38	0.39	0.42
x_{21}	0.60	0.49	0.42	0.46	0.48	0.52
x_{22}	0.60	0.58	0.57	0.56	0.57	0.62
x_{23}	0.60	0.58	0.62	0.65	0.67	0.73
x_{24}	0.60	0.75	0.75	0.76	0.78	0.84
x_{25}	0.83	0.90	0.85	0.87	0.90	0.95
x_{26}	1.10	0.90	0.98	0.99	1.02	1.08
x_{27}	1.10	1.13	1.11	1.13	1.17	1.21
x_{28}	1.10	1.23	1.27	1.29	1.33	1.35
x_{29}	1.58	1.46	1.46	1.49	1.51	1.50
x_{30}	1.58	1.71	1.74	1.74	1.72	1.66
x_{31}	2.25	2.19	2.14	2.07	1.97	1.85

FIG. 17a

PAM-32 constellations optimized for parallel decoding capacity at different

(bps)	2.0	2.5	3.0	3.5	4.0	4.5
(SNR)	12.21	15.27	18.42	21.52	24.79	28.20
x_0	-2.25	-2.16	-2.14	-2.05	-1.97	-1.85
x_1	-1.52	-1.64	-1.75	-1.74	-1.72	-1.66
x_2	2.30	2.19	-1.31	2.05	1.97	-1.35
x_3	-1.39	-1.48	-1.43	-1.49	-1.51	-1.49
x_4	1.56	1.54	2.14	-0.96	-1.03	1.85
x_5	-1.31	-1.23	1.75	-1.15	-1.17	1.66
x_6	1.67	1.65	-1.07	-0.91	-0.90	-1.21
x_7	-1.31	-1.24	-1.04	-1.28	-1.33	-1.08
x_8	-0.48	-0.43	-0.36	-0.17	-0.17	-0.42
x_9	-0.72	-0.76	-0.36	-0.34	-0.31	-0.52
x_{10}	-0.48	-0.43	-0.62	-0.17	-0.15	-0.73
x_{11}	-0.73	-0.76	-0.62	-0.34	-0.35	-0.62
x_{12}	-0.48	-0.42	-0.29	-0.71	-0.67	-0.33
x_{13}	-0.76	-0.86	-0.29	-0.52	-0.55	-0.23
x_{14}	-0.48	-0.42	-0.77	-0.72	-0.77	-0.84
x_{15}	-0.76	-0.86	-0.77	-0.52	-0.48	-0.96
x_{16}	0.87	0.98	1.07	1.49	1.51	1.21
x_{17}	0.66	0.63	1.04	1.28	1.33	1.08
x_{18}	0.87	0.98	0.77	1.74	1.72	0.84
x_{19}	0.66	0.63	0.77	1.15	1.17	0.96
x_{20}	1.07	1.13	1.31	0.72	0.77	1.35
x_{21}	0.66	0.59	1.43	0.91	0.90	1.49
x_{22}	1.05	1.10	0.62	0.71	0.67	0.73
x_{23}	0.66	0.60	0.62	0.96	1.03	0.62
x_{24}	-0.01	-0.08	0.02	0.00	0.01	0.05
x_{25}	0.17	0.25	0.02	0.17	0.15	0.14
x_{26}	-0.01	-0.08	0.29	0.00	-0.01	0.33
x_{27}	0.17	0.25	0.29	0.17	0.17	0.23
x_{28}	-0.01	-0.08	-0.02	0.52	0.48	-0.05
x_{29}	0.17	0.25	-0.02	0.34	0.35	-0.14
x_{30}	-0.01	-0.08	0.36	0.52	0.55	0.42
x_{31}	0.17	0.25	0.36	0.34	0.31	0.52

FIG. 17b

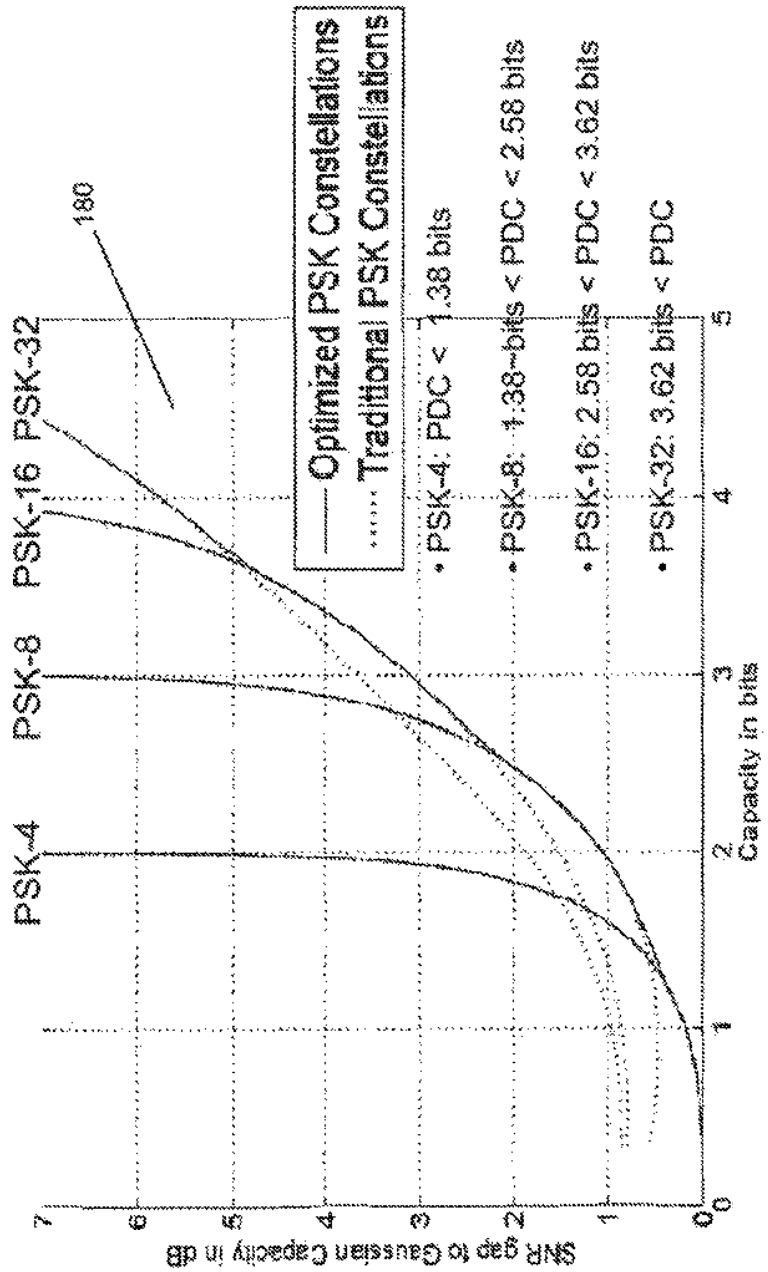

180
PSK-4
PSK-8
PSK-16
PSK-32
Optimized PSK Constellations
Traditional PSK Constellations
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38~bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
SNR gap to Gaussian Capacity in dB
Capacity in bits
0
1
2
3
4
5
6
7

FIG. 18

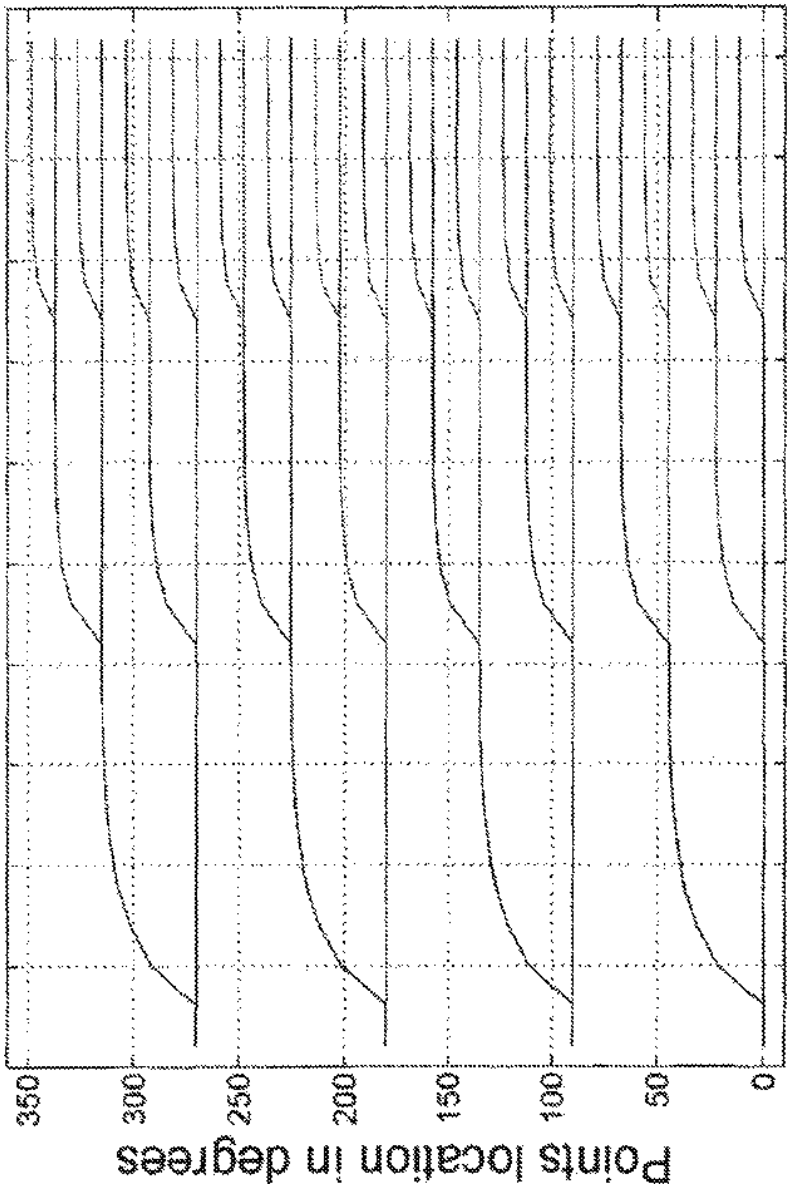

Points location in degrees
350
300
250
200
150
100
50
0
0
2
4
6
8
10
12
14
16
18
20
SNR in dB

FIG. 19

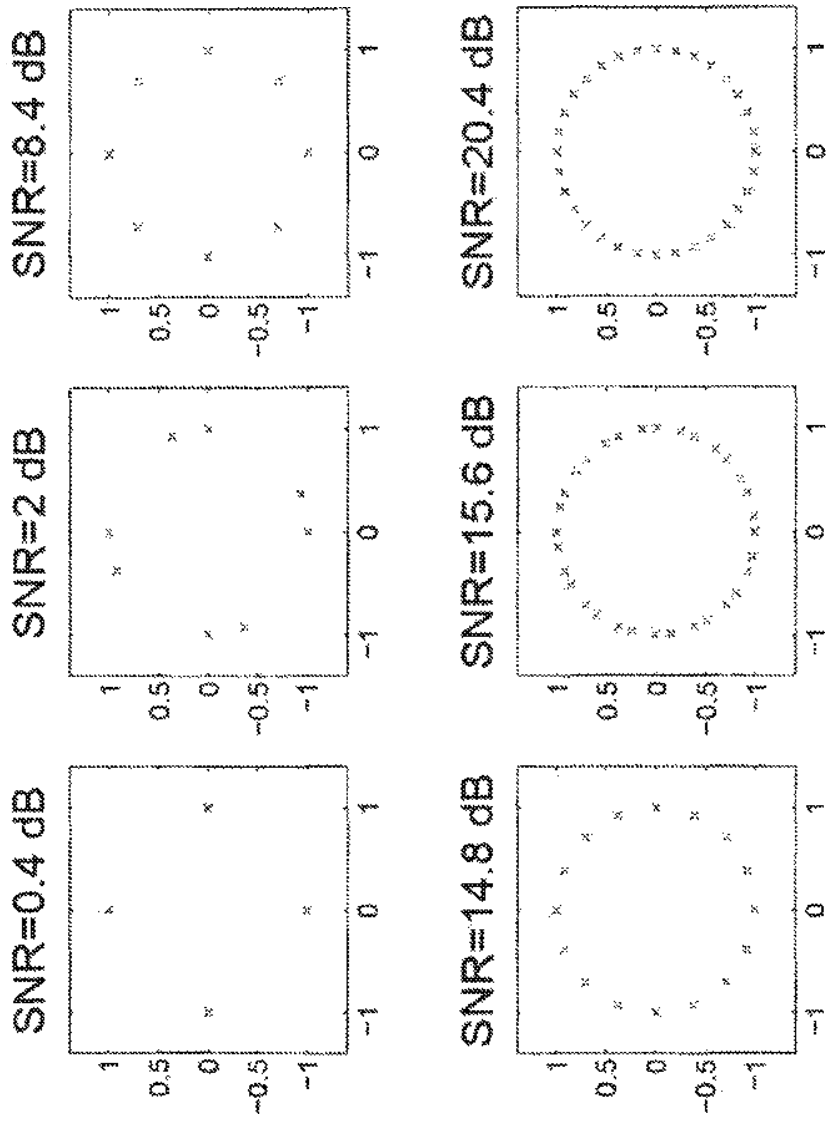


FIG. 20

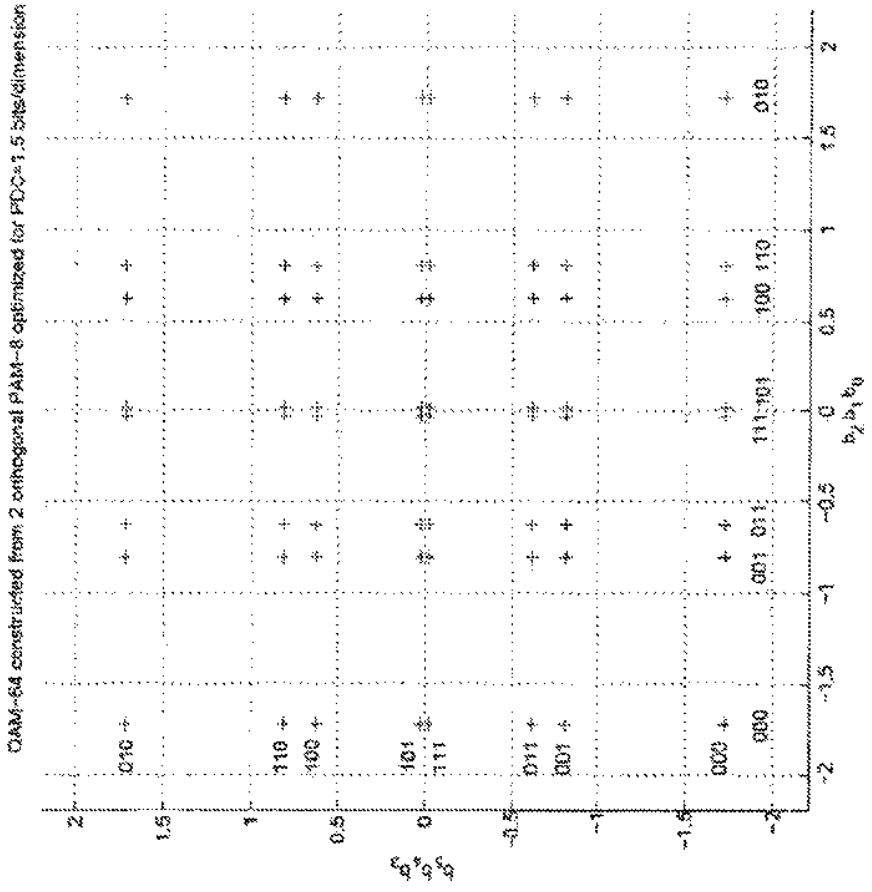

OAM-64 constructed from 2 orthogonal PAM-8 optimized for FDC=1.5 bits/dimension
010
110
100
101
111
011
001
000
010
110
100
111
101
011
001
000
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2
-1.5
-1
-0.5
0
0.5
1
1.5
2

FIG. 21

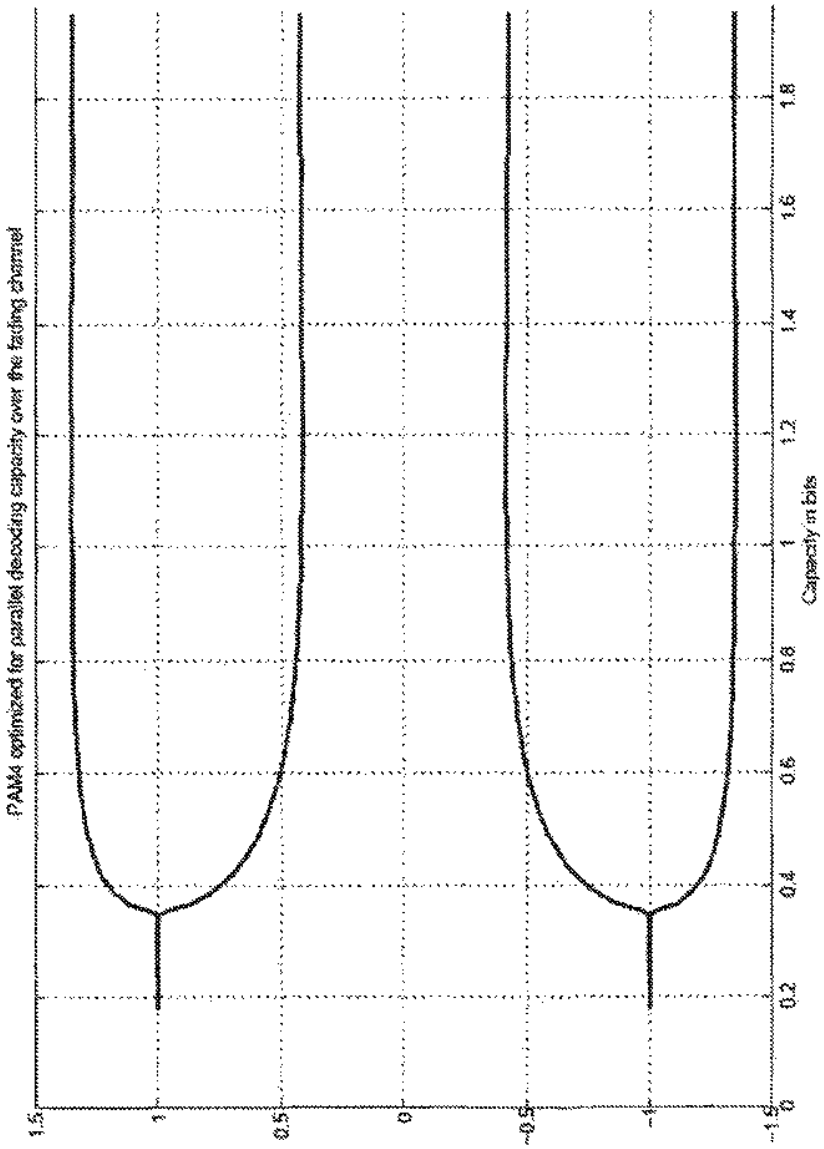

PAM4 optimized for parallel decoding capacity over the fading channel
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 22a

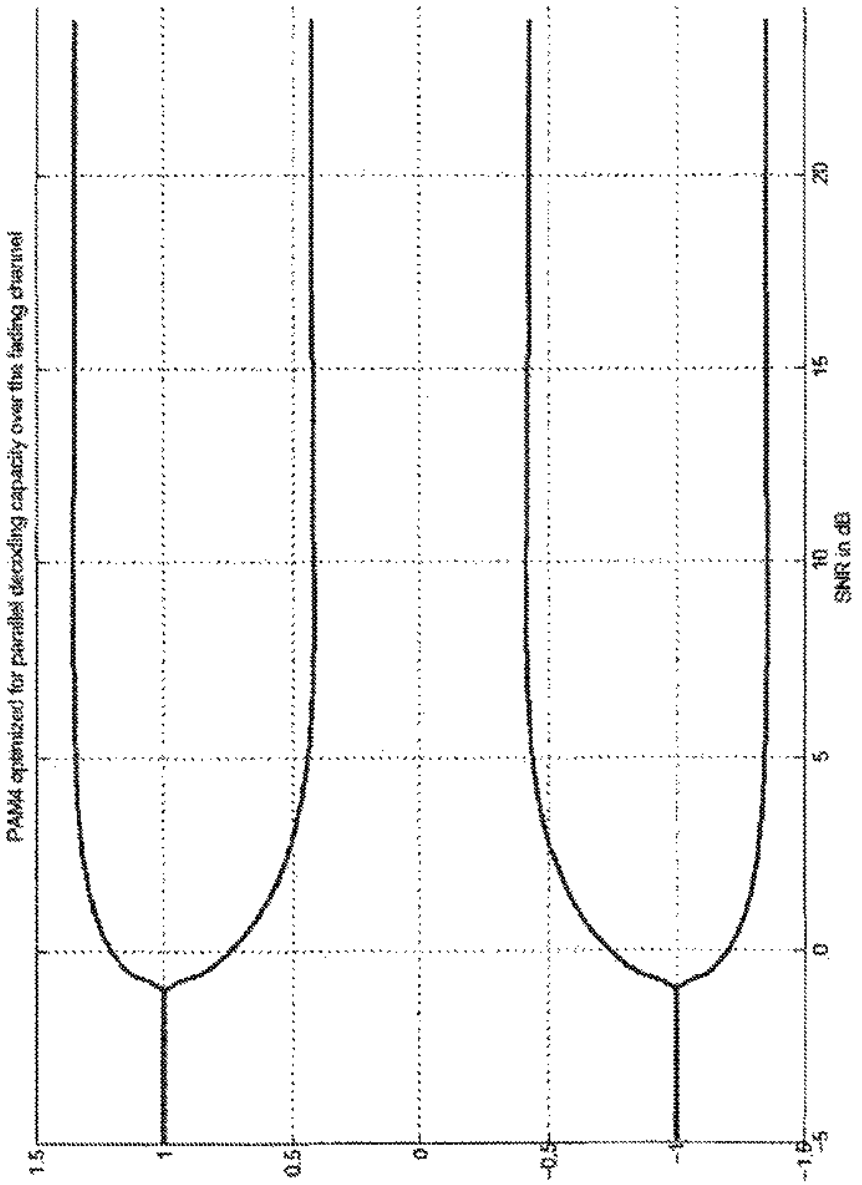

PAM4 optimized for parallel decoding capacity over the fading channel
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20
SNR in dB

FIG. 22b

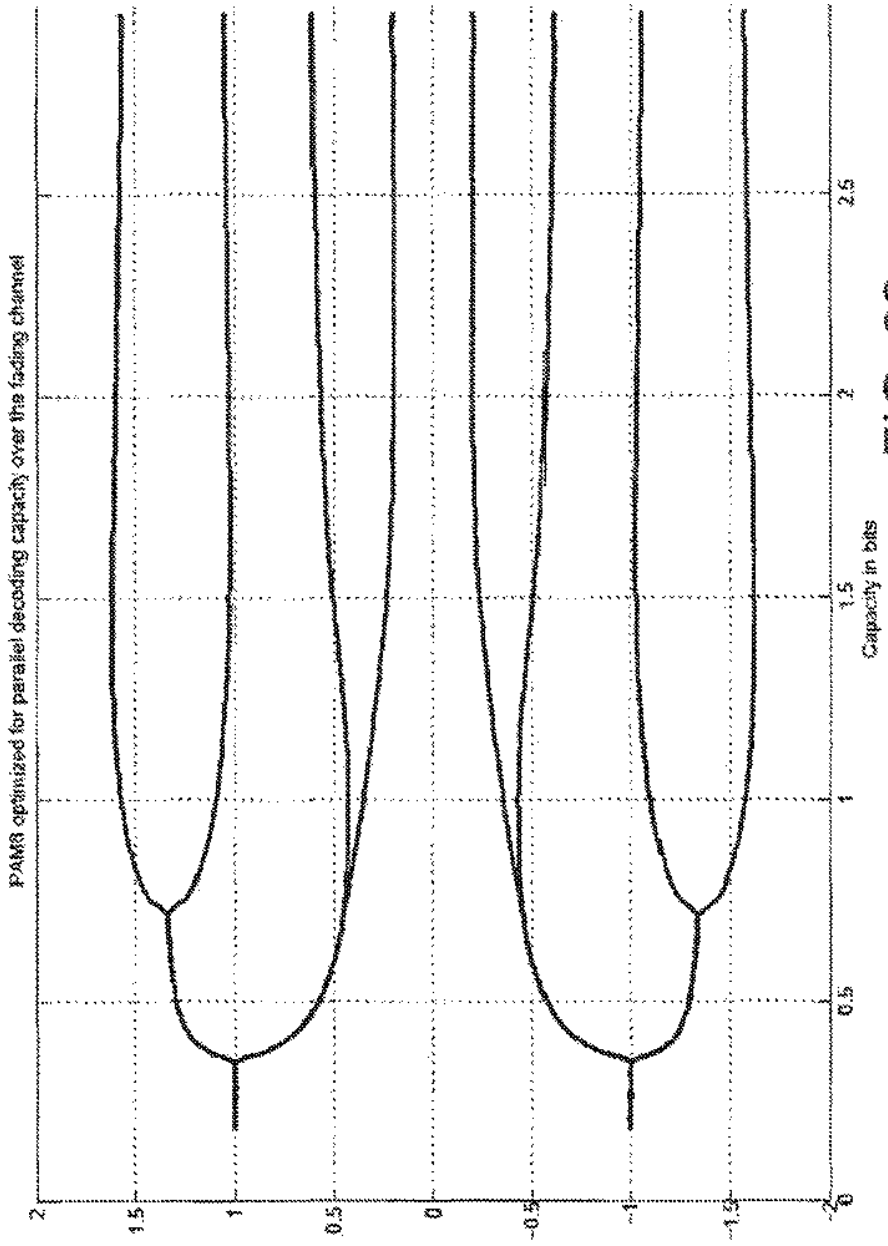

PAM8 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 23a

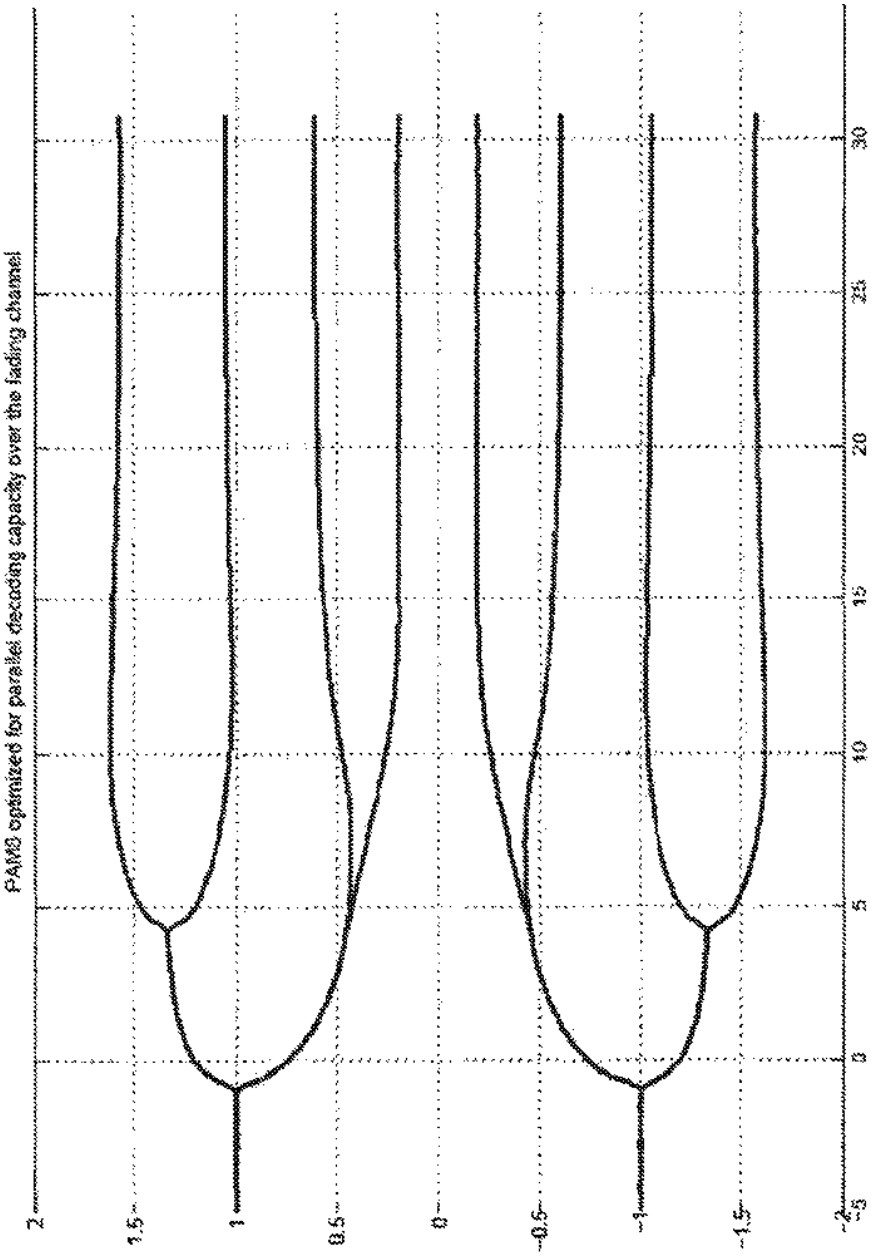

PAM8 optimized for parallel decoding capacity over the fading channel
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30
SNR in dB

FIG. 23b

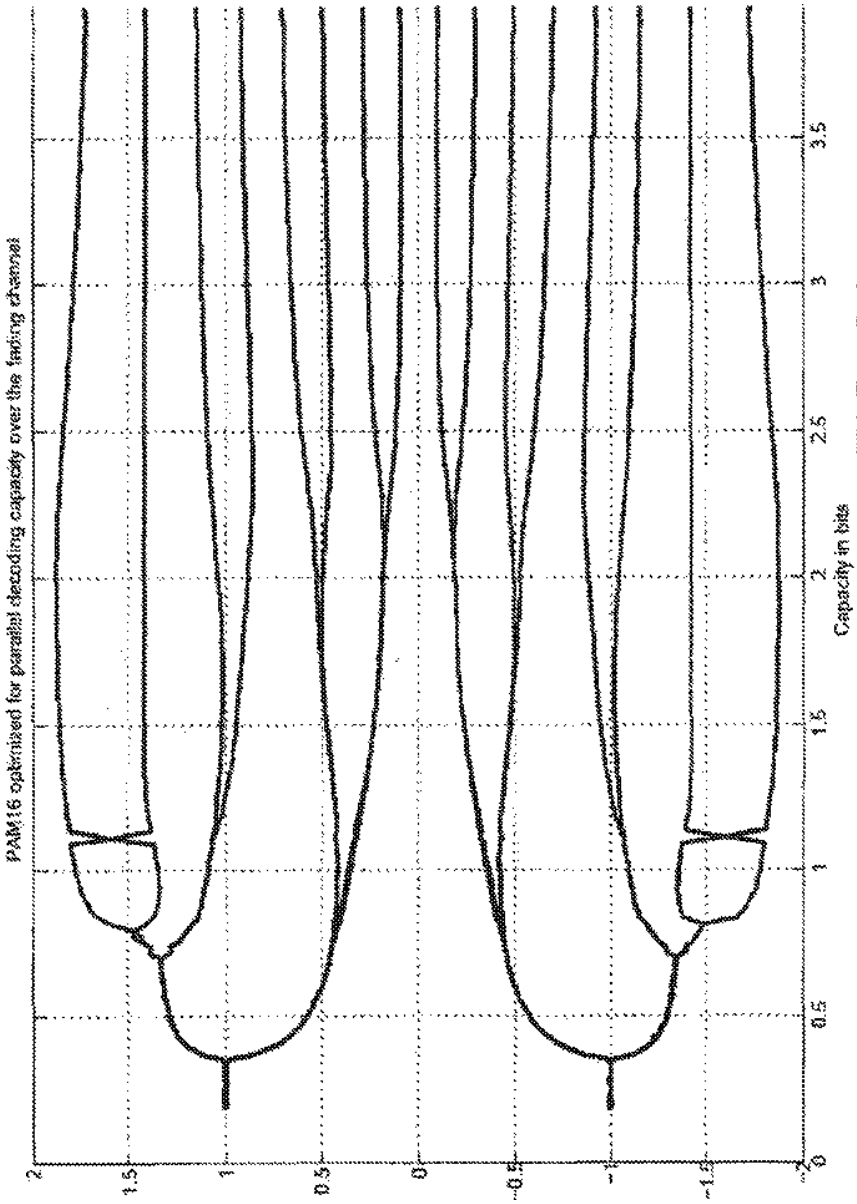

PAM16 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

FIG. 24a

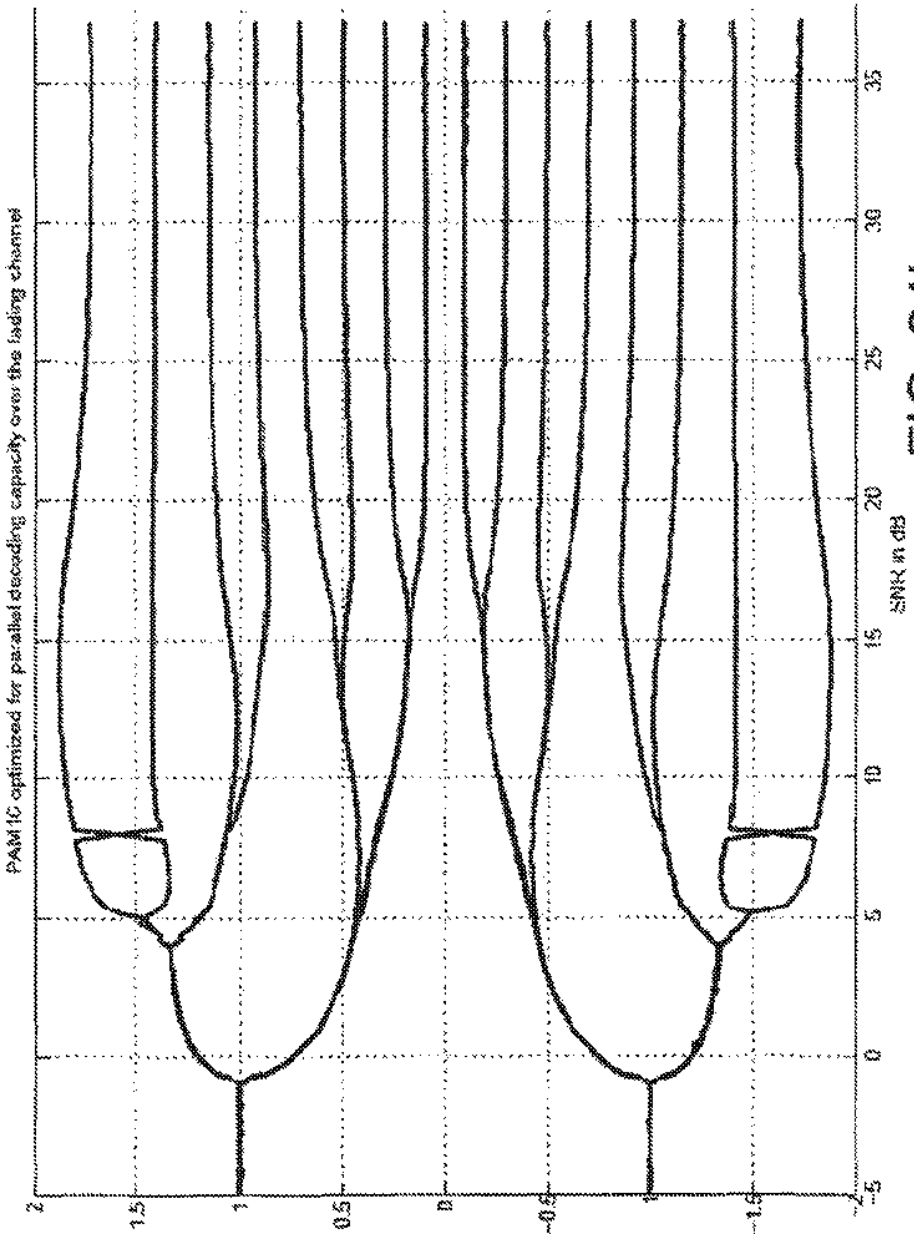

PAM10 optimized for parallel decoding capacity over the fading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
35
30
25
20
15
10
5
0
-5

FIG. 24b

EUROPEAN SEARCH REPORT

Application Number
EP 18 21 2672

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (IPC)
A	RICCARDO DE GAUDENZI ET AL: "Performance analysis of turbo-coded APSK modulations over Nonlinear Satellite Channels", IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS, IEEE SERVICE CENTER, PISCATAWAY, NJ, US. vol. 5, no. 9, 6 September 2006 (2006-09-06), pages 2396-2407, XP002457811, ISSN: 1536-1276, DOI: 10.1109/TWC.2006.1687763 * Sections II-III, VI.A *	1-24	INV. H04L27/34 H04L1/00
A	DIRK SOMMER ET AL: "Signal Shaping by Non-Uniform QAM for AWGN Channels and Applications Using Turbo Coding", PROC. ITG CONF. SOURCE AND CHANNEL CODING, 31 January 2000 (2000-01-31), pages 81-86, XP055110228, * Sections I-IV *	1-24	
A	RICCARDO DE GAUDENZI ET AL: "Turbo-coded APSK modulations design for satellite broadband communications", INTERNATIONAL JOURNAL OF SATELLITE COMMUNICATIONS AND NETWOR, JOHN WILEY & SONS, CHICHESTER, GB, vol. 24, no. 4, 1 July 2006 (2006-07-01), pages 261-281, XP008134566, ISSN: 1542-0973, DOI: 10.1002/SAT.841 [retrieved on 2006-06-19] * Sections 2-4 *	1-24	TECHNICAL FIELDS SEARCHED (IPC) H04L

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
The Hague	27 May 2019	Devillers, Bertrand

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

EUROPEAN SEARCH REPORT

Application Number
EP 18 21 2672

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (IPC)
A	FEI ZESONG ET AL: "Shaping gain by non-uniform qam constellation i with binary turbo coded modulation", PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS, 2003. PIMRC 2003. 14TH IEEE PROCEEDINGS ON SEPT. 7-10, 2003, IEEE, PISCATAWAY, NJ, USA, vol. 2, 7 September 2003 (2003-09-07), pages 1863-1867, XP010678797, ISBN: 978-0-7803-7822-3 * Sections II-III; figures 1-3 *	1-24	
X,P	NAGED F BARSOUN ET AL: "Constellation Design via Capacity Maximization", INFORMATION THEORY, 2007. ISIT 2007. IEEE INTERNATIONAL SYMPOSIUM ON, IEEE, PISCATAWAY, NJ, USA, 24 June 2007 (2007-06-24), pages 1821-1825, XP031460978, ISBN: 978-1-4244-1397-3 * the whole document *	1-24	TECHNICAL FIELDS SEARCHED (IPC)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
The Hague	27 May 2019	Devillers, Bertrand

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

PATENT COOPERATION TREATY

RECEIVED
JUL 14 2009

From the
INTERNATIONAL PRELIMINARY EXAMINING AUTHORITY

To:
DAVID J. BAILEY
KAUTH, POMERUY, PECK, & BAILEY LLP
P.O.BOX 19152
IRVINE, CA 92623

PCT

NOTIFICATION OF TRANSMITTAL OF
INTERNATIONAL PRELIMINARY
REPORT ON PATENTABILITY
(Chapter II of the Patent Cooperation Treaty)

(PCT Rule 71.1)

Date of mailing (*day/month/year*)	30 JUN 2009
Applicant's or agent's file reference: 13:00999.PCT	**IMPORTANT NOTIFICATION**

International application No.	International filing date (*day/month/year*)	Priority date (*day/month/year*)
PCT/US08/65994	05 June 2008 (05.06.2008)	05 June 2007 (05.06.2007)

Applicant

BARSOUM, MACED F.

1. The applicant is hereby notified that this International Preliminary Examining Authority transmits herewith the international preliminary report on patentability and its annexes, if any, established on the international application.

2. A copy of the report and its annexes, if any, is being transmitted to the International Bureau for communication to all the elected Offices.

3. Where required by any of the elected Offices, the International Bureau will prepare an English translation of the report (but not of any annexes) and will transmit such translation to those Offices.

4. **REMINDER**

The applicant must enter the national phase before each elected Office by performing certain acts (filing translations and paying national fees) within 30 months from the priority date (or later in some Offices)(Article 39(1))(see also the reminder sent by the International Bureau with Form PCT/IB/301).

Where a translation of the international application must be furnished to an elected Office, that translation must contain a translation of any annexes to the international preliminary report on patentability. It is the applicant's responsibility to prepare and furnish such translation directly to each elected Office concerned.

For further details on the applicable time limits and requirements of the elected Offices, see Volume II of the *PCT Applicant's Guide*.

The applicant's attention is drawn to Article 33(5), which provides that the criteria of novelty, inventive step and industrial applicability described in Article 33(2) to (4) merely serve the purposes of international preliminary examination and that "any Contracting State may apply additional or different criteria for the purposes of deciding whether, in that State, the claimed invention is patentable or not" (see also Article 27(5)). Such additional criteria may relate, for example, to exemptions from patentability, requirements for enabling disclosure, clarity and support for the claims.

Name and mailing address of the IPEA/ US Mail Stop PCT, Attn: IPEA/US Commissioner for Patents P.O. Box 1450 Alexandria, Virginia 22313-1450 Facsimile No. (571) 273-3201	Authorized officer: Ted Wang Telephone No. (571) 272-3053

Form PCT/IPEA/416 (January 2004)

CASE # _____ ACTION _____
REMINDER _____ DUE DATE _____
DEADLINE _____

PATENT COOPERATION TREATY

PCT

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

(Chapter II of the Patent Cooperation Treaty)

(PCT Article 36 and Rule 70)

Applicant's or agent's file reference 13-00999.PCT	FOR FURTHER ACTION	See Form PCT/IPEA/416
International application No. PCT/US08/65994	International filing date *(day/month/year)* 05 June 2008 (05.06.2008)	Priority date *(day/month/year)* 05 June 2007 (05.06.2007)

International Patent Classification (IPC) or national classification and IPC

IPC: **H04J 11/00**(2006.01)
USPC: 370/207

Applicant

BARSOUM, MAGED F.

1. This report is the international preliminary examination report, established by this International Preliminary Examining Authority under Article 35 and transmitted to the applicant according to Article 36.
2. This REPORT consists of a total o 4 sheets, including this cover sheet.
3. This report is also accompanied by ANNEXES, comprising:
   a. ☐ *(sent to the applicant and to the International Bureau)* a total of ___ sheets, as follows:
      ☐ sheets of the description, claims and/or drawings which have been amended and are the basis of this report and/or sheets containing rectifications authorized by this Authority (see Rule 70.16 and Section 607 of the Administrative Instructions).
      ☐ sheets which supersede earlier sheets, but which this Authority considers contain an amendment that goes beyond the disclosure in the international application as filed, as indicated in item 4 of Box No. I and the Supplemental Box.
   b. ☐ *(sent to the International Bureau only)* a total of (indicate type and number of electronic carrier(s)) _____ , containing a sequence listing and/or tables related thereto, in electronic form only, as indicated in the Supplemental Box Relating to Sequence Listing (see Section 802 of the Administrative Instructions).
4. This report contains indications relating to the following items:
   - ☒ Box No. I Basis of the report
   - ☐ Box No. II Priority
   - ☐ Box No. III Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
   - ☐ Box No. IV Lack of unity of invention
   - ☒ Box No. V Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement
   - ☐ Box No. VI Certain documents cited
   - ☐ Box No. VII Certain defects in the international application
   - ☐ Box No. VIII Certain observations on the international application

Date of submission of the demand	Date of completion of this report
21 January 2009 (21.01.2009)	18 June 2009 (18.06.2009)
Name and mailing address of the IPEA/ US Mail Stop PCT, Attn: IPEA/US Commissioner for Patents P.O. Box 1450 Alexandria, Virginia 22313-1450 Facsimile No. (571) 273-3201	Authorized officer Ted Wang Telephone No. (571) 272-3053

Form PCT/IPEA/409 (cover sheet) (April 2007)

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

International application No.
PCT/US98/65994

Box No. I Basis of the report

1. With regard to the **language**, this report is based on:

☒ the international application in the language in which it was filed.

☐ a translation of the international application into English, which is the language of a translation furnished for the purposes of:

- ☐ international search (under Rules 12.3(a) and 23.1(b))
- ☐ publication of the international application (under Rule 12.4(a))
- ☐ international preliminary examination (under Rules 55.2(a) and/or 55.3(a))

2. With regard to the **elements** of the international application, this report is based on *(replacement sheets which have been furnished to the receiving Office in response to an invitation under Article 14 are referred to in this report as "originally filed" and are not annexed to this report)*:

☐ the international application as originally filed/furnished

☒ the description:
pages 1-19 as originally filed/furnished
pages* NONE received by this Authority on _____
pages* NONE received by this Authority on _____

☒ the claims:
pages 20,21 and 23 as originally filed/furnished
pages* 22 as amended (together with any statement) under Article 19
pages* NONE received by this Authority on _____
pages* NONE received by this Authority on _____

☒ the drawings:
pages 1-43 as originally filed/furnished
pages* NONE received by this Authority on _____
pages* NONE received by this Authority on _____

☐ a sequence listing and/or any related table(s) - see Supplemental Box Relating to Sequence Listing.

3. ☐ The amendments have resulted in the cancellation of:

- ☐ the description, pages _____
- ☐ the claims, Nos. _____
- ☐ the drawings, sheets/figs _____
- ☐ the sequence listing (*specify*): _____
- ☐ any table(s) related to the sequence listing (*specify*): _____

4. ☐ This report has been established as if (some of) the amendments annexed to this report and listed below had not been made, since they have been considered to go beyond the disclosure as filed, as indicated in the Supplemental Box (Rule 70.2(c)).

- ☐ the description, pages _____
- ☐ the claims, Nos. _____
- ☐ the drawings, sheets/figs _____
- ☐ the sequence listing (*specify*): _____
- ☐ any table(s) related to the sequence listing (*specify*): _____

5. ☐ This report has been established taking into account the **rectification of an obvious mistake** authorized by or notified to this Authority under Rule 91 (Rule 70.2(e))

** If item 4 applies, some or all of those sheets may be marked "superseded."*

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

International application No.
PCT/US02/65994

Box No. V Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement

1. Statement

Novelty (N)	Claims 1-12	YES
	Claims NONE	NO
Inventive Step (IS)	Claims 1-12	YES
	Claims NONE	NO
Industrial Applicability (IA)	Claims 1-12	YES
	Claims NONE	NO

2. Citations and Explanations (Rule 70.7)

Claims 1-12 the criteria set out in PCT Article 33(2)-(3), because the prior art does not teach or fairly suggest the limitation of "wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}" as recited in claim 1 and "mapping the encoded bits of user information to a symbol constellation, wherein the symbol constellation is a capacity optimized geometrically spaced symbol constellation that provides a given capacity at a reduced signal-to-noise ratio compared to a signal constellation that maximizes d_{min}" as recited in claim 8. "selecting the points of a geometrically shaped symbol constellation, and iteratively optimizing the location of the points of the constellation to maximize a capacity measure until a predetermined improvement in the SNR performance of the constellation relative to a constellation that maximizes d_{min} has been achieved. " as recited in claims 9-10, "A geometrically shaped symbol constellation, selecting an appropriate constellation size and a desired SNR; and optimizing the location of the points of the constellation to maximize a capacity measure of the constellation." as recited in claim 11 and "transmitting information over a channel using a geometrically shaped symbol modifying the location of points within the geometrically shaped symbol constellation to change the target user data rate " as recited in claim 12.

Claims 1-12 meet the criteria set out in PCT Article 33(4), and thus have industrial applicability because the subject matter claimed can be made or used in industry.

---------- NEW CITATIONS ----------

US 7,212,582
US 6,084,915
US 2007/0025283

Form PCT/IPEA/409 (Box No. V) (April 2007)

PATENT COOPERATION TREATY

RECEIVED
JAN 19 2011
By

From the
INTERNATIONAL PRELIMINARY EXAMINING AUTHORITY

To:
DAVID J. BAILEY
KAUTH, POMEROY, PECK & BAILEY LLP
2875 MICHELLE DRIVE
SUITE 110
IRVINE, CA 92606

PCT

NOTIFICATION OF TRANSMITTAL OF
INTERNATIONAL PRELIMINARY
REPORT ON PATENTABILITY
(Chapter II of the Patent Cooperation Treaty)

(PCT Rule 71.1)

	Date of mailing (*day/month/year*) 14 JAN 2011
Applicant's or agent's file reference J3:01615.PCT	**IMPORTANT NOTIFICATION**

International application No.	International filing date (*day/month/year*)	Priority date (*day/month/year*)
PCT/US09/69861	30 December 2009 (30.12.2009)	30 December 2008 (30.12.2008)

Applicant

CONSTELLATION DESIGNS, INC.

1. The applicant is hereby notified that this International Preliminary Examining Authority transmits herewith the international preliminary report on patentability and its annexes, if any, established on the international application.

2. A copy of the report and its annexes, if any, is being transmitted to the International Bureau for communication to all the elected Offices.

3. Where required by any of the elected Offices, the International Bureau will prepare an English translation of the report (but not of any annexes) and will transmit such translation to those Offices.

4. **REMINDER**

The applicant must enter the national phase before each elected Office by performing certain acts (filing translations and paying national fees) within 30 months from the priority date (or later in some Offices)(Article 39(1))(see also the reminder sent by the International Bureau with Form PCT/IB/301).

Where a translation of the international application must be furnished to an elected Office, that translation must contain a translation of any annexes to the international preliminary report on patentability. It is the applicant's responsibility to prepare and furnish such translation directly to each elected Office concerned.

For further details on the applicable time limits and requirements of the elected Offices, see Volume II of the *PCT Applicant's Guide*.

The applicant's attention is drawn to Article 33(5), which provides that the criteria of novelty, inventive step and industrial applicability described in Article 33(2) to (4) merely serve the purposes of international preliminary examination and that "any Contracting State may apply additional or different criteria for the purposes of deciding whether, in that State, the claimed invention is patentable or not" (see also Article 27(5)). Such additional criteria may relate, for example, to exemptions from patentability, requirements for enabling disclosure, clarity and support for the claims.

Name and mailing address of the IPEA/ US	Authorized officer
Mail Stop PCT, Attn: IPEA/US Commissioner for Patents P.O. Box 1450 Alexandria, Virginia 22313-1450	Ken Wieder
Facsimile No.	Telephone No. 571-272-2986

Form PCT/IPEA/416 (January 2004)

PATENT COOPERATION TREATY

PCT

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

(Chapter II of the Patent Cooperation Treaty)

(PCT Article 36 and Rule 70)

Applicant's or agent's file reference	FOR FURTHER ACTION	See Form PCT/IPEA/416
J3:01615.PCT		
International application No.	International filing date *(day/month/year)*	Priority date *(day/month/year)*
PCT/US09/69881	30 December 2009 (30.12.2009)	30 December 2008 (30.12.2008)

International Patent Classification (IPC) or national classification and IPC

IPC: **H04B 7/185**(2006.01)
USPC: 455/12.1

Applicant

CONSTELLATION DESIGNS, INC.

1. This report is the international preliminary examination report, established by this International Preliminary Examining Authority under Article 35 and transmitted to the applicant according to Article 36.
2. This REPORT consists of a total of 10 sheets, including this cover sheet.
3. This report is also accompanied by ANNEXES, comprising:
   a. ☐ *(sent to the applicant and to the International Bureau)* a total of ____ sheets, as follows:
      ☐ sheets of the description, claims and/or drawings which have been amended and are the basis of this report and/or sheets containing rectifications authorized by this Authority (see Rule 70.16 and Section 607 of the Administrative Instructions).
      ☐ sheets which supersede earlier sheets, but which this Authority considers contain an amendment that goes beyond the disclosure in the international application as filed, as indicated in item 4 of Box No. I and the Supplemental Box.
   b. ☐ *(sent to the International Bureau only)* a total of (indicate type and number of electronic carrier(s)) _____ , containing a sequence listing and/or tables related thereto, in electronic form only, as indicated in the Supplemental Box Relating to Sequence Listing (see Section 802 of the Administrative Instructions).
4. This report contains indications relating to the following items:
   - ☒ Box No. I Basis of the report
   - ☐ Box No. II Priority
   - ☐ Box No. III Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
   - ☐ Box No. IV Lack of unity of invention
   - ☒ Box No. V Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement
   - ☐ Box No. VI Certain documents cited
   - ☒ Box No. VII Certain defects in the international application
   - ☐ Box No. VIII Certain observations on the international application

Date of submission of the demand	Date of completion of this report
01 November 2010 (01.11.2010)	07 January 2011 (07.01.2011)
Name and mailing address of the IPEA/ US Mail Stop PCT, Attn: IPEA/US Commissioner for Patents P.O. Box 1450 Alexandria, Virginia 22313-1450 Facsimile No.	Authorized officer Ken Wieder Telephone No. 571-272-2986

Form PCT/IPEA/409 (cover sheet)(April 2007)

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

International application No.
PCT/US09/69881

Box No. I Basis of the report

1. With regard to the **language**, this report is based on:

☒ the international application in the language in which it was filed.

☐ a translation of the international application into English, which is the language of a translation furnished for the purposes of:

☐ international search (under Rules 12.3(a) and 23.1(b))

☐ publication of the international application (under Rule 12.4(a))

☐ international preliminary examination (under Rules 55.2(a) and/or 55.3(a))

2. With regard to the **elements** of the international application, this report is based on *(replacement sheets which have been furnished to the receiving Office in response to an invitation under Article 14 are referred to in this report as "originally filed" and are not annexed to this report)*:

☒ the international application as originally filed/furnished

☒ the description:
pages 1-30 as originally filed/furnished
pages* NONE received by this Authority on
pages* NONE received by this Authority on

☒ the claims:
pages 31-40 as originally filed/furnished
pages* NONE as amended (together with any statement) under Article 19
pages* NONE received by this Authority on
pages* NONE received by this Authority on

☒ the drawings:
pages 1-167 as originally filed/furnished
pages* NONE received by this Authority on
pages* NONE received by this Authority on

☐ a sequence listing and/or any related table(s) - see Supplemental Box Relating to Sequence Listing.

3. ☒ The amendments have resulted in the cancellation of:

☒ the description, pages NONE
☒ the claims, Nos. NONE
☒ the drawings, sheets/figs NONE
☒ the sequence listing (*specify*): NONE
☒ any table(s) related to the sequence listing (*specify*): NONE

4. ☐ This report has been established as if (some of) the amendments annexed to this report and listed below had not been made, since they have been considered to go beyond the disclosure as filed, as indicated in the Supplemental Box (Rule 70.2(c)).

☐ the description, pages
☐ the claims, Nos.
☐ the drawings, sheets/figs
☐ the sequence listing (*specify*):
☐ any table(s) related to the sequence listing (*specify*):

5. ☐ This report has been established taking into account the **rectification of an obvious mistake** authorized by or notified to this Authority under Rule 91 (Rule 70.2(e)).

** If item 4 applies, some or all of those sheets may be marked "superseded."*

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

International application No.
PCT/US09/69881

Box No. V Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement

1. Statement

Novelty (N)	Claims 4-15, 19-28, 30-31, 35-46	YES
	Claims 1-3, 16-18, 32-34, 47	NO
Inventive Step (IS)	Claims NONE	YES
	Claims 1-28 and 30-47	NO
Industrial Applicability (IA)	Claims 1-28 and 30-47	YES
	Claims NONE	NO

2. Citations and Explanations (Rule 70.7)
Please See Continuation Sheet

Form PCT/IPEA/409 (Box No. V) (April 2007)

INTERNATIONAL PRELIMINARY REPORT ON PATENTABILITY

International application No.
PCT/US09/69881

Box No. VII Certain defects in the international application

The following defects in the form or contents of the international application have been noted:

Claim 29 is missing.

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.

Continuation of:

V. 2. Citations and Explanations:
Claims 1-3, 16-18, 32-34, and 47 lack novelty under PCT Article 33(2) as being anticipated by US 2004/0258177 A1 to Shen et al. (hereinafter 'Shen').

Regarding claim 1, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (Information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-8 symbol constellation and at least one of the

Supplemental Box

constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 25 - 46 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0318]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 2, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

Regarding claim 3, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems, abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

Regarding claim 16, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (Information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-16 symbol constellation and at least one of the constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 47 - 84 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0318]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 17, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

Regarding claim 18, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems, abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

Regarding claim 32, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (Information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the

Supplemental Box

channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-32 symbol constellation and at least one of the constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 85 - 148 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0318]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 33, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

Regarding claim 34, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems, abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

Regarding claim 47, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (Information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]. The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a N-Dimensional symbol constellation, where the constellation points in at least one dimension are within at least one of the ranges specified in FIGS. 25 - 167 (a 4-D (4-dimensional) QPSK (Quadrature Phase Shift Key) constellation, para [0058]; m-D (m-dimensional) Gray code map construction for 2m-D M PSK modulation, para [0061]; an m-D (m-dimensional) Gray code map (shown where m=2) constructed using two 1-D (1-dimensional) 3 bit basic Gray code maps (shown using 8 PSK shaped constellations) according to the invention, para [0065]).

Claims 4-15, 19-28, 30-31, and 35-46 lack an inventive step under PCT Article 33(3) as being obvious over Shen, in view of US 2007/0280147 A1 to Catreux-Erceg et al. (hereinafter 'Catreux-Erceg').

Regarding claims 4, 19, and 35, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (Such diversity gains improve system performance by increasing received signal-to-noise ratio, . . . the significant increase in system capacity . . . increase system capacity by nearly M additional bits/cycle for each 3-dB increase in SNR, para [0014]; efficient optimization of network capacity, para [0004]; The spacing of the constellation changes constantly due to transmission power control (TPC) . . . This makes an accurate estimation of the constellation very difficult, para [0012]; The gain values 310a and 312a of CLMI using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.65 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 5, 20, and 36, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at

Supplemental Box

least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (channel gain as seen at finger i at the receive antenna 206 of UE 202a may be represented as h.sub.i1+w.sub.2h.sub.2i, para [0051]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 6, 21, and 37, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 7, 22, and 38, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain values 310a and 312a of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.65 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 8, 23, and 39, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 9, 24, and 40, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-

Supplemental Box

Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 10, 25, and 41, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; For example, the weight factors W.sub.1 and W.sub.2 may correspond to phase and/or amplitude component feedback adjustments that may be generated by the receiver based on the type of space-time coding, para [0040]; Such diversity gains improve system performance by increasing received signal-to-noise ratio. . . . the significant increase in system capacity . . . increase system capacity by nearly M additional bits/cycle for each 3-dB increase in SNR, para [0014]; The gain values 310a and 312a of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.65 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 11, 26, and 42, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; phase and/or amplitude component feedback adjustments that may be generated by the receiver based on the type of space-time coding, para [0040]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 12, 27, and 43, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Supplemental Box

Regarding claims 13, 28, and 44, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 14, 30, and 45, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 15, 31, and 46, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Claims 1-28 and 30-47 have industrial applicability as defined by PCT Article 33(4) because the subject matter can be made or used in industry.

---------- NEW CITATIONS ----------

PATENT COOPERATION TREATY

PCT

INTERNATIONAL SEARCH REPORT

(PCT Article 18 and Rules 43 and 44)

Applicant's or agent's file reference J3:01615.PCT	FOR FURTHER ACTION	see Form PCT/ISA/220 as well as, where applicable, item 5 below.
International application No. PCT/US 08/89881	International filing date *(day/month/year)* 30 December 2008 (30.12.2008)	(Earliest) Priority Date *(day/month/year)* 30 December 2008 (30.12.2008)
Applicant CONSTELLATION DESIGNS, INC.		

This international search report has been prepared by this International Searching Authority and is transmitted to the applicant according to Article 18. A copy is being transmitted to the International Bureau.

This international search report consists of a total of 7 sheets.

☐ It is also accompanied by a copy of each prior art document cited in this report.

1. **Basis of the report**

a. With regard to the language, the international search was carried out on the basis of:

☒ the international application in the language in which it was filed.

☐ a translation of the international application into _____ which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b)).

b. ☐ This international search report has been established taking into account the rectification of an obvious mistake authorized by or notified to this Authority under Rule 91 (Rule 43.6*bis*(a)).

c. ☐ With regard to any nucleotide and/or amino acid sequence disclosed in the international application, see Box No. I.

2. ☐ Certain claims were found unsearchable (see Box No. II).

3. ☐ Unity of invention is lacking (see Box No. III).

4. With regard to the title,

☒ the text is approved as submitted by the applicant.

☐ the text has been established by this Authority to read as follows:

5. With regard to the abstract,

☒ the text is approved as submitted by the applicant.

☐ the text has been established, according to Rule 38.2, by this Authority as it appears in Box No. IV. The applicant may, within one month from the date of mailing of this international search report, submit comments to this Authority.

6. With regard to the drawings,

a. the figure of the drawings to be published with the abstract is Figure No. 1

☒ as suggested by the applicant.

☐ as selected by this Authority, because the applicant failed to suggest a figure.

☐ as selected by this Authority, because this figure better characterizes the invention.

b. ☐ none of the figures is to be published with the abstract.

Form PCT/ISA/210 (first sheet) (July 2009)

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US 09/69661

A. CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H04B 7/185 (2010.01)
USPC - 455/12.1
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC(8): H04B 7/185 (2010.01)
USPC: 455/12.1

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
IPC(8): H04B 7/185 (2010.01) (keyword limited; terms below)
USPC: 455/7, 12.1 (keyword limited; terms below)

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
PubWEST(PGPB, USPT, EPAB, JPAB); Google; Search Terms: signal constellationor diagram point transmitter transceiver receiver channel communicat network order coding encoder encoding bit byte mapping mapper maps map modulat signal$ transmission symbol bits demodulat de-modulat estimat predict likelihood likely probabilit PAM pulse amplitude modulat

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X --- Y	US 2004/0258177 A1 (SHEN et al.) 23 December 2004 (23.12.2004), entire document, especially abstract, para. [0009], [0016], [0019], [0059], [0061], [0065], [0075], [0149], [0155], [0157], [0174], [0318], and [0330].	1-3, 16-18, 32-34, 47 --- 4-15, 19-26, 30-31, 35-46
Y	US 2007/0280147 A1 (CATREUX-ERCEG et al.) 06 December 2007 (06.12.2007), entire document, especially para. [0004], [0012], [0014], [0030], [0040], [0051], [0062], and [0064]-[0065].	4-15, 19-26, 30-31, 35-46
A	US 2007/0054614 A1 (WALKER et al.) 08 March 2007 (08.03.2007), entire document	1-26, 30-47
A	US 2005/0286409 A1 (YOON et al.) 29 December 2005 (29.12.2005), entire document	1-26, 30-47

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
12 April 2010 (12.04.2010)	03 MAY 2010
Name and mailing address of the ISA/US Mail Stop PCT, Attn: ISA/US, Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Authorized officer: Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/210 (second sheet) (July 2009)

RECEIVED
OCT 24 2008
By

PATENT COOPERATION TREATY

PCT

INTERNATIONAL SEARCH REPORT

(PCT Article 18 and Rules 43 and 44)

Applicant's or agent's file reference J3:00999.PCT	FOR FURTHER ACTION	see Form PCT/ISA/220 as well as, where applicable, item 5 below.
International application No. PCT/US2008/065994	International filing date *(day/month/year)* 05 June 2008	(Earliest) Priority Date *(day/month/year)* 05 June 2007
Applicant BARSOUM, MAGED F.		

This international search report has been prepared by this International Searching Authority and is transmitted to the applicant according to Article 18. A copy is being transmitted to the International Bureau.

This international search report consists of a total of 2 sheets.

☐ It is also accompanied by a copy of each prior art document cited in this report.

1. **Basis of the report**

   a. With regard to the language, the international search was carried out on the basis of:

   ☒ the international application in the language in which it was filed

   ☐ a translation of the international application into _____, which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b))

   b ☐ With regard to any nucleotide and/or amino acid sequence disclosed in the international application, see Box No. I.

2. ☐ Certain claims were found unsearchable (see Box No. II)

3. ☐ Unity of invention is lacking (see Box No. III)

4. With regard to the title,

   ☒ the text is approved as submitted by the applicant

   ☐ the text has been established by this Authority to read as follows:

5. With regard to the abstract,

   ☐ the text is approved as submitted by the applicant

   ☒ the text has been established, according to Rule 38.2(b), by this Authority as it appears in Box No. IV. The applicant may, within one month from the date of mailing of this international search report, submit comments to this Authority

6. With regard to the drawings,

   a. the figure of the drawings to be published with the abstract is Figure No. 1

   ☒ as suggested by the applicant

   ☐ as selected by this Authority, because the applicant failed to suggest a figure

   ☐ as selected by this Authority, because this figure better characterizes the invention

   b. ☐ none of the figures is to be published with the abstract

Form PCT/ISA/210 (first sheet) (April 2005)

CASE # J3:00999.PCT ACTION _____
REMINDER _____ DUE DATE _____
DEADLINE _____

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US2008/065994

A. CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H04J11/00 (2008.04)
USPC - 370/207
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC(8) - H04J 11/00 (2008.04)
USPC - 370/207

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
PatBase, Google Patents

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X ..... Y	US 2004/0022179 A1 (GIANNAKIS et al) 05 February 2004 (05.02.2004) entire document	9-11 ..... 1-8, 12
Y	US 2007/0116161 A1 (TOKORO et al) 24 May 2007 (24.05.2007) entire document	1-8, 12
Y	US 6,603,801 B1 (ANDREN et al) 05 August 2003 (05.08.2003) entire document	4
Y	US 2006/0155843 A1 (GLASS et al) 13 July 2006 (13.07.2006) entire document	3

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
03 October 2008	22 OCT 2008
Name and mailing address of the ISA/US Mail Stop PCT, Attn: ISA/US, Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Authorized officer: Blaine R. Copenheaver PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/210 (second sheet) (April 2005)

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US2008/065994

Box No. I Basis of this opinion

1 With regard to the language, this opinion has been established on the basis of:

☒ the international application in the language in which it was filed.

☐ a translation of the international application into _____ which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b)).

2. ☐ This opinion has been established taking into account the rectification of an obvious mistake authorized by or notified to this Authority under Rule 91 (Rule 43*bis*.1(a))

3. With regard to any nucleotide and/or amino acid sequence disclosed in the international application, this opinion has been established on the basis of:

a type of material

☐ a sequence listing

☐ table(s) related to the sequence listing

b. format of material

☐ on paper

☐ in electronic form

c. time of filing/furnishing

☐ contained in the international application as filed

☐ filed together with the international application in electronic form

☐ furnished subsequently to this Authority for the purposes of search

4. ☐ In addition, in the case that more than one version or copy of a sequence listing and/or table(s) relating thereto has been filed or furnished, the required statements that the information in the subsequent or additional copies is identical to that in the application as filed or does not go beyond the application as filed, as appropriate, were furnished.

5. Additional comments:

Form PCT/ISA/237 (Box No. I) (April 2007)

WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY

International application No
PCT/US2008/065994

Box No. V Reasoned statement under Rule 43bis.1(a)(i) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement

1. Statement

Novelty (N)	Claims	1-8, 12	YES
	Claims	9-11	NO
Inventive step (IS)	Claims	None	YES
	Claims	1-12	NO
Industrial applicability (IA)	Claims	1-12	YES
	Claims	None	NO

2. Citations and explanations:

Claims 9-11 lack novelty under PCT Article 33(2) as being anticipated by Giannakis et al. (US 2004/0022179 A1).

Regarding claim 9, Giannakis et al. disclose a method of selecting the points of a geometrically shaped symbol constellation used in a communication system, comprising: selecting an appropriate constellation size (para. 0068, constellation size is denoted by Q) and a desired capacity per dimension (para. 0036, per real dimension); estimating an initial SNR at which the system is likely to operate (para 0044, measures SNR); and iteratively optimizing the location of the points (para. 0008, complex field linear precoding with small size precoders to improve diversity gain and, as a result, system performance) of the constellation (para. 0049, constellation) to maximize a capacity measure (para. 0051, maximum coding and diversity gains) until a predetermined improvement in the SNR performance of the constellation relative to a constellation (para. 0074) that maximizes dmin has been achieved (para. 0052, dmin).

Regarding claim 10, Giannakis et al. disclose a geometrically shaped symbol constellation used in a communication system, obtained by a process of: selecting an appropriate constellation size (para. 0068, constellation size is denoted by Q) and a desired capacity per dimension (para. 0036, per real dimension); estimating an initial SNR at which the system is likely to operate (para. 0044, measures SNR); and iteratively optimizing the location of the points (para. 0008, complex field linear precoding with small size precoders to improve diversity gain and, as a result, system performance) of the constellation (para. 0049, constellation) to maximize a capacity measure (para. 0051, maximum coding and diversity gains) until a predetermined improvement in the SNR performance of the constellation relative to a constellation (para. 0074) that maximizes dmin has been achieved (para. 0052, dmin).

Regarding claim 11, Giannakis et al. disclose a geometrically shaped symbol constellation used in a communication system, obtained by a process of: selecting an appropriate constellation size (para. 0049, size constellations) and a desired SNR (para. 0074, run for each SNR); and optimizing the location of the points (para. 0008, complex field linear precoding with small size precoders to improve diversity gain and, as a result, system performance) of the constellation (para. 0049, constellation) to maximize a capacity measure (para. 0051, maximum coding and diversity gains) of the constellation (para. 0074, constellation).

(Cont. in Supplemental Box)

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US2008/065994

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:
Box No. V
2. Citations and explanations.

Claims 1, 2, 5-8 lack an inventive step under PCT Article 33(3) as being obvious over Giannakis et al. in view of Tokoro et al. (US 2007/0116161 A1).

Regarding claim 1, Giannakis et al. disclose a digital communication system, comprising, a transmitter configured to transmit signals to a receiver via a communication channel (para. 0018, telecommunication system in which a transmitter and receiver); wherein the transmitter (para. 0018), comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (para. 0010, non-linear coder encodes a data stream to produce an encoded data stream); a mapper configured to map encoded bits to symbols in a symbol constellation (para. 0026, Mapping unit 16 maps the symbols d to constellation symbols); a modulator configured to generate a signal for transmission via the communication channel (para. 0010, modulator produces an output waveform in accordance with the precoded data stream for transmission through a wireless channel) using symbols generated by the mapper (para. 0026, Mapping unit 16); wherein the receiver (para. 0012, receiver), comprises: a demodulator configured to demodulate the received signal via the communication channel (para. 0011, demodulator receives a waveform carrying a joint coded-linearly precoded transmission and produces a demodulated data stream); a decoder that is configured to estimate decoded bits from the likelihoods (para. 0048, universal lattice decoder (a.k.a. sphere decoder) can approach ML performance in polynomial time); wherein the symbol constellation (para. 0026, constellation symbol) is a capacity optimized (para. 0008, complex field linear precoding with small size precoders to improve diversity gain and, as a result, system performance) that provides a given capacity at a reduced signal-to-noise ratio (para. 0044, savings in SNR) compared to a signal constellation (para. 0074, constellation) that maximizes (para. 0050, maximum) dmin (para. 0052, dmin), but is silent on a demapper configured to estimate likelihoods from the demodulated signal; generated by the demapper; geometrically spaced symbol constellation and, Tokoro et al. in discussing diversity receiver and method disclose a demapper configured to estimate the likelihoods from the demodulated signal (para. 0045, through demapping processing); generated by the demapper (para. 0045, through demapping processing); geometrically spaced symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with the a demapper configured to estimate likelihoods from the demodulated signal; generated by the demapper; geometrically spaced symbol constellation of Tokoro et al. for the purpose of reproducing error-free data.

Regarding claim 2, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1, Giannakis et al. further disclose capacity optimized (para. 0008, system performance) subject to additional constraints (para. 0008, utilize linear precoding in combination with non-linear error-control coding to achieve maximum diversity), but is silent on wherein the geometrically spaced symbol constellation. However, Tokoro et al. disclose wherein the geometrically spaced symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with wherein the geometrically spaced symbol constellation of Tokoro et al. for the purpose of reproducing error-free data.

Regarding claim 5, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1, Giannakis et al. further disclose wherein the channel is an AWGN channel (para. 0035, additive white Gaussian noise (AWGN)).

Regarding claim 6, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1, Giannakis et al. further disclose wherein the channel is a fading channel (para. 0041, fading channel).

Regarding claim 7, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1, Giannakis et al. further disclose wherein are variable (para. 0045, signal space diversity) depending on the user rate (para. 0045, transmission rate), but is silent on the constellations points of the geometrically spaced symbol constellation. However, Tokoro et al. disclose wherein the geometrically spaced symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with wherein the geometrically spaced symbol constellation of Tokoro et al. for the purpose of reproducing error-free data.

Regarding claim 8, Giannakis et al. disclose a method of transmitting bits of user information over a communication channel, comprising: encoding the bits of user information (para. 0010, encodes a data stream) using a coding scheme (para. 0009, coding-precoding scheme); mapping the encoded bits (para. 0010, encodes) of user information to a symbol constellation (para. 0026, Mapping unit 16 maps the symbols d to constellation symbols), wherein the symbol constellation (para. 0026, constellation symbol) is a capacity optimized (para. 0008, complex field linear precoding with small size precoders to improve diversity gain and, as a result, system performance) that provides a given capacity at a reduced signal-to-noise ratio (para. 0044, savings in SNR) compared to a signal constellation (para. 0074, constellation) that maximizes (para. 0050, maximum) dmin (para. 0052, d min) modulating the symbols (para. 0010, modulator) in accordance with a modulation scheme (para. 0026, quadrature amplitude modulation (QAM)); transmitting the modulated signal via the communication channel (para. 0015, modulate the precoded data stream to produce an output waveform; and transmit the output waveform through a wireless channel); receiving (para. 0033, Receiver 6) a modulated signal (para. 0015, modulate the precoded data stream); demodulating the modulated signal (para. 0011, demodulator) in accordance with the modulation scheme (para. 0026, QAM); to produce likelihoods (para. 0048, ML); decoding the likelihoods (para. 0048, ML) to obtain an estimate of the decoded bits (para. 0011, The decoder applies iterative decoding techniques to decode the demodulated data and produce estimated data), but is silent on geometrically spaced symbol constellation; demapping the demodulated signal using the geometrically shaped signal constellation. However, Tokoro et al. disclose geometrically spaced symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged); demapping (para. 0045, through demapping processing) the demodulated signal (para. 0011, demodulated) using the geometrically shaped signal constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with the geometrically spaced symbol constellation; demapping the demodulated signal using the geometrically shaped signal constellation of Tokoro et al. for the purpose of reproducing error-free data.

(Cont. in Next Supplemental Box)

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US2006/065994

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:
Previous Supplemental Box:

Claim 3 lacks an inventive step under PCT Article 33(3) as being obvious over Giannakis et al. in view of Tokoro et al. in further view of Glass et al. (US 2006/0155843 A1).

Regarding claim 3, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1, Giannakis et al. further disclose optimized for capacity (para. 0008, system performance), but is silent on wherein the geometrically shaped symbol constellation, using parallel decode capacity. However, Tokoro et al. disclose wherein the geometrically spaced symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with wherein the geometrically spaced symbol constellation of Tokoro et al. for the purpose of reproducing error-free data. Neither Giannakis et al. and Tokoro et al. discloses parallel decode capacity. However, Glass et al. in discussing information transportation scheme from high functionality probe to logic analyzer, disclose parallel decode capacity (para. 0042, parallel decode). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. in view of Tokoro et al. with the parallel decode capacity of Glass et al. for the purpose of improving transport of information.

Claim 4 lacks an inventive step under PCT Article 33(3) as being obvious over Giannakis et al. in view of Tokoro et al. in further view of Andren et al. (US 6,603,801 B1).

Regarding claim 4, Giannakis et al. in view of Tokoro et al. disclose the communication system of claim 1. Giannakis et al. further disclose optimized (para. 0008, system performance), but is silent on wherein the geometrically shaped symbol constellation for capacity using joint capacity. However, Tokoro et al. disclose wherein the geometrically shaped symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. with wherein the geometrically shaped symbol constellation of Tokoro et al. for the purpose of reproducing error-free data. Neither Giannakis et al. in view of Tokoro et al. discloses capacity using joint capacity. However, Andren et al. in discussing spread spectrum transceiver for use in wireless local area network and having multipath mitigation, disclose capacity using joint capacity (col. 19, ins. 56-65, joint capacity). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Giannakis et al. in view of Tokoro et al. with the capacity using joint capacity of Andren et al. for the purpose of improving throughput.

Claim 12 lacks an inventive step under PCT Article 33(3) as being obvious over Tokoro et al. in view of Giannakis et al.

Regarding claim 12, Tokoro et al. disclose a method of transmitting information, comprising: transmitting information (para. 0045, transmitter) over a channel using a geometrically shaped symbol constellation (para. 0018, constellations that represent respective situations in which symbols are geometrically arranged); and symbol constellation (para. 0018), within the geometrically shaped (para. 0018) but is silent on modifying the location of points to change the target user data rate. However, Giannakis et al. disclose modifying the location of points (para. 0045, Signal space diversity) to change the target user data rate (para. 0045, transmission rate). Therefore, it would have been obvious to one of ordinary skill in the art at the time of the invention to incorporate the invention of Tokoro et al. with the modifying the location of points to change the target user data rate of Giannakis et al. for the purpose of providing multiplicative benefit to the data rate.

Claims 1-12 meet the criteria set out in PCT Article 33(4), and thus have industrial applicability because the subject matter claimed can be made or used in industry.

WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/68561

Box No. I Basis of this opinion

1. With regard to the language, this opinion has been established on the basis of:

   ☒ the international application in the language in which it was filed.

   ☐ a translation of the international application into _____ which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b)).

2. ☐ This opinion has been established taking into account the rectification of an obvious mistake authorized by or notified to this Authority under Rule 91 (Rule 43*bis*.1(a))

3. With regard to any nucleotide and/or amino acid sequence disclosed in the international application, this opinion has been established on the basis of a sequence listing filed or furnished:

   a. (means)

      ☐ on paper

      ☐ in electronic form

   b. (time)

      ☐ in the international application as filed

      ☐ together with the international application in electronic form

      ☐ subsequently to this Authority for the purposes of search

4. ☐ In addition, in the case that more than one version or copy of a sequence listing has been filed or furnished, the required statements that the information in the subsequent or additional copies is identical to that in the application as filed or does not go beyond the application as filed, as appropriate, were furnished.

5. Additional comments:

Form PCT/ISA/237 (Box No. I) (July 2009)

WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/69881

Box No. V **Reasoned statement under Rule 43*bis*.1(a)(i) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement**

1. Statement

Novelty (N)	Claims	4-15, 19-28, 30-31, 35-46	YES
	Claims	1-3, 16-18, 32-34, 47	NO
Inventive step (IS)	Claims	None.	YES
	Claims	1-28 and 30-47	NO
Industrial applicability (IA)	Claims	1-28 and 30-47	YES
	Claims	None.	NO

2. Citations and explanations:

Claims 1-3, 16-18, 32-34, and 47 lack novelty under PCT Article 33(2) as being anticipated by US 2004/0258177 A1 to Shen et al. (hereinafter 'Shen').

Regarding claim 1, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-8 symbol constellation and at least one of the constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 25 - 46 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0316]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 2, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

Regarding claim 3, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems, abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

-- Please See Supplemental Sheet --

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/68881

Box No. VII Certain defects in the international application

The following defects in the form or contents of the international application have been noted:
Claim 29 is missing.

Form PCT/ISA/237 (Box No. VII) (July 2009)

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/69881

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Box V. 2. Citations and explanations:

Regarding claim 16, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-16 symbol constellation and at least one of the constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 47 - 84 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0318]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 17, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

Regarding claim 18, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems, abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

Regarding claim 32, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 6 such that each symbol of the plurality of symbols includes 6 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a PAM-32 symbol constellation and at least one of the constellation points is a distance greater than 0 and less than the maximum distance specified in FIGS. 85 - 148 from the constellation point of a constellation optimized for capacity at a specific SNR (Examples of such modulations may include BPSK (Binary Phase Shift Key), QPSK (Quadrature Phase Shift Key), 8 PSK (8 Phase Shift Key), 16 QAM (16 Quadrature Amplitude Modulation), and even higher order modulation types, para [0155]; 8 PSK constellation set may be viewed as being a 64 QAM (64 Quadrature Amplitude Modulation), para [0318]; distance, d.sub.H, of the any adjacent elements is one, para [0330]).

Regarding claim 33, Shen further teaches the communication system wherein the code is a Turbo code (a communication system that employs turbo codes, para [0009]).

– Please See Continuation Sheet –

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/69681

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Box V. 2. Citations and explanations:

Regarding claim 34, Shen further teaches the communication system wherein the code is a LDPC code (LDPC (Low Density Parity Check) coded modulation. A novel approach is provided within LDPC coded modulation communication systems. abstract; Another type of communication system that has also received interest is a communication system that employs LDPC (Low Density Parity Check) code, para [0009]).

Regarding claim 47, Shen teaches a digital communication system, comprising: a transmitter configured to transmit signals to a receiver via a communication channel (transmitter employs the satellite dish to communicate to the satellite via a wireless communication channel. The satellite is able to communicate with one or more satellite receivers, para [0075]); wherein the transmitter, comprises: a coder configured to receive user bits and output encoded bits at an expanded output encoded bit rate (information bits that are to be transmitted are encoded using an encoder of the channel encoder/decoder, para [0157]; encoder employs a rate 5/6 LDPC code, and n is equal to 8 such that each symbol of the plurality of symbols includes 8 bits, para [0019]); a mapper configured to map encoded bits to symbols in a symbol constellation (The multi-dimensional space Gray code mapper is operable to map each symbol of the plurality of symbols to a 2m-D (2m-dimensional) M PSK (Phase Shift Key) modulation, para [0016]); a modulator configured to generate a signal for transmission via the communication channel using symbols generated by the mapper (These encoded bits are provided to a modulator of the modulator/demodulator where modulation encoding/symbol mapping may be performed according to the modulation of interest, para [0157]); wherein the receiver, comprises: a demodulator configured to demodulate the received signal via the communication channel (a demodulator portion of the modulator/demodulator where any modulation decoding/symbol mapping . . . These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder, para [0155]); a demapper configured to estimate likelihoods from the demodulated signal (constellation de-mapping, decoding, para [0143]; The digital receiver processing module decodes, descrambles, demaps, and/or demodulates the digital reception formatted, para [0149]); a decoder that is configured to estimate decoded bits from the likelihoods generated by the demapper (These demodulated symbols are then provided to a decoder portion of the channel encoder/decoder where best estimates of the information bits contained within the originally received continuous-time signal, para [0155]; The LLR (Log-Likelihood Ratio) decoding of LDPC codes . . . This LLR will give a measure of the degree to which the communication channel, para [0174]); and wherein the symbol constellation is a N-Dimensional symbol constellation, where the constellation points in at least one dimension are within at least one of the ranges specified in FIGS. 25 - 167 (a 4-D (4-dimensional) QPSK (Quadrature Phase Shift Key) constellation, para [0058]; m-D (m-dimensional) Gray code map construction for 2m-D M PSK modulation, para [0061]; an m-D (m-dimensional) Gray code map (shown where m=2) constructed using two 1-D (1-dimensional) 3 bit basic Gray code maps (shown using 8 PSK shaped constellations) according to the invention, para [0065]).

Claims 4-15, 19-28, 30-31, and 35-46 lack an inventive step under PCT Article 33(3) as being obvious over Shen, in view of US 2007/0280147 A1 to Catreux-Erceg et al. (hereinafter 'Catreux-Erceg').

Regarding claims 4, 19, and 35, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (Such diversity gains improve system performance by increasing received signal-to-noise ratio, . . . the significant increase in system capacity . . . increase system capacity by nearly M additional bits/cycle for each 3-dB increase in SNR, para [0014]; efficient optimization of network capacity, para [0004]; The spacing of the constellation changes constantly due to transmission power control (TPC) . . . This makes an accurate estimation of the constellation very difficult, para [0012]; The gain values 310a and 312a of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.65 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 5, 20, and 36, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (channel gain as seen at finger l at the receive antenna 206 of UE 202a may be represented as h.sub.l+w.sub.2h.sub.2l, para [0061]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 15% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

-- Please See Continuation Sheet --

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/69881

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Box V. 2. Citations and explanations:

Regarding claims 6, 21, and 37, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]. The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). it would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 30% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 7, 22, and 38, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain values 310a and 312a of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.66 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]. The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 45% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 8, 23, and 39, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). it would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 9, 24, and 40, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR (The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). it would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for joint capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

– Please See Continuation Sheet –

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No
PCT/US 09/59861

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Box V. 2. Citations and explanations:

Regarding claims 10, 25, and 41, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; For example, the weight factors W.sub.1 and W.sub.2 may correspond to phase and/or amplitude component feedback adjustments that may be generated by the receiver based on the type of space-time coding, para [0040]; Such diversity gains improve system performance by increasing received signal-to-noise ratio. . . . the significant increase in system capacity . . . increase system capacity by nearly M additional bits/cycle for each 3-dB increase in SNR, para [0014]; The gain values 310a and 312a of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.65 dB and 1.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0062]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 5% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 11, 26, and 42, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; phase and/or amplitude component feedback adjustments that may be generated by the receiver based on the type of space-time coding, para [0040]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 40% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 12, 27, and 43, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 50% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

-- Please See Continuation Sheet --

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/69861

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Box V. 2. Citations and explanations.

Regarding claims 13, 28, and 44, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 60% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 14, 30, and 45, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 70% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Regarding claims 15, 31, and 46, Shen teaches the communication system of claims 1, 16, and 32 as described above. But, Shen does not explicitly teach the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR. However, Catreux-Erceg teaches the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR (parallel code channels . . . reducing power beyond the 20 dB may therefore have only a marginal impact on capacity, para [0030]; The gain of CLM1 using maximizing the SINR criterion as illustrated in graph 304d, over the 1-antenna system, graph 302d, is 2.5 dB across the entire range of system geometry, para [0065]; The gain values 310c and 312c of CLM1 using the maximized SINR criterion over prior art under these channel conditions may be estimated as 0.33 dB and 0.75 dB, respectively, at system geometry values 5 and 10 dB, respectively, para [0064]). It would have been obvious to one skilled in the art to combine the teachings of Shen with those of Catreux-Erceg in order to provide the communication system wherein the constellation provides an increase in capacity at a predetermined SNR that is at least 100% of the gain in capacity achieved by a constellation optimized for PD capacity at the predetermined SNR, because both of them teach encoding, decoding, and processing of signals for multi-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation in communication systems, and to do so would improve the performance of coding and processing of wireless communication signals.

Claims 1-28 and 30-47 have industrial applicability as defined by PCT Article 33(4) because the subject matter can be made or used in industry.

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 08/69881

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

Form PCT/ISA/237 (Supplemental Box) (July 2009)

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T 5
		FRAGOULI ET AL., "Turbo Codes with Non-Uniform Constellations", IEEE Int. Conf. Commun., June 2001, pp. 70-73.	
		KSCHISCHANG ET AL., "Optimal Nonuniform Signaling for Gaussian Channels", IEEE, pp. 913-929, 1993.	
		GOFF ET AL., "Channel capacity of bit-interleaved coded modulation schemes using 8-ary signal constellations", Electronics Letters, Vol. 38, Issue 4, Feb 14, 2002, pp. 187-188.	
		GOFF ET AL., "Signal Constellations for Bit-Interleaved Coded Modulation", IEEE Transactions on Information Theory, Vol. 49, Issue 1, January 2003, pp. 307-313.	
		HAMKINS ET AL., "Asymptotically Dense Spherical Codes - Part I: Wrapped Spherical Codes", IEEE Transactions on Information Theory, November 1997, Vol. 43, No. 6, pp. 1774-1785.	
		HAMKINS ET AL., "Asymptotically Dense Spherical Codes - Part II: Laminated Spherical Codes", IEEE Transactions on Information Theory, November 1997, Vol. 43, No. 6, pp. 1786-1798.	
		HOSSAIN ET AL., "BICM Transmission using Non-Uniform QAM Constellations: Performance Analysis and Design", IEEE International Conference on Communications (ICC), 2010, Conference: May 23- 27, 2010, 7 pgs.	
		HOSSAIN ET AL., "Constellation and Interleaver Design for BICM", Global Telecommunications Conference (GLOBECOM 2011), 2011 IEEE, Conference: December 5-9, 2011, 7 pgs.	
		HOSSAIN ET AL., "Towards Fully Optimized BICM Transceivers", Arxiv.org, December 8, 2010, 31 pgs.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		ISAKA ET AL., "Error performance analysis of multilevel coded asymmetric 8-PSK modulation with multistage decoding and unequal error protection", 1998 IEEE INTERNATIONAL SYMPOSIUM ON CAMBRIDGE, MA, USA 16-21 AUG, 16 August 1998 page 210.	
		JO ET AL., "An Advanced Hierarchical Modulation with Rotated Constellation", The 12th International Conference on Advanced Communication Technology (ICACT), 2010, Conference: February 7-10, 2010, pp. 515-518.	
		KAYHAN ET AL., "Joint Signal-Labeling Optimization for Pragmatic Capacity under Peak-Power Constraint", Global Telecommunications Conference (GLOBECOM 2010), 2010 IEEE, Conference December 6-10, 2010, Miami, FL, USA, 6 pgs.	
		KHANDANI ET AL., "Application of Shaping Technique to Multi-level Turbo-coded Modulation", 2002, 5 pgs.	
		KHANDANI ET AL., "Application of Shaping Technique to Multi-level Turbo-coded Modulation", University of Waterloo, Department of Electrical and Computer Engineering, Waterloo, Ontario, Canada, Technical Report UW-E&CE#02-07, May 30, 2002, 24 pgs.	
		KHOO ET AL., "Bit Interleaved Coded Modulation with Iterative Decoding Using Constellation Shaping", IEEE Transactions on Communications, October 2006, 5 pgs.	
		LEE ET AL., "Hierarchical Constellation Based Adaptive Relay Scheme in Multi-Hop Networks", IEEE Communication Letters, Vol. 11, No. 3, March 2007, pp. 225-227.	
		LIOLIS ET AL., "Amplitude Phase Shift Keying Constellation Design and its Applications to Satellite Digital Video Broadcasting", Retrieved from: http://www.dtic.upf.edu/~aguillen/home_upf/Publications_files/apsk_chapter.pdf, June 1, 2009, 28 pgs.	
		LIU ET AL., "APSK Constellation with Gray Mapping", IEEE Communications Letters, Vol. 15, Issue 12, December 2011, pp.127-1273, Date of Publication October 31, 2011.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		LOGHIN ET AL., "Non-Uniform Constellations for ATSC 3.0", IEEE Transactions on Broadcasting, Vol. 62, No. 1, March 2016, pp. 197-203.	
		LONG DUAN ET AL., "Approaching the AWGN Channel Capacity without Active Shaping", Proceedings of the International Symposium on Information Theory, pp. 374, 1997.	
		MA ET AL., "Coded Modulation Using Superimposed Binary Codes", IEEE Transactions of Information Theory, Vol. 50, No. 12, December 2004, pp. 3331-3343.	
		MAKOWSKI, "On the Optimality of Uniform Pulse Amplitude Modulation", IEEE Transactions on information Theory, December 2006, Vol. 52, No. 12, pp. 5546-5549.	
		MARTINEZ ET AL., "Bit-Interleaved Coded Modulation in the Wideband Regime", Retrieved from: https://arxiv.org/pdf/0710.4046.pdf, Draft, October 22, 2007, 23 pgs.	
		MARTINEZ ET AL., "Coding and Modulation for the Additive Exponential Noise Channel", IEEE International Symposium on Information Theory, 2008. ISIT 2008, Conference: 6-11 July 2008, 5 pgs.	
		MERIC ET AL., "Generic Approach for Hierarchical Modulation Performance Analysis: Application to DVB-SH and DVB-S2", Retrieved from: https://arxiv.org/abs/1103.1742, Submitted March 9, 2011, Draft March 10, 2011, 17 pgs.	
		MHEICH ET AL., "Constellation Shaping for Broadcast Channels in Practical Situations", 19th European Signal Processing Conference (EUSIPCO 2011), Barcelona, Spain, August 29- September 2, 2011, pp. 96-100.	
		MILOVANOVIC ET AL., "Simple Optimization Method of One-Dimensional M-PAM Constellations for the AWGN Channels", 4th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Services, October 13-15, 1999, 4 pgs.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		MOORE ET AL., "Pairwise optimization of modulation constellations for non-uniform sources Modulation", Can. J. Elect. Computer Eng. Vol. 34, pp 167-177, 2009.	
		MUHAMMAD, NABIL SVEN, "Coding and Modulation for Spectral Efficient Transmission", University of Stuttgart. August 25, 2010 (http://dx.doi.org/10.18419/opus-2676).	
		MUHAMMAD ET AL., "Joint Optimization of Signal Constellation and Bit Labeling for Bit-Interleaved Coded Modulation with Iterative Decoding", IEEE Communications Letters, September 2005, Vol. 9, No. 9, pp. 775-777.	
		NGO ET AL., "A New Iterative Decoder for Turbo Codes on the Nonlinear Channel with Non-uniform 16QAM Modulation", Turbo Coding 2006, April 3-7, 2006, Munich, 7 pgs.	
		NGO ET AL., "Performance of non-uniform 16QAM modulation over linear and nonlinear channels", Electronics Letters, Vol. 42, Issue 9, April 27, 2006, 2 pgs.	
		OTNES ET AL., "Adaptive Data Rate using ARQ and Nonuniform Constellations", Vehicular Technology Conference pp 1211-1215, 2001.	
		RAPHAELI ET AL., "An Improved Pragmatic Turbo Encoding Scheme for High Spectral Efficiency Using Constellation Shaping", IEEE International Conference on Communications, ICC 2003, Conference May 11-15, 2003, Anchorage, AK, USA, 6 pgs.	
		RAPHAELI ET AL., "Constellation Shaping for Pragmatic Turbo-Coded Modulation with High Spectral Efficiency", IEEE Transactions on Communications, March 2004, Vol. 52, No. 3, pp. 341-345.	
		RUOTSALAINEN ET AL., "On the construction of the higher dimensional constellations", ISIT 2000, Lausanne, Switzerland, June 30-July 5, 2002, pg. 490.	
		SAWAYA, "Performance optimization for capacity-approaching channel coding schemes", XP055604929 Ph.D. report, ENST Paris, France 31 March 2002.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		SAWAYA ET AL., "Multilevel coded modulations based on asymmetric constellations", PROCEEDINGS OF THE 2001 IEEE INTERNATIONAL SYMPOSIUM ON INFORMATION THEORY NEW YORK, NY: IEEE, US, 24 June 2001 pages 281-281.	
		SCHRECKENACH ET AL., "Signal Shaping Using Non-Unique Symbol Mappings", Proceedings of the 43rd Annual Allerton Conference on Communication, Control and Computing, September 2005, 10 pgs.	
		SHANNON ET AL., "A Mathematical Theory of Communication", The Bell System Technical Journal, Vol. 27, July, October 1948, 379-423, 623-656.	
		SHEN ET AL., "On the Design of Modern Multilevel Coded Modulation for Unequal Error Protection", IEEE International Conference on Communications, 2008. ICC '08., Conference: 19-23 May 2008, 6 pgs.	
		SOMMER ET AL., "Signal Shaping by Non-Uniform QAM for AWGN Channels and Applications Using Turbo Coding", ITG Conference on Source and Channel Coding, January 2000, pp. 81-86.	
		SOUTO ET AL., "Iterative Detection and Channel Estimation for WCDMA Systems Employing Non-Uniform QAM Constellations", IST Mobile and Wireless Communications Summit, June 2006, 6 pgs.	
		SOUTO ET AL., "Iterative Turbo Multipath Interference Cancellation for WCDMA System with Non-Uniform Modulation", IEEE, 2005, 5 pgs.	
		SOUTO ET AL., "Non-Uniform Constellations for Broadcasting and Multicasting Services in WCDMA Systems", Retrieved from: http://www.eurasip.org/Proceedings/Ext/IST05/papers/424.pdf, June 19-23, 2005, 5 pgs.	
		STIERSTORFER ET AL., "Asymptotically Optimal Mappings for BICM with M-PAM and M¬¬2QAM", Retrieved from: http://www.lit.lnt.de/papers/elet_set_part_cst_2009.pdf, Draft, July 13, 2009, 6 pgs.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		STIERSTORFER ET AL., "Optimizing BICM with convolutional codes for transmission over the AWGN channel", Int. Zurich Seminar on Communications (IZS), March 3-5, 2010, 5 pgs.	
		SUN ET AL., "Approaching Capacity by Equiprobable Signaling on the Gaussian Channel", IEEE Transactions on Information Theory, September 1993, Vol. 39, No. 5, pp. 1714-1716.	
		THEODORAKOPOULOS ET AL., "Comparison of Uniform and Non-uniform M-QAM Schemes for Mobile Video Applications", Proceedings of the 2005 Systems Communications, 2005, Conference: August 14-17, 2005, 6 pgs.	
		TRAN ET AL., "Signal Mappings of 8-Ary Constellations for BICM-ID Systems Over a Rayleigh Fading Channel", Canadian Conference on Electrical and Computer Engineering, 2004, pp. 1809-1813.	
		UNGERBOECK ET AL., "Channel Coding with Multilevel/Phase Signals", IEEE Transactions on Information Theory, January 1982, Vol. IT-28, No. 1, pp. 55-67.	
		VALLES ET AL., "Constellation Design for Improved Iterative LDPC Decoding", Aerospace Conference, 2008 IEEE, Conference: March 1-8, 2008, Big Sky, MT, USA, DOI: 10.1109/AERO.2008.4526371, 7 pgs.	
		VITTHALADEVUNI ET AL., "A Recursive Algorithm for the Exact BER Computation of Generalized Hierarchical QAM Constellations", IEEE Transactions on Information Theory, Vol. 49, No. 1, January 14, 2003, pp. 297-307.	
		VON DEETZEN ET AL., "On Code Design for Unequal Error Protection Multilevel Coding", 7th International ITG Conference on Source and Channel Coding (SCC), 2008, Conference: 14-16 Jan. 2008, 4 pgs.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

EXAMINER SIGNATURE			
Examiner Signature		Date Considered	

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF EFS WEB TRANSMISSION

I hereby certify that this correspondence is being electronically filed with the United States Patent and Trademark Office on January 9, 2020 at or before 11:59 p.m. Pacific Time under the Rules of 37 C.F.R. 1.8.

/Katrina Lee/
Katrina Lee

Appl No.	: 16/728,384	Confirmation No. 6350
Applicant	: Constellation Designs, LLC	
Filed	: December 27, 2019	
Title	: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	
TC/A.U.	: Unassigned	
Examiner	: Unassigned	
Docket No.	: C44-00999.CON15	
Customer No.	: 71897	

INFORMATION DISCLOSURE STATEMENT § 1.97(b)

MAIL STOP AMENDMENT
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806
January 9, 2020

Commissioner

In compliance with the duty of disclosure under 37 C.F.R. §§ 1.56, 1.97 and 1.98, and in accordance with the provisions in the MPEP §§ 609 and 707.05(b), enclosed is FORM PTO/SB/08A/B with a listing of references that are known to applicant. Copies of each of the references listed under “Non-Patent Literature Documents" are enclosed.

It is respectfully requested that these references be considered in the examination of this application and identified on the list of references cited on the patent issuing on this application. Applicant also requests that an initialed copy of said FORM PTO/SB/08A/B be entered in the application file, and returned to the Applicant with the next official communication from the Office in accordance with MPEP § 609.

It is believed that there are no fees due in connection with this filing. If there are any fees due in connection with this filing, the Commissioner is authorized to charge Deposit Account 50-4407. Please show our docket number with any charge to the Deposit Account.

Respectfully submitted,

KPPB LLP

/Isaac M. Fine/
Isaac M. Fine
Registration No. 74,274
949.852.0000

IMF/kl

Enclosures

Electronic Acknowledgement Receipt

EFS ID:	38241866
Application Number:	16728384
International Application Number:	
Confirmation Number:	6350
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Customer Number:	71897
Filer:	Isaac M. Fine/Katrina Lee
Filer Authorized By:	Isaac M. Fine
Attorney Docket Number:	C44-00999.CON15
Receipt Date:	09-JAN-2020
Filing Date:	
Time Stamp:	14:27:19
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Non Patent Literature	fraggouliturbocodes.pdf	390410 621ad51ce88fe66fd91e2b703ebca3153e85d561	no	4

Warnings:

Information:					
2	Non Patent Literature	kschischang.pdf	6122211 239d559a97760e35283e617f3aa2bca6145b9d78	no	17
Warnings:					
Information:					
3	Non Patent Literature	goffchannel.pdf	608813 8300d29d7e33e450eea7755f98bdc1af90d44fde	no	2
Warnings:					
Information:					
4	Non Patent Literature	goffsignal.pdf	533386 08682211250094e0d3133079f0ddb4e4e99427c9	no	7
Warnings:					
Information:					
5	Non Patent Literature	hamkinsasymptotically part1.pdf	6026510 a0799aa7fb2d0f23327e603b28f651b34cd2f814	no	12
Warnings:					
Information:					
6	Non Patent Literature	hamkinsasymptotically part2.pdf	5818777 a0f42b6144daf4f408d8504256ea09d6383d8f2e	no	13
Warnings:					
Information:					
7	Non Patent Literature	hossainbicm.pdf	323686 aaf4a06aa1201a65e16e38138ffcd02e3f032577	no	7
Warnings:					
Information:					
8	Non Patent Literature	hossainconstellation.pdf	182678 04b729ee609b0ec84a02c4daf511fc6d8de75f63	no	7
Warnings:					
Information:					

9	Non Patent Literature	hossaintowards.pdf	375211 ccbc4ecf5df1d466ee5a31a5cd34535f1d9d8b31	no	31
Warnings:					
Information:					
10	Non Patent Literature	isakaerror.pdf	74898 03c7acaddcdb1983b741feb0d9102bb51e33bc3e	no	1
Warnings:					
Information:					
11	Non Patent Literature	joanadvanced.pdf	561247 9e772ce4ceaa267baece6d8d065e5b26ded41c07	no	4
Warnings:					
Information:					
12	Non Patent Literature	kayhanjoint.pdf	129692 55fdda53da7878cd3a24e0a90a9001e17735b88d	no	5
Warnings:					
Information:					
13	Non Patent Literature	khandanietal5pgs.pdf	281428 f4e483e3d247dbbe421a8c28e5c203927588ae48	no	5
Warnings:					
Information:					
14	Non Patent Literature	khandanietal24pgs.pdf	365828 beaf09d0bb5ad238c65fc6048f46cfe6caa79ab5	no	24
Warnings:					
Information:					
15	Non Patent Literature	khoo.pdf	179157 e0e86f5208b402048e59269eff16de70f9e7c142	no	4
Warnings:					
Information:					

16	Non Patent Literature	leehierarchical.pdf	146113 fef3ed181ecc57d6d0cd86e41c3512c6bfa1874b	no	3
Warnings:					
Information:					
17	Non Patent Literature	liolis.pdf	586320 0f7e8e85037f7cf1aeddadd928a7f8744ac1a305	no	28
Warnings:					
Information:					
18	Non Patent Literature	liuapsk.pdf	348918 a856c29d381c50575d1691d3b549fa9411e5752d	no	3
Warnings:					
Information:					
19	Non Patent Literature	loghin.pdf	309076 f3a9489e2f28baa6f7be067ff23aaec7d79b610a	no	7
Warnings:					
Information:					
20	Non Patent Literature	longduan.pdf	264622 cc3cc2ac0152ae47257009ce4d195fc1e243a329	no	1
Warnings:					
Information:					
21	Non Patent Literature	maetal2004.pdf	446397 57d731573bd30d5537cf375d0e59e6d4405c09fb	no	13
Warnings:					
Information:					
22	Non Patent Literature	makowskiontheoptimality.pdf	1088878 bed6acf845390c6a0af54e50cb47ece45f4a6aa6	no	4
Warnings:					
Information:					

23	Non Patent Literature	martinezbit.pdf	197201 78db06e649d8c80352b9202e910581ee73ccd5e5	no	23
Warnings:					
Information:					
24	Non Patent Literature	martinezcoding.pdf	103108 78eeac279c38c548f8537182051afb2e07b9d672	no	5
Warnings:					
Information:					
25	Non Patent Literature	mericgeneric.pdf	193862 448458e03f19e47114642aecf171478fddc145c0	no	17
Warnings:					
Information:					
26	Non Patent Literature	mheichconstellation.pdf	224912 02f47ea9be73246ff08379741e45e55d9c8c2961	no	5
Warnings:					
Information:					
27	Non Patent Literature	milovanovicsimple.pdf	327498 0fd343ef9bb84c5ed468de42f843bc8e3d7afa33	no	4
Warnings:					
Information:					
28	Non Patent Literature	moore.pdf	3156605 9466ac134bae98381a6c68eb0f28910b87d2eb78	no	11
Warnings:					
Information:					
29	Non Patent Literature	muhammadcoding.pdf	11987405 d6c471138eb64d34c4543e73adccad84c929f2d1	no	176
Warnings:					
Information:					

30	Non Patent Literature	muhammadjointoptimization.pdf	935159 99a6ab7e1db0e3442645abaa42c09419fcbb6473	no	3
Warnings:					
Information:					
31	Non Patent Literature	ngoanew.pdf	407326 76cc2fe91579af1b75fe0101edbca5046f8ef152	no	6
Warnings:					
Information:					
32	Non Patent Literature	ngoperformanceofnon.pdf	59924 dd0cd2b40aedd6fdeee4a6c434527bc82f22dc6c	no	2
Warnings:					
Information:					
33	Non Patent Literature	otnes.pdf	1354384 4bc653a1e0ee6a1b1a5229d5dc5eea1a3648578b	no	5
Warnings:					
Information:					
34	Non Patent Literature	raphaelianimproved.pdf	230842 f0dbc43b897d16edaa921041b88628d394b2a879	no	6
Warnings:					
Information:					
35	Non Patent Literature	raphaeliconstellationshaping.pdf	1439924 fdc4fa28292192ac91efae5b403f643ffd5ab681	no	5
Warnings:					
Information:					
36	Non Patent Literature	ruotsalainenontheconstruction.pdf	261120 df0d8be263ff83a0be4e24a8856561ef7e24d346	no	1
Warnings:					
Information:					

37	Non Patent Literature	sawayaperformance.pdf	6304841 32e1db98deaf77b65f709d32f1ae7496e4a0fa4b	no	164
Warnings:					
Information:					
38	Non Patent Literature	sawayamultilevel.pdf	88973 24f0edeb2715b9fc7f157d681d942a7ebef4ebda	no	1
Warnings:					
Information:					
39	Non Patent Literature	schreckenachsignal.pdf	209659 8bc4111c93bf1e3292ac81e44b21840c02428dbf	no	10
Warnings:					
Information:					
40	Non Patent Literature	shannon.pdf	10301926 1b932e60cdd70f4e93888336699faa7745ed2be2	no	45
Warnings:					
Information:					
41	Non Patent Literature	shen.pdf	135583 57d1485f0eacdc3895dc9821a10c0d9c070d178b	no	5
Warnings:					
Information:					
42	Non Patent Literature	sommersignal.pdf	727402 7404b7855c4f6dfdcda39d98195ea7b4dcf7d7aa	no	6
Warnings:					
Information:					
43	Non Patent Literature	soutopp811-815.pdf	8680146 ae276f75ce5d7b3c81d6e54dec04ff2a4b208476	no	5
Warnings:					
Information:					

44	Non Patent Literature	soutoiterative.pdf	299131 2b7fa6de11d69d67dd7bca3c68fe2a5dccc443be	no	5
Warnings:					
Information:					
45	Non Patent Literature	soutononuniform.pdf	142767 b1ca46a38b92c3aad5ecb68284a0886ad6626803	no	5
Warnings:					
Information:					
46	Non Patent Literature	stierstorferasympt.pdf	151318 1ea232ba33ef3181498598a0a01caf1a68097dae	no	6
Warnings:					
Information:					
47	Non Patent Literature	stierstorferopt.pdf	124559 89490d4419acf21ca69f4992c8ef7b0b7c72169c	no	4
Warnings:					
Information:					
48	Non Patent Literature	sunapproachingcapacity.pdf	842100 ad8370cfa2aa04fbfbaf2113c44b961a0127e35f	no	3
Warnings:					
Information:					
49	Non Patent Literature	theodorakopouloscomp.pdf	964630 8d446ed2d5655ea71096cb65b0b11d6343b9fb3b	no	6
Warnings:					
Information:					
50	Non Patent Literature	transignal.pdf	1548337 4157eaabbee681232e8aaf661bc9107d30bd285a	no	5
Warnings:					
Information:					

51	Non Patent Literature	ungerboeckchannelcoding.pdf	3786640 86cf233d058fc18c00e7e60aedf6f641c60ba498	no	13
Warnings:					
Information:					
52	Non Patent Literature	valles.pdf	3664467 5689140bf1637412ab2f471be6e3e3134a54c0e5	no	7
Warnings:					
Information:					
53	Non Patent Literature	vitthaladevuni.pdf	1046325 c8a05cdd2209389313871008 98e7da9bf90aff30	no	11
Warnings:					
Information:					
54	Non Patent Literature	vondeetzen.pdf	242144 87d96e8073167305a00031940431558ae930d731	no	4
Warnings:					
Information:					
55	Information Disclosure Statement (IDS) Form (SB08)	00999con15ids1449part2.pdf	137088 21502eefdd977629b3ed87a38da4bc87f0a2cdc9	no	7
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					
56	Transmittal Letter	00999con15idsxtmlpart2.pdf	120006 6d54f39c7430526e7a6eb181f2990cf728eefc2f	no	2
Warnings:					
Information:					
			Total Files Size (in bytes):	85561568	

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

New Applications Under 35 U.S.C. 111

If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371

If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office

If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T 5
		WANG ET AL, "Shaping Gain for AWGN Channel by Non-Uniform Constellation in LDPC-Coded System", 11th IEEE Singapore International Conference on Communication Systems, 2008, ICCS 2008, Conference: November 19-21, 2008, pp. 1302-1306.	
		WU ET AL., "Non-uniform and Large Distance Constellation Design for Hierarchical Modulation", 2010 IEEE International Conference on Communications (ICC), Conference: May 23-27, 2010, 5 pgs.	
		XIE ET AL., "Bit-Interleaved LDPC-Coded Modulation with Iterative Demapping and Decoding", IEEE 69th Vehicular Technology Conference, Barcelona, Spain, Date of Conference April 26-29, 2009, 5 pgs.	
		XIE ET AL., "On the Channel Capacity and Iterative Demapping of Generalized 4PAM over AWGN Channel", 11th IEEE Singapore International Conference on Communication Systems, 2008. ICCS 2008. December 2008, pp. 860-863.	
		YANG ET AL., "A Novel BICM-ID System Approaching Shannon-Limit at High Spectrum Efficiency", IEICE Trans. Commun., Vol. E94-B, No. 3, March 2011, pp. 793-795.	
		YUAN ET AL., "Robust Hierarchical Broadcasting for AWGN and Flat Rayleigh Fading Channels using Multilevel Codes", Proceedings of IEEE Fifth International Symposium on Communication Theory & Applications, ISCTA'99, Ambleside, UK, 1999, S. 3 pgs.	
		ZESONG ET AL., "Shaping Gain by Non-Uniform QAM Constellation with Binary Turbo Coded Modulation", PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS, IEEE, vol. 2. pp. 1863-1867, 7 September 2003.	
		ZHANG ET AL., "A New Constellation Shaping Method and Its Performance Evaluation in BICM-ID", IEEE 70th Vehicular Technology Conference Fall (VTC 2009-Fall). Conference: Sept. 20-23, 2009, 5 pgs.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		ZHENG ET AL., "Shaping Gain of LDPC Coded-QAM Transmitting Systems with Non-Uniform Constellation", IEEE, pp. 6-9, 2007.	

EXAMINER SIGNATURE			
Examiner Signature		Date Considered	

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF EFS WEB TRANSMISSION

I hereby certify that this correspondence is being electronically filed with the United States Patent and Trademark Office on January 9, 2020 at or before 11:59 p.m. Pacific Time under the Rules of 37 C.F.R. 1.8.

/Katrina Lee/
Katrina Lee

Appl No.	: 16/728,384	Confirmation No. 6350
Applicant	: Constellation Designs, LLC	
Filed	: December 27, 2019	
Title	: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	
TC/A.U.	: Unassigned	
Examiner	: Unassigned	
Docket No.	: C44-00999.CON15	
Customer No.	: 71897	

INFORMATION DISCLOSURE STATEMENT § 1.97(b)

MAIL STOP AMENDMENT
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806
January 9, 2020

Commissioner

In compliance with the duty of disclosure under 37 C.F.R. §§ 1.56, 1.97 and 1.98, and in accordance with the provisions in the MPEP §§ 609 and 707.05(b), enclosed is FORM PTO/SB/08A/B with a listing of references that are known to applicant. Copies of each of the references listed under “Non-Patent Literature Documents" are enclosed.

It is respectfully requested that these references be considered in the examination of this application and identified on the list of references cited on the patent issuing on this application. Applicant also requests that an initialed copy of said FORM PTO/SB/08A/B be entered in the application file, and returned to the Applicant with the next official communication from the Office in accordance with MPEP § 609.

It is believed that there are no fees due in connection with this filing. If there are any fees due in connection with this filing, the Commissioner is authorized to charge Deposit Account 50-4407. Please show our docket number with any charge to the Deposit Account.

Respectfully submitted,

KPPB LLP

/Isaac M. Fine/
Isaac M. Fine
Registration No. 74,274
949.852.0000

IMF/kl

Enclosures

Electronic Acknowledgement Receipt	
EFS ID:	38241901
Application Number:	16728384
International Application Number:	
Confirmation Number:	6350
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Customer Number:	71897
Filer:	Isaac M. Fine/Katrina Lee
Filer Authorized By:	Isaac M. Fine
Attorney Docket Number:	C44-00999.CON15
Receipt Date:	09-JAN-2020
Filing Date:	
Time Stamp:	14:30:27
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Non Patent Literature	wang.pdf	115141 a10fd0f0f3efbbfa2aa1354c97dd8733d1341bba	no	5
Warnings:					

Information:					
2	Non Patent Literature	wu.pdf	259648 1845442dbb456c6b6dd7fa7ccf9901777afc5b42	no	5
Warnings:					
Information:					
3	Non Patent Literature	xiebit.pdf	2291075 65fbed5638e027cfb76e55f189445ccfb080ebc1	no	5
Warnings:					
Information:					
4	Non Patent Literature	xieonthe.pdf	731331 91121ee77f4c53ad68bb92ce06b84526eb33892d	no	4
Warnings:					
Information:					
5	Non Patent Literature	yang.pdf	759114 9ca8e86d1d4db4cd7af9d31c3b42584b495d0037	no	3
Warnings:					
Information:					
6	Non Patent Literature	yuan.pdf	294100 6499ddeb266e23c74d012f2e2f0ffa2179710ca2	no	3
Warnings:					
Information:					
7	Non Patent Literature	zesongpp1863-1867.pdf	850437 1f536566b7fcaeac1a9c0c614a67662e3e939480	no	5
Warnings:					
Information:					
8	Non Patent Literature	zhang.pdf	220648 c6bb4cac0581128b0b0b3bf5c2993b7cbb6f17b6	no	5
Warnings:					
Information:					

9	Non Patent Literature	zheng.pdf	864321 0be5b8a24fca3ea6a556e5f81609e16863541cfe	no	4
Warnings:					
Information:					
10	Information Disclosure Statement (IDS) Form (SB08)	00999con15ids1449part3.pdf	110793 ee6289af7b1adb3574796ea9426082a14f8a4e10	no	2
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					
11	Transmittal Letter	00999con15idsxtmlpart3.pdf	120004 bae08a5a4c09cdb99790b118dd24be9c9d2c7165	no	2
Warnings:					
Information:					
		Total Files Size (in bytes):	6616612		

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

New Applications Under 35 U.S.C. 111

If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371

If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office

If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

United States Patent and Trademark Office

Office of the Chief Financial Officer

Document Code:WFEE

User :C46575

Sale Accounting Date:01/28/2020

Sale Item Reference Number	Effective Date
16728384	12/27/2019

Document Number	Fee Code	Fee Code Description	Amount Paid	Payment Method
I20201RF46174080	1830	PROCESSING FEE, EXCEPT PROV. APPLS.	$140.00	Deposit Account

PATENT APPLICATION FEE DETERMINATION RECORD

Substitute for Form PTO-875

Application or Docket Number: 16/728,384

APPLICATION AS FILED - PART I

FOR	(Column 1) NUMBER FILED	(Column 2) NUMBER EXTRA	SMALL ENTITY RATE($)	SMALL ENTITY FEE($)	OR	OTHER THAN SMALL ENTITY RATE($)	OTHER THAN SMALL ENTITY FEE($)
BASIC FEE (37 CFR 1.16(a), (b), or (c))	N/A	N/A	N/A			N/A	300
SEARCH FEE (37 CFR 1.16(k), (i), or (m))	N/A	N/A	N/A			N/A	660
EXAMINATION FEE (37 CFR 1.16(o), (p), or (q))	N/A	N/A	N/A			N/A	760
TOTAL CLAIMS (37 CFR 1.16(i))	30 minus 20 =	* 10			OR	x 100 =	1000
INDEPENDENT CLAIMS (37 CFR 1.16(h))	3 minus 3 =	*				x 460 =	0.00
APPLICATION SIZE FEE (37 CFR 1.16(s))	If the specification and drawings exceed 100 sheets of paper, the application size fee due is $310 ($155 for small entity) for each additional 50 sheets or fraction thereof. See 35 U.S.C. 41(a)(1)(G) and 37 CFR 1.16(s).						0.00
MULTIPLE DEPENDENT CLAIM PRESENT (37 CFR 1.16(j))							0.00
* If the difference in column 1 is less than zero, enter "0" in column 2.			TOTAL			TOTAL	2720

APPLICATION AS AMENDED - PART II

		(Column 1) CLAIMS REMAINING AFTER AMENDMENT		(Column 2) HIGHEST NUMBER PREVIOUSLY PAID FOR	(Column 3) PRESENT EXTRA	SMALL ENTITY RATE($)	SMALL ENTITY ADDITIONAL FEE($)	OR	OTHER THAN SMALL ENTITY RATE($)	OTHER THAN SMALL ENTITY ADDITIONAL FEE($)
AMENDMENT A	Total (37 CFR 1.16(i))	*	Minus	**	=	x =		OR	x =	
	Independent (37 CFR 1.16(h))	*	Minus	***	=	x =		OR	x =	
	Application Size Fee (37 CFR 1.16(s))									
	FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							OR		
						TOTAL ADD'L FEE		OR	TOTAL ADD'L FEE	

		(Column 1) CLAIMS REMAINING AFTER AMENDMENT		(Column 2) HIGHEST NUMBER PREVIOUSLY PAID FOR	(Column 3) PRESENT EXTRA	RATE($)	ADDITIONAL FEE($)	OR	RATE($)	ADDITIONAL FEE($)
AMENDMENT B	Total (37 CFR 1.16(i))	*	Minus	**	=	x =		OR	x =	
	Independent (37 CFR 1.16(h))	*	Minus	***	=	x =		OR	x =	
	Application Size Fee (37 CFR 1.16(s))									
	FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							OR		
						TOTAL ADD'L FEE		OR	TOTAL ADD'L FEE	

* If the entry in column 1 is less than the entry in column 2, write "0" in column 3.
** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 20, enter "20".
*** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 3, enter "3".
The "Highest Number Previously Paid For" (Total or Independent) is the highest found in the appropriate box in column 1.

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

APPLICATION NUMBER	FILING or 371(c) DATE	GRP ART UNIT	FIL FEE REC'D	ATTY.DOCKET.NO	TOT CLAIMS	IND CLAIMS
16/728,384	12/27/2019		2720	C44-00999.CON15	30	3

CONFIRMATION NO. 6350

FILING RECEIPT

71897
KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

OC000000114292135

Date Mailed: 01/30/2020

Receipt is acknowledged of this non-provisional utility patent application. The application will be taken up for examination in due course. Applicant will be notified as to the results of the examination. Any correspondence concerning the application must include the following identification information: the U.S. APPLICATION NUMBER, FILING DATE, NAME OF FIRST INVENTOR, and TITLE OF INVENTION. Fees transmitted by check or draft are subject to collection.

Please verify the accuracy of the data presented on this receipt. If an error is noted on this Filing Receipt, please submit a written request for a corrected Filing Receipt, including a properly marked-up ADS showing the changes with strike-through for deletions and underlining for additions. If you received a "Notice to File Missing Parts" or other Notice requiring a response for this application, please submit any request for correction to this Filing Receipt with your reply to the Notice. When the USPTO processes the reply to the Notice, the USPTO will generate another Filing Receipt incorporating the requested corrections provided that the request is grantable.

Inventor(s)

Maged F. Barsoum, San Jose, CA;
Christopher R. Jones, Pacific Palisades, CA;

Applicant(s)

Constellation Designs, LLC, Anaheim, CA;

Assignment For Published Patent Application

Constellation Designs, LLC, Anaheim, CA

Power of Attorney: None

Domestic Priority data as claimed by applicant

This application is a CON of 16/206,991 11/30/2018 PAT 10567980
which is a CON of 15/682,475 08/21/2017 PAT 10149179
which is a CON of 15/200,800 07/01/2016 PAT 9743292
which is a CON of 14/491,731 09/19/2014 PAT 9385832
which is a CON of 13/618,630 09/14/2012 PAT 8842761
which is a CON of 13/118,921 05/31/2011 PAT 8270511
which is a CON of 12/156,989 06/05/2008 PAT 7978777
which claims benefit of 60/933,319 06/05/2007

Foreign Applications for which priority is claimed (You may be eligible to benefit from the **Patent Prosecution Highway** program at the USPTO. Please see http://www.uspto.gov for more information.) - None.
Foreign application information must be provided in an Application Data Sheet in order to constitute a claim to foreign priority. See 37 CFR 1.55 and 1.76.

Permission to Access Application via Priority Document Exchange: Yes

Permission to Access Search Results: Yes

Applicant may provide or rescind an authorization for access using Form PTO/SB/39 or Form PTO/SB/69 as appropriate.

If Required, Foreign Filing License Granted: 01/28/2020

The country code and number of your priority application, to be used for filing abroad under the Paris Convention, is **US 16/728,384**

Projected Publication Date: 05/07/2020

Non-Publication Request: No

Early Publication Request: No

Title

Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

Preliminary Class

Statement under 37 CFR 1.55 or 1.78 for AIA (First Inventor to File) Transition Applications: No

PROTECTING YOUR INVENTION OUTSIDE THE UNITED STATES

Since the rights granted by a U.S. patent extend only throughout the territory of the United States and have no effect in a foreign country, an inventor who wishes patent protection in another country must apply for a patent in a specific country or in regional patent offices. Applicants may wish to consider the filing of an international application under the Patent Cooperation Treaty (PCT). An international (PCT) application generally has the same effect as a regular national patent application in each PCT-member country. The PCT process **simplifies** the filing of patent applications on the same invention in member countries, but **does not result** in a grant of "an international patent" and does not eliminate the need of applicants to file additional documents and fees in countries where patent protection is desired.

Almost every country has its own patent law, and a person desiring a patent in a particular country must make an application for patent in that country in accordance with its particular laws. Since the laws of many countries differ in various respects from the patent law of the United States, applicants are advised to seek guidance from specific foreign countries to ensure that patent rights are not lost prematurely.

Applicants also are advised that in the case of inventions made in the United States, the Director of the USPTO must issue a license before applicants can apply for a patent in a foreign country. The filing of a U.S. patent application serves as a request for a foreign filing license. The application's filing receipt contains further information and guidance as to the status of applicant's license for foreign filing.

Applicants may wish to consult the USPTO booklet, "General Information Concerning Patents" (specifically, the section entitled "Treaties and Foreign Patents") for more information on timeframes and deadlines for filing foreign patent applications. The guide is available either by contacting the USPTO Contact Center at 800-786-9199, or it can be viewed on the USPTO website at http://www.uspto.gov/web/offices/pac/doc/general/index.html.

For information on preventing theft of your intellectual property (patents, trademarks and copyrights), you may wish to consult the U.S. Government website, http://www.stopfakes.gov. Part of a Department of Commerce initiative, this website includes self-help "toolkits" giving innovators guidance on how to protect intellectual property in specific countries such as China, Korea and Mexico. For questions regarding patent enforcement issues, applicants may call the U.S. Government hotline at 1-866-999-HALT (1-866-999-4258).

LICENSE FOR FOREIGN FILING UNDER

Title 35, United States Code, Section 184

Title 37, Code of Federal Regulations, 5.11 & 5.15

GRANTED

The applicant has been granted a license under 35 U.S.C. 184, if the phrase "IF REQUIRED, FOREIGN FILING LICENSE GRANTED" followed by a date appears on this form. Such licenses are issued in all applications where the conditions for issuance of a license have been met, regardless of whether or not a license may be required as set forth in 37 CFR 5.15. The scope and limitations of this license are set forth in 37 CFR 5.15(a) unless an earlier license has been issued under 37 CFR 5.15(b). The license is subject to revocation upon written notification. The date indicated is the effective date of the license, unless an earlier license of similar scope has been granted under 37 CFR 5.13 or 5.14.

This license is to be retained by the licensee and may be used at any time on or after the effective date thereof unless it is revoked. This license is automatically transferred to any related applications(s) filed under 37 CFR 1.53(d). This license is not retroactive.

The grant of a license does not in any way lessen the responsibility of a licensee for the security of the subject matter as imposed by any Government contract or the provisions of existing laws relating to espionage and the national security or the export of technical data. Licensees should apprise themselves of current regulations especially with respect to certain countries, of other agencies, particularly the Office of Defense Trade Controls, Department of State (with respect to Arms, Munitions and Implements of War (22 CFR 121-128)); the Bureau of Industry and Security, Department of Commerce (15 CFR parts 730-774); the Office of Foreign AssetsControl, Department of Treasury (31 CFR Parts 500+) and the Department of Energy.

NOT GRANTED

No license under 35 U.S.C. 184 has been granted at this time, if the phrase "IF REQUIRED, FOREIGN FILING LICENSE GRANTED" DOES NOT appear on this form. Applicant may still petition for a license under 37 CFR 5.12, if a license is desired before the expiration of 6 months from the filing date of the application. If 6 months has lapsed from the filing date of this application and the licensee has not received any indication of a secrecy order under 35 U.S.C. 181, the licensee may foreign file the application pursuant to 37 CFR 5.15(b).

SelectUSA

The United States represents the largest, most dynamic marketplace in the world and is an unparalleled location for business investment, innovation, and commercialization of new technologies. The U.S. offers tremendous resources and advantages for those who invest and manufacture goods here. Through SelectUSA, our nation works to promote and facilitate business investment. SelectUSA provides information assistance to the international investor

community; serves as an ombudsman for existing and potential investors; advocates on behalf of U.S. cities, states, and regions competing for global investment; and counsels U.S. economic development organizations on investment attraction best practices. To learn more about why the United States is the best country in the world to develop technology, manufacture products, deliver services, and grow your business, visit http://www.SelectUSA.gov or call +1-202-482-6800.

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

APPLICATION NO.	FILING DATE	FIRST NAMED INVENTOR	ATTORNEY DOCKET NO.	CONFIRMATION NO.
16/728,384	12/27/2019	Maged F. Barsoum	C44-00999.CON15	6350

71897 7590 02/04/2020

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER	
ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
02/04/2020	ELECTRONIC

Please find below and/or attached an Office communication concerning this application or proceeding.

The time period for reply, if any, is set in the attached communication.

Notice of the Office communication was sent electronically on above-indicated "Notification Date" to the following e-mail address(es):

PAIR@KPPB.COM

Decision Granting Request for Prioritized Examination (Track I)	**Application No.** 16/728,384	**Applicant(s)** Barsoum et al.	
	Examiner CHERYL P GIBSON BAYLOR	**Art Unit** OPET	**AIA (FITF) Status** No

1. THE REQUEST FILED 27 December 2019 IS **GRANTED** .

The above-identified application has met the requirements for prioritized examination

A. ☑ for an original nonprovisional application (Track I).
B. ☐ for an application undergoing continued examination (RCE).

2. **The above-identified application will undergo prioritized examination**. The application will be accorded special status throughout its entire course of prosecution until one of the following occurs:

A. filing a **petition for extension of time** to extend the time period for filing a reply;

B. filing an **amendment to amend the application to contain more than four independent claims, more than thirty total claims** , or a multiple dependent claim;

C. filing a **request for continued examination** ;

D. filing a notice of appeal;

E. filing a request for suspension of action;

F. mailing of a notice of allowance;

G. mailing of a final Office action;

H. completion of examination as defined in 37 CFR 41.102; or

I. abandonment of the application.

Telephone inquiries with regard to this decision should be directed to CHERYL GIBSON BAYLOR at (571)272-3213. In his/her absence, calls may be directed to Petition Help Desk at (571) 272-3282.

/CHERYL GIBSON BAYLOR/ Paralegal Specialist, OPET	

U.S. Patent and Trademark Office
PTO-2298 (Rev. 02-2012)

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

APPLICATION NO.	FILING DATE	FIRST NAMED INVENTOR	ATTORNEY DOCKET NO.	CONFIRMATION NO.
16/728,384	12/27/2019	Maged F. Barsoum	C44-00999.CON15	6350

71897 7590 04/09/2020

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER
BOCURE, TESFALDET

ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
04/09/2020	ELECTRONIC

Please find below and/or attached an Office communication concerning this application or proceeding.

The time period for reply, if any, is set in the attached communication.

Notice of the Office communication was sent electronically on above-indicated "Notification Date" to the following e-mail address(es):

PAIR@KPPB.COM

Office Action Summary	**Application No.** 16/728,384	**Applicant(s)** Barsoum et al.	
	Examiner TESFALDET BOCURE	**Art Unit** 2633	**AIA (FITF) Status** No

-- The MAILING DATE of this communication appears on the cover sheet with the correspondence address --

Period for Reply

A SHORTENED STATUTORY PERIOD FOR REPLY IS SET TO EXPIRE 3 MONTHS FROM THE MAILING DATE OF THIS COMMUNICATION.

- Extensions of time may be available under the provisions of 37 CFR 1.136(a). In no event, however, may a reply be timely filed after SIX (6) MONTHS from the mailing date of this communication.
- If NO period for reply is specified above, the maximum statutory period will apply and will expire SIX (6) MONTHS from the mailing date of this communication.
- Failure to reply within the set or extended period for reply will, by statute, cause the application to become ABANDONED (35 U.S.C. § 133). Any reply received by the Office later than three months after the mailing date of this communication, even if timely filed, may reduce any earned patent term adjustment. See 37 CFR 1.704(b).

Status

1)☑ Responsive to communication(s) filed on 12/27/2019.

☐ A declaration(s)/affidavit(s) under **37 CFR 1.130(b)** was/were filed on _____.

2a)☐ This action is **FINAL.** 2b) ☑ This action is non-final.

3)☐ An election was made by the applicant in response to a restriction requirement set forth during the interview on _____; the restriction requirement and election have been incorporated into this action.

4)☐ Since this application is in condition for allowance except for formal matters, prosecution as to the merits is closed in accordance with the practice under *Ex parte Quayle*, 1935 C.D. 11, 453 O.G. 213.

Disposition of Claims*

5) ☑ Claim(s) 1-30 is/are pending in the application.

5a) Of the above claim(s) _____ is/are withdrawn from consideration.

6) ☐ Claim(s) _____ is/are allowed.

7) ☑ Claim(s) 1-30 is/are rejected.

8) ☐ Claim(s) _____ is/are objected to.

9) ☐ Claim(s) _____ are subject to restriction and/or election requirement

* If any claims have been determined allowable, you may be eligible to benefit from the **Patent Prosecution Highway** program at a participating intellectual property office for the corresponding application. For more information, please see http://www.uspto.gov/patents/init_events/pph/index.jsp or send an inquiry to **PPHfeedback@uspto.gov.**

Application Papers

10)☑ The specification is objected to by the Examiner.

11)☑ The drawing(s) filed on 12/27/2019 is/are: a)☑ accepted or b)☐ objected to by the Examiner.

Applicant may not request that any objection to the drawing(s) be held in abeyance. See 37 CFR 1.85(a).

Replacement drawing sheet(s) including the correction is required if the drawing(s) is objected to. See 37 CFR 1.121(d).

Priority under 35 U.S.C. § 119

12)☐ Acknowledgment is made of a claim for foreign priority under 35 U.S.C. § 119(a)-(d) or (f).

Certified copies:

a)☐ All b)☐ Some** c)☐ None of the:

1.☐ Certified copies of the priority documents have been received.

2.☐ Certified copies of the priority documents have been received in Application No. _____.

3.☐ Copies of the certified copies of the priority documents have been received in this National Stage application from the International Bureau (PCT Rule 17.2(a)).

** See the attached detailed Office action for a list of the certified copies not received.

Attachment(s)

1) ☑ Notice of References Cited (PTO-892)

2) ☑ Information Disclosure Statement(s) (PTO/SB/08a and/or PTO/SB/08b) Paper No(s)/Mail Date _____.

3) ☐ Interview Summary (PTO-413) Paper No(s)/Mail Date _____.

4) ☐ Other: _____.

DETAILED ACTION

Notice of Pre-AIA or AIA Status

1. The present application is being examined under the pre-AIA first to invent provisions.

2. This office action (First Action on the Merit) is in response to the application as originally filed on 12/27/2019. The detail office action to the pending claims 1-30 are as shown below.

3. The certification and request for prioritized examination (TrackOne) Under 37 CFR 1.102(e) is noted by the Examiner.

Information Disclosure Statement

4. The information disclosure statements (IDSs) submitted on 01/09/2020 (three IDSs filed on the same date) are in compliance with the provisions of 37 CFR 1.97. Accordingly, the examiner is considering the information disclosure statement. Attached with this correspondence are the initialed copies of the IDSs.

Drawings

5. The drawings were received on 12/27/2019. These drawings are accepted by the Examiner.

Specification

6. The disclosure is objected to because of the following informalities:

In par. [0001], line 2, after "November 30, 2018"---and issued on February 18, 2020 as US patent number 10,567,980---should be added.

Appropriate correction is required.

Remarks

7. The above Examiner's Amendment is required to update the status of the parent US patent application number 16/206,991.

Claim Rejections - 35 USC § 112

8. The following is a quotation of 35 U.S.C. 112(b):

(b) CONCLUSION.—The specification shall conclude with one or more claims particularly pointing out and distinctly claiming the subject matter which the inventor or a joint inventor regards as the invention.

The following is a quotation of 35 U.S.C. 112 (pre-AIA), second paragraph:

The specification shall conclude with one or more claims particularly pointing out and distinctly claiming the subject matter which the applicant regards as his invention.

Claims 22 -30 are rejected under 35 U.S.C. 112(b) or 35 U.S.C. 112 (pre-AIA), second paragraph, as being indefinite for failing to particularly point out and distinctly claim the subject matter which the inventor or a joint inventor, or for pre-AIA the applicant regards as the invention.

Even though the claimed subject matter in claims 22 is not drafted as a means-plus-function, the claimed "receiver" using *a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels* to transform the received signals into received bits" is performing every conceivable claimed limitations in claims 22, therefore, claim 22 is a single means Claim.

The same is true with the claimed subject matter in the dependent claims 23-30, where the same single receiver is performing every conceivable claimed limitations.

Claim Rejections - 35 USC § 112

9. The following is a quotation of the first paragraph of 35 U.S.C. 112(a):

> (a) IN GENERAL.—The specification shall contain a written description of the invention, and of the manner and process of making and using it, in such full, clear, concise, and exact terms as to enable any person skilled in the art to which it pertains, or with which it is most nearly connected, to make and use the same, and shall set forth the best mode contemplated by the inventor or joint inventor of carrying out the invention.

The following is a quotation of the first paragraph of pre-AIA 35 U.S.C. 112:

> The specification shall contain a written description of the invention, and of the manner and process of making and using it, in such full, clear, concise, and exact terms as to enable any person skilled in the art to which it pertains, or with which it is most nearly connected, to make and use the same, and shall set forth the best mode contemplated by the inventor of carrying out his invention.

10. Claims 1-30 rejected under 35 U.S.C. 112(a) or 35 U.S.C. 112 (pre-AIA), first paragraph, as failing to comply with the written description requirement. The claim(s) contains subject matter which was not described in the specification in such a way as to reasonably convey to one skilled in the relevant art that the inventor or a joint inventor, or for pre-AIA the inventor(s), at the time the application was filed, had possession of the claimed invention.

The specification does not support the claimed:

"A demapper, coupled to the demodulator, capable of determining likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels" in claims 1 and 11; and

"A communication system, comprising a receiver that receives signals via a communication channel having a channel signal-to-noise ratio (SNR), wherein the receiver uses *a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels* to transform the received signals into received bits" in claim 22.

The mappers (a receiver claimed in claimed 21) as shown in figures 2, 3, 4a and 4b do not disclose the claimed "determining likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels". The same is true with the claimed "the non-uniform symbol constellation" in claims 2,5,6,8,9,12,15-19, 23, 26, 27, 29 as the claimed "*at least two constellation points having identical locations and different labels*" is referring or further limiting to the once claimed to the "non-uniform constellation" in the independent claims 1, 11 and 21.

Claims 3, 4, 7, 10, 11, 14, 20, 21, 24, 25 and 28 are inherently rejected as being dependent on the rejected base claims.

Claim Rejections - 35 USC § 102

11. The following is a quotation of the appropriate paragraphs of pre-AIA 35 U.S.C. 102 that form the basis for the rejections under this section made in this Office action:

> A person shall be entitled to a patent unless –
>
> (e) the invention was described in (1) an application for patent, published under section 122(b), by another filed in the United States before the invention by the applicant for patent or (2) a patent granted on an application for patent by another filed in the United States before the invention by the applicant for patent, except that an international application filed under the treaty defined in section 351(a) shall have the effects for purposes of this subsection of an application filed in the United States

only if the international application designated the United States and was published under Article 21(2) of such treaty in the English language.

12. Claim(s) 1, 10, 11, 20, 21 and 22 are rejected under pre-AIA 35 U.S.C. 102(e) as being anticipated by Eberlein et al., Eberlein hereinafter (US patent publication number 2008/0200114-A1).

Eberlein teaches a communication system (see figures 11 and 12), comprising:

a receiver (figure 14) capable of receiving signals via a communication channel having a channel signal-to-noise ratio (SNR) ([0014], [0030], [0064] and [0162]), wherein the receiver comprises:

a demodulator (830 in figure 14) capable of demodulating a received signal into a demodulated signal;

a demapper (840), coupled to the demodulator, capable of determining likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels (see figures 3A and 3B, where the signal points located in the first quadrant shown in figure 3A, claimed identical location, having different labelling 0010, 0000, 001 and 0011); and

a decoder (860 and 866), coupled to the demapper, capable of using the likelihoods determined by the demapper to provide a sequence of received bits based upon a low density parity check (LDPC) code (see elements 860 and 866 in figure 14, where the decoder can be LPDC as disclosed in par. [0150]), as in claim 1.

Further to the independent claims, 11 and 22, Eberlein also teaches:

A communication system (see figures 11 and 12), comprising:

a receiver (figure 14) that receives signals via a communication channel having a channel signal-to-noise ratio (SNR) ([0014], [0030], [0064] and [0162]),, wherein the receiver comprises:

a demodulator that demodulates a received signal into a demodulated signal;

a demapper (840) that determines likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels see figures 3A and 3B, where the signal points in the first quadrant shown in figure 3A, claimed identical location, having different labelling 0010, 0000, 0001 and 0011); and

a decoder (860 and 866) that uses the likelihoods determined by the demapper to provide a sequence of received bits based upon a low density parity check (LDPC) code (see elements 860 and 866 in figure 14, where the decoder can be LPDC as disclosed in par. [0150]);

wherein the demapper (840)is interposed between the demodulator, the decoder, and the demapper receives information from the demodulator and provides information to the decoder, as in claim 11.

A communication system (see figures 11 and 12), comprising a receiver (figure 14) that receives signals via a communication channel having a channel signal-to-noise ratio (SNR) ([0014], [0030], [0064] and [0162]), wherein the receiver uses a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels to transform the received signals into received bits see figures 3A and 3B, where the signal points in the first quadrant shown in figure 3A, claimed identical location, having different labelling 0010, 0000, 0001 and 0011), as in claim 22.

Further to the dependent claims 10, 20 and 21, Eberlein also teaches:

The communication system of claim 1, further comprising a transmitter (figure 13) capable of transmitting signals via the communication channel, where the transmitter comprises:

a coder (710) capable of receiving bits and outputting encoded bits using a Low Density Parity Check (LDPC) code;

a mapper (760), coupled to the coder, capable of mapping the encoded bits to symbols in the non-uniform symbol constellation; and

a modulator (770), coupled to the mapper, capable of generating a signal for transmission via the communication channel based upon symbols selected by the mapper, as in claim 10.

The communication system of claim 11, further comprising a transmitter (figure 13) that transmits signals via the communication channel, where the transmitter comprises:

a coder (714 and 732) that receives bits and outputs encoded bits using a Low Density Parity Check (LDPC) code;

a mapper (760) that maps the encoded bits to symbols in the selected symbol constellation; and

a modulator (770) that generates a signal for transmission via the communication channel based upon symbols selected by the mapper;

wherein the mapper (760) is interposed between the modulator (770) and the coder(710) and the mapper receives information from the coder and provides information to the modulator, as in claim 20; and

The communication system of claim 11, further comprising a transmitter (figure 13) that transmits signals via the communication channel, where the transmitter uses the non-

uniform symbol constellation to transform encoded bits into the transmitted signals (see abstract of the disclosure for Hierarchal modulation, claimed non-uniform symbol constellation and as shown in figure 3A and 3B), as in claim 21.

Claim Rejections - 35 USC § 103

12. The following is a quotation of pre-AIA 35 U.S.C. 103(a) which forms the basis for all obviousness rejections set forth in this Office action:

> (a) A patent may not be obtained though the invention is not identically disclosed or described as set forth in section 102, if the differences between the subject matter sought to be patented and the prior art are such that the subject matter as a whole would have been obvious at the time the invention was made to a person having ordinary skill in the art to which said subject matter pertains. Patentability shall not be negatived by the manner in which the invention was made.

The factual inquiries set forth in *Graham v. John Deere Co.*, 383 U.S. 1, 148 USPQ 459 (1966), that are applied for establishing a background for determining obviousness under pre-AIA 35 U.S.C. 103(a) are summarized as follows:

1. Determining the scope and contents of the prior art.

2. Ascertaining the differences between the prior art and the claims at issue.

3. Resolving the level of ordinary skill in the pertinent art.

4. Considering objective evidence present in the application indicating obviousness or nonobviousness.

This application currently names joint inventors. In considering patentability of the claims under pre-AIA 35 U.S.C. 103(a), the examiner presumes that the subject matter of the various claims was commonly owned at the time any inventions covered therein were made absent any evidence to the contrary. Applicant is advised of the obligation under 37 CFR 1.56 to point out the inventor and invention dates of each claim that was not commonly owned at

the time a later invention was made in order for the examiner to consider the applicability of pre-AIA 35 U.S.C. 103(c) and potential pre-AIA 35 U.S.C. 102(e), (f) or (g) prior art under pre-AIA 35 U.S.C. 103(a).

13. Claims 9, 19 and 30 are rejected under pre-AIA 35 U.S.C. 103(a) as being unpatentable over Eberlein et al., Eberlein hereinafter (US patent publication number 2008/0200114-A1) in view of Walker et al., Walker hereinafter, (US patent publication number 2005/011581-A1).

Eberlein teaches the claimed subject matter in the independent claims 1,11 and 21 in which the dependent claimed 9, 19 and 30 respectively depends. However, Eberlein fails to teach the claimed wherein at least one of the plurality of non-uniform symbol constellations is an orthogonal non-uniform Pulse Amplitude Modulated (PAM) constellation in claims 9, 19 ad 30.

Walker for the same endeavor as the instant application and that of Eberlein teaches a communication system for communicating a digitally modulated non-uniformly modulated signal (see figure 6) compromising modulating the digitally modulated two-dimensional PSK and QAM signals using orthogonal PAM signal (see figure 6 and par. [0032], [0041] and [004]).

To use the teaching of Walker in the system of Eberlein would enable the system to minimize intersymbol interference between the symbols transmitted.

Therefore, it would have been obvious to one of an ordinary skill in the art to use the teaching of Walker at the time the invention was made.

Conclusion

13. The prior art made of record and not relied upon is considered pertinent to applicant's disclosure.

- US patent numbers 7,660,368 and 8,675,754 issued to Walker, Ling and Yonge respectively and US patent publication numbers 2005/0089068, 2007/0111581, and 2007/0195868 and 2007/0104293 published to Sun, Walker et al. and Hiatt respectively teaches a communication system having a transmitter having encoder and decoder for LPDC encoding and decoding the source signal using and non-uniform.

14. Any inquiry concerning this communication or earlier communications from the examiner should be directed to TESFALDET BOCURE whose telephone number is (571)272-3015. The examiner can normally be reached on M-to-F, 9:00am-5:30pm.

Examiner interviews are available via telephone, in-person, and video conferencing using a USPTO supplied web-based collaboration tool. To schedule an interview, applicant is encouraged to use the USPTO Automated Interview Request (AIR) at http://www.uspto.gov/interviewpractice.

If attempts to reach the examiner by telephone are unsuccessful, the examiner's supervisor, Sam K AHN can be reached on 571-272-3044. The fax phone number for the organization where this application or proceeding is assigned is 571-273-8300.

Information regarding the status of an application may be obtained from the Patent Application Information Retrieval (PAIR) system. Status information for published applications may be obtained from either Private PAIR or Public PAIR. Status information for unpublished

applications is available through Private PAIR only. For more information about the PAIR system, see http://pair-direct.uspto.gov. Should you have questions on access to the Private PAIR system, contact the Electronic Business Center (EBC) at 866-217-9197 (toll-free). If you would like assistance from a USPTO Customer Service Representative or access to the automated information system, call 800-786-9199 (IN USA OR CANADA) or 571-272-1000.

/TESFALDET BOCURE/
Primary Examiner, Art Unit 2633

Notice of References Cited	Application/Control No. 16/728,384	Applicant(s)/Patent Under Reexamination Barsoum et al.	
	Examiner TESFALDET BOCURE	Art Unit 2633	Page 1 of 1

U.S. PATENT DOCUMENTS

*		Document Number Country Code-Number-Kind Code	Date MM-YYYY	Name	CPC Classification	US Classification
*	A	US-7215713-B2	05-2007	Walker; Glenn A.	H04B7/18513	267/246
*	B	US-7660368-B2	02-2010	Ling; Fuyun	H04L1/0045	341/143
*	C	US-8675754-B1	03-2014	Yonge, III; Lawrence Winston	H04L5/12	375/261
*	D	US-20050089068-A1	04-2005	Sun, Feng-Wen	H03M13/1165	370/509
*	E	US-20050111581-A1	05-2005	Walker, Glenn A.	H04B7/18513	375/308
*	F	US-20070195868-A1	08-2007	Walker; Glenn A.	H04B7/18513	375/211
*	G	US-20070104293-A1	05-2007	Hiatt; Michael L. JR.	H04L27/3488	375/329
*	H	US-20080200114-A1	08-2008	Eberlein; Ernst	H04H20/06	455/3.02
	I					
	J					
	K					
	L					
	M					

FOREIGN PATENT DOCUMENTS

*		Document Number Country Code-Number-Kind Code	Date MM-YYYY	Country	Name	CPC Classification
	N					
	O					
	P					
	Q					
	R					
	S					
	T					

NON-PATENT DOCUMENTS

*		Include as applicable: Author, Title Date, Publisher, Edition or Volume, Pertinent Pages)
	U	
	V	
	W	
	X	

*A copy of this reference is not being furnished with this Office action. (See MPEP § 707.05(a).)
Dates in MM-YYYY format are publication dates. Classifications may be US or foreign.

Index of Claims	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

✓	Rejected	-	Cancelled	N	Non-Elected	A	Appeal
=	Allowed	÷	Restricted	I	Interference	O	Objected

CLAIMS

☐ Claims renumbered in the same order as presented by applicant ☐ CPA ☐ T.D. ☐ R.1.47

CLAIM		DATE								
Final	Original	04/03/2020								
	1	✓								
	2	✓								
	3	✓								
	4	✓								
	5	✓								
	6	✓								
	7	✓								
	8	✓								
	9	✓								
	10	✓								
	11	✓								
	12	✓								
	13	✓								
	14	✓								
	15	✓								
	16	✓								
	17	✓								
	18	✓								
	19	✓								
	20	✓								
	21	✓								
	22	✓								
	23	✓								
	24	✓								
	25	✓								
	26	✓								
	27	✓								
	28	✓								
	29	✓								
	30	✓								

Search Notes	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

CPC - Searched*		
Symbol	**Date**	**Examiner**
(H04W24/02|H04B15/00|H04L1/0003|H04L1/0009|H04L27/3405| H04L27/3483).CPC.	04/02/2020	TB
CPC on EAST (H04L27/3405|H04L1/0003|H04L1/0009|H04L1/0002| H04L1/0003|H04W24/02|H04B15/00|H04L1/0003|H04L1/0009| H04L27/3405|H04L27/3483).CPC.	04/02/2020	TB
CPC updated	04/05/2020	TB

CPC Combination Sets - Searched*		
Symbol	**Date**	**Examiner**

US Classification - Searched*			
Class	**Subclass**	**Date**	**Examiner**
375, 329, 332	(375/261.ccls. or 375/269.ccls. or 375/279.ccls. or 375/295.ccls. or 375/308.ccls. or 375/329.ccls. or 375/332.ccls. or 375/358.ccls. or 329/103.ccls. or 332/104.ccls. 329/304.ccls. 329/305.ccls.)	04/02/2020	TB
375, 455, 329, 332, 370	Searched ALL, (375/$8.ccls. or 455/$8.ccls. or 370/ $8.ccls. or 714/$8.ccls. or 329/$8.ccls. or 332/$8.ccls .)	04/02/2020	TB
375, 455, 329, 332, 370	Search updated ALL, see EAST search result	04/05/2020	TB
375, 455, 329, 332, 370	Search updated ALL, see WEST search result	04/06/2020	TB

* See search history printout included with this form or the SEARCH NOTES box below to determine the scope of the search.

Search Notes	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

Search Notes		
Search Notes	**Date**	**Examiner**
EAST and Inventor's Name searched (unable to download the search result)	04/02/2020	TB
Google Scholar searched	04/05/2020	TB
EAST updated and IP5 and Global Dossier Searched	04/05/2020	TB
WEST searched	04/06/2020	TB
See related US patent application S/Ns 16/726,065; 16/724,047, 16/ 729,216, 16/729,226, 16/206,991, 15/ 682,475, 15/200,800, 14/ 491,731, 13/618,630, 13/118,921, 12/ 156,989 and 16/726,226	04/02/2020	TB

Interference Search			
US Class/CPC Symbol	**US Subclass/CPC Group**	**Date**	**Examiner**

EAST Search History

EAST Search History (Prior Art)

Ref #	Hits	Search Query	DBs	Default Operator	Plurals	Time Stamp
L1	306	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L2	0	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and (((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (QAM or (quadrature near2 amplitude near3 modulat$4) or PSK or (phase near2 shift near2 keying near2 modulation)) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L3	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L4	421	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L5	7	L4 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L6	328	L4 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12

L7	4	L6 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L8	5	L6 and L1	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L9	4	L6 and (375/261.ccls. or 375/269.ccls. or 375/279.ccls. or 375/295.ccls. or 375/308.ccls. or 375/329.ccls. or 375/332.ccls. or 375/358.ccls. or 329/103.ccls. or 332/104.ccls. 329/304.ccls. 329/305.ccls.)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L10	43	L6 and (375/$8.ccls. or 455/$8.ccls. or 370/$8.ccls. or 714/$8.ccls. or 329/$8.ccls. or 332/$8.ccls.)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
L11	154	("10149179" | "10524139" | "10530629" | "20020044597" | "20020106010" | "20030231715" | "20040022179" | "20040066738" | "20040161050" | "20040258177" | "20050180531" | "20050207507" | "20050276343" | "20050286409" | "20060045169" | "20060144843" | "20060155843" | "20070022179" | "20070025283" | "20070054614" | "20070116161" | "20070147530" | "20070280147" | "20080200114" | "20090097582" | "20090161786" | "20100195743" | "20100303174" | "20110090948" | "20110228869" | "20110305300" | "20120147983" | "20130083862" | "20130170571" | "5289501" | "5862179" | "6084915" | "6115415" | "6157678" | "6603801" | "6606355" | "6611554" | "6665831" | "7212582" | "7245666" | "7376203" | "7620067" | "7978777" | "8265175" | "8270511" | "8842761" | "9191148" | "9385832" | "9743290" | "9743292" | "9887870" | "9967127").PN.	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:14
L12	4	L6 and L11	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:14
S1	306	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.)	US-PGPUB; USPAT; USOCR;	OR	ON	2020/04/05 20:28

			FPRS; EPO; JPO; DERWENT; IBM_TDB			
S2	0	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and (((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (QAM or (quadrature near2 amplitude near3 modulat$4) or PSK or (phase near2 shift near2 keying near2 modulation)) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:29
S3	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:30
S4	421	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:33
S5	7	S4 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:34
S6	328	S4 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:41
S8	4	S6 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:42
S9	5	S6 and S1	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:50

S10	4	S6 and (375/261.ccls. or 375/269.ccls. or 375/279.ccls. or 375/295.ccls. or 375/308.ccls. or 375/329.ccls. or 375/332.ccls. or 375/358.ccls. or 329/103.ccls. or 332/104.ccls. 329/304.ccls. 329/305.ccls.)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:51
S11	43	S6 and (375/$8.ccls. or 455/$8.ccls. or 370/$8.ccls. or 714/$8.ccls. or 329/$8.ccls. or 332/$8.ccls.)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 20:51

4/5/2020 10:22:08 PM
C:\Users\tbocure\Documents\EAST\Workspaces\16728384.wsp

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned 2633
	Examiner Name	Unassigned Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

U.S.PATENTS

Examiner Initial*	Cite No	Patent Number	Kind Code 1	Issue Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		5289501		1994-02-22	Seshadri et al.	
		5862179		1999-01-19	Goldstein et al.	
		6084915	A	2000-07-04	Williams et al.	
		6115415		2000-09-05	Goldstein et al.	
		6157678		2000-12-05	Wei	
		6603801	B1	2003-08-05	Andren et al.	
		6606355		2003-08-12	Wei	
		6611554		2003-08-26	Chouly et al.	
		6665831		2003-12-16	Yoshida et al.	
		7212582	B2	2007- 05-01	Zhang et al.	
		7245666	B1	2007-07-17	Gardner, James et al.	
		7376203		2008- 05-20	Brunel et al.	
		7620067		2009-11-17	Niu et al.	
		7978777	B2	2011-07-12	Barsoum et al.	
		8265175		2012-09-11	Barsoum, M. F., et al.	
		8270511		2012-09-18	Barsoum, M F., et al.	
		8842761		2014-09-23	Barsoum, Maged F., et al.	
		9191148		2015-11-17	Barsoum, Maged F.	
		9385832		2016-07-05	Barsoum et al., Maged	
		9743290	B2	2017-08-22	Barsoum et al.	
		9743292	B2	2017-08-22	Barsoum et al.	
		9887870	B2	2018-02-06	Barsoum et al.	
		9967127	B1	2018- 05-08	Sun et al.	
		10149179	B2	2018-12-04	Barsoum et al.	
		10524139		2019-12-31	Barsoum et al.	
		10530629		2020-01-07	Barsoum et al.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

U.S.PATENT APPLICATION PUBLICATIONS

Examiner Initial*	Cite No	Publication Number	Kind Code [1]	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		20020044597	A1	2002-04-18	Shively	
		20020106010	A1	2002-08-08	Jones	
		20030231715	A1	2003-12-18	Shoemake, Matthew B.	
		20040022179	A1	2004-02-05	Giannakis et al.	
		20040066738	A1	2004-04-08	Stopler, Daniel	
		20040161050	A1	2004-08-19	Larsson et al.	
		20040258177	A1	2004-12-23	Shen, Ba-zhong et al.	
		20050180531		2005-08-18	Wellig et al.	
		20050207507	A1	2005-09-22	Mitsutani, Naoki	
		20050276343	A1	2005-12-15	Jones	
		20050286409	A1	2005-12-29	Yoon et al.	
		20060045169	A1	2006-03-02	Kim, Byoung-hoon	
		20070022179	A1	2007-01-25	Kim et al.	
		20070116161	A1	2007- 05-24	Tokoro et al.	
		20060144843	A1	2006-07-06	Vandal et al.	
		20060155843	A1	2006-07-13	Glass et al.	
		20070025283	A1	2007-02-01	Koslov et al.	
		20070054614	A1	2007-03-08	Walker et al.	
		20070147530	A1	2007-06-28	Li	
		20070280147	A1	2007-12-06	Catreux-erceg, Severine et al.	
		20080200114	A1	2008-08-21	Eberlein et al.	
		20090097582	A1	2009-04-16	Barsoum et al., F.	
		20090161786	A1	2009-06-25	Nakagawa, Takao et al.	
		20100195743	A1	2010-08-05	Barsoum, M F., et al.	
		20100303174	A1	2010-12-02	Oh, Min Seok et al.	
		20110090948	A1	2011-04-21	Zhou et al.	
		20110228869	A1	2011-09-22	Barsoum, M et al.	
		20110305300	A1	2011-12-15	Ko	
		20120147983	A1	2012-06-14	Barsoum, M F., et al.	
		20130083862	A1	2013-04-04	Barsoum, Maged F., et al.	
		20130170571	A1	2013-07-04	Barsoum, Maged F., et al.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		20140314177	A1	2014-10-23	Choi et al.	
		20150236812	A1	2015-08-20	Barsoum, Maged F., et al.	
		20160316382		2015-10-27	Barsoum et al., Maged	
		20160204967	A1	2016-07-14	Choi et al.	
		20160309342	A1	2016-10-20	Barsoum et al.	
		20170374564	A1	2017-12-28	Barsoum et al.	
		20170374565	A1	2017-12-28	Barsoum et al.	
		20180102933	A1	2018-04-12	Schneider et al.	
		20180191548	A1	2018-07-05	Barsoum et al.	
		20190116509	A1	2019-04-18	Barsoum et al.	
		20190342772	A1	2019-11-07	Barsoum et al.	

FOREIGN PATENT DOCUMENTS

Examiner Initial*	Cite No	Foreign Document Number[3]	Country Code[2]	Kind Code[4]	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines where Relevant Passages or Relevant Figures Appear	T[5]
		09130438	JP	A	1997- 05-16	Tomabechi, Akitaka et al.		
		9832257	WO		1998-07-23	Motorola, Inc.		
		2003229835	JP		2003-08-15	Mitsuhiro, Suzuki		
		1578021	EP	A1	2005-09-21	Mitsutani, Naoki		
		2005269258	JP	A	2005-09-29	Mitsuya, Naoki		
		2007074524	WO	A1	2007-07-05	Nakagawa, Takao et al.		
		1971098	EP	A1	2008-09-17	Nakagawa, Takao et al.		
		2008151308	WO	A1	2008-12-11	Barsoum, Maged et al.		
		2010078472	WO	A1	2010-07-08	Barsoum et al.		
		2010538502	JP	A	2010-12-09	Barsoum et al.		
		5513377	JP	B2	2014-06-04	Barsoum et al.		
		2153561	EP	B1	2019-01-09	Barsoum et al.		
		201905158	TR	T4	2019- 05-21	Barsoum et al.		
		3518485	EP	A1	2019-07-31	Barsoum et al.		

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T 5
		European Supplementary Search Report for Application No. EP 08795885, International Filing Date June 5, 2008, Search Completed April 1, 2014, 8 pgs.	
		Extended European Search Report for European Application No. 18212572.4, Search completed May 27, 2019, Mailed June 06, 2019, 13 Pgs.	
		International Preliminary Report on Patentability for International Application No. PCT/US/2008/065994, Report Completed June 18, 2009, Report Mailed June 30, 2009, 4 pgs.	
		International Preliminary Report on Patentability for International Application No. PCT/US2009/069881, Report Issued January 7, 2011, Report Mailed January 14, 2011, 11 pgs.	
		International Search Report for International Application No. PCT/US 09/69881, date completed April 12, 2010, date mailed May 3, 2010, 2 pgs.	
		International Search Report for International Application No. PCT/US2008/065994, Report completed October 3, 2008, mailed October 22, 2008, 2 pgs.	
		Written Opinion for International Application No. PCT/US2008/065994, completed October 3, 2008, mailed October 22, 2008, 5 pgs.	
		Written Opinion of the International Searching Authority for International Application No. PCT/US 09/69881, date completed April 13, 2010, date mailed May 3, 2010, 10 pgs.	
		"ETSI EN 302 307 V1.1.1, European Standard (Telecommunications series), Digital Video Braodcasting (DVB); Second generation framing structure,", channel coding and modulation systems for Broadcasting, Interactive Services, News Gathering and other broadband satellite applications, 2005-03, Retrieved from: http://www.etsi.org/deliver/etsi_en/302300_302399/302307/01.01.01_60/en_302307v010101p.pdf.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		13/618,630, "Notice of Allowance Received", May 15, 2014, 10 pgs.	
		AGRELL ET AL., "Gray Coding for Multilevel Constellations in Gaussian Noise", To appear in IEEE Transactions on Information Theory, Vol. 53, Issue 1, January 2007, Published December 26, 2006, 14 pgs.	
		AGRELL ET AL., "On optimal constellations for BICM at low SNR", Information Theory Workshop, 2009. ITW 2009. IEEE, Conference: October 11-16, 2009, 6 pgs.	
		AGRELL ET AL., "On the BICM Capacity", ArXiv.org, December 8, 2010, 53 pgs.	
		AGRELL ET AL., "On the Optimality of the Binary Reflected Gray Code", IEEE Transactions on Information Theory, December 2004, Vol. 50, No. 12, pp. 3170-3182.	
		AGRELL ET AL., "Optimal Signal Sets and Binary Labelings for BICM at low SNR", To Appear in IEEE Transactions on Information Theory, Vol. 57, Issue 10, October 2011, Published October 6, 2011, 24 pgs.	
		ALVARADO ET AL., "On the BICM Capacity—Part I: Binary Labelings, Arbitrary Input Distributions, and First-Order Asymptotics", Arxiv.org, January 25, 2010, 45 pgs.	
		ALVARADO ET AL., "On the capacity of BICM with QAM constellations (Invited Paper)", Retrieved from: http://publications.lib.chalmers.se/records/fulltext/local_92655.pdf, IWCMC'09, June 21-24, 2009, Leipzig, Germany, 8 pgs.	
		ARAFA ET AL., "Non-Uniform Signal Constellation for Iteratively Decoded Bit Interleaved Coded Modulation (BICM-ID) with Convolution and LDPC Codes", 2011 International Conference on Innovations in Information Technology, Conference: April 25-27, 2011, pp. 23-28.	
		BARSOUM, "On Constellation Design and Iterative Codes", University of California, Los Angeles, Dissertation, 2008, 119 pgs.	
		BARSOUM ET AL., "Constellation Design via Capacity maximization", IEEE International Symposium on Information Theory, 24 June 2007, pp.1821-1825.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		BATSHON ET AL., "Iterative Polar Quantization-Based Modulation to Achieve Channel Capacity in Ultrahigh-Speed Optical Communication Systems", IEEE Photonics Journal, Vol 2, No. 4, pp 593-599 August 2010.	
		BETTS ET AL., "Performance of Nonuniform Constellations on the Gaussian Channel", IEEE Transactions on Information Theory, September 1994, Vol. 40, No. 5, pp. 1633-1638.	
		CHEN ET AL., "Increasing achievable information rates via geometric shaping", arXiv.org, e-Print Archive, Physics, arXiv:1804.08850v1, April 24, 2018.	
		CHOI ET AL., "Channel Capacity Enhancement Scheme for Satellite Communication System", 2007 6th International Conference on Information, Communications & Signal Processing, Conference: December 10-13, 2007, 4 pgs.	
		CHOI ET AL., "Satellite-DMB applied to the 8PSK hierarchical modulation", Digest of Technical Papers. International Conference on Consumer Electronics, 2007. ICCE 2007, Conference: January 10-14, 2007, 2 pgs.	
		CONWAY ET AL., "A Fast Encoding Method for Lattice Codes and Quantizers", IEEE Transactions on Information Theory, November 1983, Vol. IT-29, No. 6, pp. 820-824.	
		DE GAUDENZI ET AL., "Performance Analysis of Turbo-Coded APSK Modultions Over Nonlinear Satellite Channels", IEEE Transactions of Wireless Communications, September 2006, Vol. 5, No. 5, pp. 2396-2407.	
		DE GAUDENZI ET AL., "APSK Coded Modulation Schemes for Nonlinear Satellite Channels with High Power and Spectral Efficiency", American Institute of Aeronautics and Astronautics Paper AIAA, 2002-1861, 2002, Retrieved from: http://www.dtic.upf.edu/~aguillen/home_upf/Publications_files/aiaa2001.pdf.	
		DE GAUDENZI ET AL., "Performance Analysis of Turbo-Coded APSK Modulations over Nonlinear Satellite Channels", IEEE Transactions on Wireless Communications, Vol. 5, No. 9, September 6, 2006,1536-1276,12 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		DE GAUDENZI ET AL., "Turbo-Coded APSK Modulations Design for Satellite Broadband Communications", International Journal of Satellite Communications and Networking, Vol. 24, No. 4, July 1, 2006, pp. 261-281.	
		FABREGAS ET AL., "Bit-Interleaved Coded Modulation", Foundations and Trends® in Communications and Information Theory: Vol. 5: No. 1–2, pp 1-153. http://dx.doi.org/10.1561/0100000019 , Published: 30 Nov 2008.	
		FABREGAS ET AL., "Bit-Interleaved Coded Modulation with Shaping", IEEE Information Theory Workshop (ITW), 2010, Conference: August 30 - September 3, 2010, 6 pgs.	
		FABREGAS ET AL., "Error Probability of Bit-Interleaved Coded Modulation using the Gaussian Approximation", 2004 Conference on Information Sciences and Systems, Princeton University, March 17–19, 2004, 7 pgs.	
		FABREGAS ET AL., "Impact of Signal Constellation Expansion on the Achievable Diversity of Pragmatic Bit-interleaved Space-Time Codes", IEEE Transactions on Wireless Communications, Vol. 5, Issue 8, August 2006, 15 pgs.	
		FARID, ET AL., "Design of Non-Uniform Capacity-Approaching Signaling for Optical Wireless Intensity Channels", IEEE International Symposium on Information Theory, 2008, Conference: July 6-11, 2008, pp. 2327-2331.	
		FARID ET AL., "Channel Capacity and Non-Uniform Signalling for Free-Space Optical Intensity Channels", IEEE Journal on Selected Areas in Communications, Vol. 27, No. 9, December 2009, 12 pgs.	
		FORNEY, ET AL., "Multidimensional Constellations - Part II: Voronoi Constellations", IEEE Journal on Selected Areas in Communications, August 1989, Vol. 7, No. 6, pp. 941-958.	
		FORNEY, JR. ET AL., "Efficient Modulation for Band-Limited Channels", IEEE Journal on Selected Areas in Communications, September 1984, Vol. SAC-2, No. 5, pp. 632-647.	
		FORNEY, JR. ET AL., "Multidimensional Constellations - Part I: Introduction, Figures of Merit, and Generalized Cross Constellations", IEEE Journal on Selected Areas in Communication, August 1989, Vol. 7, No. 6, pp. 877-892.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		FOSCHINI ET AL., "Optimization of Two-Dimensional Signal Constellations in the Presence of Gaussian Noise", IEEE Transactions on Communications, January 1974, Vol. Com-22, No. 1, pp. 28-38.	
		FRAGOULI ET AL., "Serially Concatenated Coding for Broadcasting S-UMTS Applications", IEEE 7th Int. Symp. on Spread-Spectrum Tech. & Appl., Prague, Czech Republic, September 2-5, 2002, pp. 697-701.	

EXAMINER SIGNATURE

Examiner Signature	/TESFALDET BOCURE/	Date Considered	03/28/2020

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

WEST Search History for Application 16728384

Creation Date: 2020040612:04

Prior Art Searches

Query	DB	Hits	Op.	Plur.	Thes.	Date
(((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (orthogonal$4 same (PAM or (pulse adj2 amplitude adj2 modulat$4)))) and @ad<=20070605)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	22	OR	YES		04-06-2020
(((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (orthogonal$4 same (PAM or (pulse adj2 amplitude adj2 modulat$4)))) and (parallel near4 decod$4) and @ad<=20070605)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	0	OR	YES		04-06-2020
((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	306	OR	YES		04-06-2020
((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	5	OR	YES		04-06-2020
(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2	PGPB, USPT, USOC, EPAB, JPAB, DWPI,	328	OR	YES		04-06-2020

amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding)	TDBD, FPRS					
(((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (orthogonal$4 same (PAM or (pulse adj2 amplitude adj2 modulat$4)))) and (parallel near4 decod$4) and @ad<=20070605)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	0	OR	YES		04-06-2020
(((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and (orthogonal$4 same (PAM or (pulse adj2 amplitude adj2 modulat$4)))) and @ad<=20070605)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	22	OR	YES		04-06-2020
(((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.)) and ((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding))	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	5	OR	YES		04-06-2020
((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding)) and (375/261.ccls. or 375/269.ccls. or 375/279.ccls. or 375/295.ccls. or 375/308.ccls. or 375/329.ccls. or 375/332.ccls. or 375/358.ccls. or 329/103.ccls. or 332/104.ccls. 329/304.ccls. 329/305.ccls.)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	4	OR	YES		04-06-2020

((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding)) and (375/$8.ccls. or 455/$8.ccls. or 370/$8.ccls. or 714/$8.ccls. or 329/$8.ccls. or 332/$8.ccls.)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	43	OR	YES		04-06-2020
((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding)) and (H04W24/02 | H04B15/00 | H04L1/0003 | H04L1/0009 | H04L27/3405 | H04L27/3483 | H03M13/255 | H03M13/1165 | H04L1/0071).CPC.	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	313	OR	YES		04-06-2020
((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((signal near2 noise near2 ratio) or SNR) and ((PSK or (phase near2 shift near2 keying near2 modulation)) or (QAM or (quadrature near2 amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code)) and (demapper or demapping) and (decoder or decoding)) and (H04L27/3405 | H04L1/0003 | H04L1/0009 | H04L1/0002 | H04L1/0003 | H04W24/02 | H04B15/00 | H04L1/0003 | H04L1/0009 | H04L27/3405 | H04L27/3483).CPC.	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	27	OR	YES		04-06-2020

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned 2633
	Examiner Name	--Unassigned Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T 5
		FRAGOULI ET AL., "Turbo Codes with Non-Uniform Constellations", IEEE Int. Conf. Commun., June 2001, pp. 70-73.	
		KSCHISCHANG ET AL., "Optimal Nonuniform Signaling for Gaussian Channels", IEEE, pp. 913-929, 1993.	
		GOFF ET AL., "Channel capacity of bit-interleaved coded modulation schemes using 8-ary signal constellations", Electronics Letters, Vol. 38, Issue 4, Feb 14, 2002, pp. 187-188.	
		GOFF ET AL., "Signal Constellations for Bit-Interleaved Coded Modulation", IEEE Transactions on Information Theory, Vol. 49, Issue 1, January 2003, pp. 307-313.	
		HAMKINS ET AL., "Asymptotically Dense Spherical Codes - Part I: Wrapped Spherical Codes", IEEE Transactions on Information Theory, November 1997, Vol. 43, No. 6, pp. 1774-1785.	
		HAMKINS ET AL., "Asymptotically Dense Spherical Codes - Part II: Laminated Spherical Codes", IEEE Transactions on Information Theory, November 1997, Vol. 43, No. 6, pp. 1786-1798.	
		HOSSAIN ET AL., "BICM Transmission using Non-Uniform QAM Constellations: Performance Analysis and Design", IEEE International Conference on Communications (ICC), 2010, Conference: May 23- 27, 2010, 7 pgs.	
		HOSSAIN ET AL., "Constellation and Interleaver Design for BICM", Global Telecommunications Conference (GLOBECOM 2011), 2011 IEEE, Conference: December 5-9, 2011, 7 pgs.	
		HOSSAIN ET AL., "Towards Fully Optimized BICM Transceivers", Arxiv.org, December 8, 2010, 31 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		ISAKA ET AL., "Error performance analysis of multilevel coded asymmetric 8-PSK modulation with multistage decoding and unequal error protection", 1998 IEEE INTERNATIONAL SYMPOSIUM ON CAMBRIDGE, MA, USA 16-21 AUG, 16 August 1998 page 210.	
		JO ET AL., "An Advanced Hierarchical Modulation with Rotated Constellation", The 12th International Conference on Advanced Communication Technology (ICACT), 2010, Conference: February 7-10, 2010, pp. 515-518.	
		KAYHAN ET AL., "Joint Signal-Labeling Optimization for Pragmatic Capacity under Peak-Power Constraint", Global Telecommunications Conference (GLOBECOM 2010), 2010 IEEE, Conference December 6-10, 2010, Miami, FL, USA, 6 pgs.	
		KHANDANI ET AL., "Application of Shaping Technique to Multi-level Turbo-coded Modulation", 2002, 5 pgs.	
		KHANDANI ET AL., "Application of Shaping Technique to Multi-level Turbo-coded Modulation", University of Waterloo, Department of Electrical and Computer Engineering, Waterloo, Ontario, Canada, Technical Report UW-E&CE#02-07, May 30, 2002, 24 pgs.	
		KHOO ET AL., "Bit Interleaved Coded Modulation with Iterative Decoding Using Constellation Shaping", IEEE Transactions on Communications, October 2006, 5 pgs.	
		LEE ET AL., "Hierarchical Constellation Based Adaptive Relay Scheme in Multi-Hop Networks", IEEE Communication Letters, Vol. 11, No. 3, March 2007, pp. 225-227.	
		LIOLIS ET AL., "Amplitude Phase Shift Keying Constellation Design and its Applications to Satellite Digital Video Broadcasting", Retrieved from: http://www.dtic.upf.edu/~aguillen/home_upf/Publications_files/apsk_chapter.pdf, June 1, 2009, 28 pgs.	
		LIU ET AL., "APSK Constellation with Gray Mapping", IEEE Communications Letters, Vol. 15, Issue 12, December 2011, pp.127-1273, Date of Publication October 31, 2011.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		LOGHIN ET AL., "Non-Uniform Constellations for ATSC 3.0", IEEE Transactions on Broadcasting, Vol. 62, No. 1, March 2016, pp. 197-203.	
		LONG DUAN ET AL., "Approaching the AWGN Channel Capacity without Active Shaping", Proceedings of the International Symposium on Information Theory, pp. 374, 1997.	
		MA ET AL., "Coded Modulation Using Superimposed Binary Codes", IEEE Transactions of Information Theory, Vol. 50, No. 12, December 2004, pp. 3331-3343.	
		MAKOWSKI, "On the Optimality of Uniform Pulse Amplitude Modulation", IEEE Transactions on information Theory, December 2006, Vol. 52, No. 12, pp. 5546-5549.	
		MARTINEZ ET AL., "Bit-Interleaved Coded Modulation in the Wideband Regime", Retrieved from: https://arxiv.org/pdf/0710.4046.pdf, Draft, October 22, 2007, 23 pgs.	
		MARTINEZ ET AL., "Coding and Modulation for the Additive Exponential Noise Channel", IEEE International Symposium on Information Theory, 2008. ISIT 2008, Conference: 6-11 July 2008, 5 pgs.	
		MERIC ET AL., "Generic Approach for Hierarchical Modulation Performance Analysis: Application to DVB-SH and DVB-S2", Retrieved from: https://arxiv.org/abs/1103.1742, Submitted March 9, 2011, Draft March 10, 2011, 17 pgs.	
		MHEICH ET AL., "Constellation Shaping for Broadcast Channels in Practical Situations", 19th European Signal Processing Conference (EUSIPCO 2011), Barcelona, Spain, August 29- September 2, 2011, pp. 96-100.	
		MILOVANOVIC ET AL., "Simple Optimization Method of One-Dimensional M-PAM Constellations for the AWGN Channels", 4th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Services, October 13-15, 1999, 4 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		MOORE ET AL., "Pairwise optimization of modulation constellations for non-uniform sources Modulation", Can. J. Elect. Computer Eng. Vol. 34, pp 167-177, 2009.	
		MUHAMMAD, NABIL SVEN, "Coding and Modulation for Spectral Efficient Transmission", University of Stuttgart. August 25, 2010 (http://dx.doi.org/10.18419/opus-2676).	
		MUHAMMAD ET AL., "Joint Optimization of Signal Constellation and Bit Labeling for Bit-Interleaved Coded Modulation with Iterative Decoding", IEEE Communications Letters, September 2005, Vol. 9, No. 9, pp. 775-777.	
		NGO ET AL., "A New Iterative Decoder for Turbo Codes on the Nonlinear Channel with Non-uniform 16QAM Modulation", Turbo Coding 2006, April 3-7, 2006, Munich, 7 pgs.	
		NGO ET AL., "Performance of non-uniform 16QAM modulation over linear and nonlinear channels", Electronics Letters, Vol. 42, Issue 9, April 27, 2006, 2 pgs.	
		OTNES ET AL., "Adaptive Data Rate using ARQ and Nonuniform Constellations", Vehicular Technology Conference pp 1211-1215, 2001.	
		RAPHAELI ET AL., "An Improved Pragmatic Turbo Encoding Scheme for High Spectral Efficiency Using Constellation Shaping", IEEE International Conference on Communications, ICC 2003, Conference May 11-15, 2003, Anchorage, AK, USA, 6 pgs.	
		RAPHAELI ET AL., "Constellation Shaping for Pragmatic Turbo-Coded Modulation with High Spectral Efficiency", IEEE Transactions on Communications, March 2004, Vol. 52, No. 3, pp. 341-345.	
		RUOTSALAINEN ET AL., "On the construction of the higher dimensional constellations", ISIT 2000, Lausanne, Switzerland, June 30-July 5, 2002, pg. 490.	
		SAWAYA, "Performance optimization for capacity-approaching channel coding schemes", XP055604929 Ph.D. report, ENST Paris, France 31 March 2002.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		SAWAYA ET AL., "Multilevel coded modulations based on asymmetric constellations", PROCEEDINGS OF THE 2001 IEEE INTERNATIONAL SYMPOSIUM ON INFORMATION THEORY NEW YORK, NY: IEEE, US, 24 June 2001 pages 281-281.	
		SCHRECKENACH ET AL., "Signal Shaping Using Non-Unique Symbol Mappings", Proceedings of the 43rd Annual Allerton Conference on Communication, Control and Computing, September 2005, 10 pgs.	
		SHANNON ET AL., "A Mathematical Theory of Communication", The Bell System Technical Journal, Vol. 27, July, October 1948, 379-423, 623-656.	
		SHEN ET AL., "On the Design of Modern Multilevel Coded Modulation for Unequal Error Protection", IEEE International Conference on Communications, 2008. ICC '08., Conference: 19-23 May 2008, 6 pgs.	
		SOMMER ET AL., "Signal Shaping by Non-Uniform QAM for AWGN Channels and Applications Using Turbo Coding", ITG Conference on Source and Channel Coding, January 2000, pp. 81-86.	
		SOUTO ET AL., "Iterative Detection and Channel Estimation for WCDMA Systems Employing Non-Uniform QAM Constellations", IST Mobile and Wireless Communications Summit, June 2006, 6 pgs.	
		SOUTO ET AL., "Iterative Turbo Multipath Interference Cancellation for WCDMA System with Non-Uniform Modulation", IEEE, 2005, 5 pgs.	
		SOUTO ET AL., "Non-Uniform Constellations for Broadcasting and Multicasting Services in WCDMA Systems", Retrieved from: http://www.eurasip.org/Proceedings/Ext/IST05/papers/424.pdf, June 19-23, 2005, 5 pgs.	
		STIERSTORFER ET AL., "Asymptotically Optimal Mappings for BICM with M-PAM and M¬¬2QAM", Retrieved from: http://www.lit.lnt.de/papers/elet_set_part_cst_2009.pdf, Draft, July 13, 2009, 6 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		STIERSTORFER ET AL., "Optimizing BICM with convolutional codes for transmission over the AWGN channel", Int. Zurich Seminar on Communications (IZS), March 3-5, 2010, 5 pgs.	
		SUN ET AL., "Approaching Capacity by Equiprobable Signaling on the Gaussian Channel", IEEE Transactions on Information Theory, September 1993, Vol. 39, No. 5, pp. 1714-1716.	
		THEODORAKOPOULOS ET AL., "Comparison of Uniform and Non-uniform M-QAM Schemes for Mobile Video Applications", Proceedings of the 2005 Systems Communications, 2005, Conference: August 14-17, 2005, 6 pgs.	
		TRAN ET AL., "Signal Mappings of 8-Ary Constellations for BICM-ID Systems Over a Rayleigh Fading Channel", Canadian Conference on Electrical and Computer Engineering, 2004, pp. 1809-1813.	
		UNGERBOECK ET AL., "Channel Coding with Multilevel/Phase Signals", IEEE Transactions on Information Theory, January 1982, Vol. IT-28, No. 1, pp. 55-67.	
		VALLES ET AL., "Constellation Design for Improved Iterative LDPC Decoding", Aerospace Conference, 2008 IEEE, Conference: March 1-8, 2008, Big Sky, MT, USA, DOI: 10.1109/AERO.2008.4526371, 7 pgs.	
		VITTHALADEVUNI ET AL., "A Recursive Algorithm for the Exact BER Computation of Generalized Hierarchical QAM Constellations", IEEE Transactions on Information Theory, Vol. 49, No. 1, January 14, 2003, pp. 297-307.	
		VON DEETZEN ET AL., "On Code Design for Unequal Error Protection Multilevel Coding", 7th International ITG Conference on Source and Channel Coding (SCC), 2008, Conference: 14-16 Jan. 2008, 4 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

EXAMINER SIGNATURE			
Examiner Signature	/TESFALDET BOCURE/	Date Considered	03/28/2020

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	~~Unassigned~~ 2633
	Examiner Name	~~Unassigned~~ Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T 5
		WANG ET AL, "Shaping Gain for AWGN Channel by Non-Uniform Constellation in LDPC-Coded System", 11th IEEE Singapore International Conference on Communication Systems, 2008, ICCS 2008, Conference: November 19-21, 2008, pp. 1302-1306.	
		WU ET AL., "Non-uniform and Large Distance Constellation Design for Hierarchical Modulation", 2010 IEEE International Conference on Communications (ICC), Conference: May 23-27, 2010, 5 pgs.	
		XIE ET AL., "Bit-Interleaved LDPC-Coded Modulation with Iterative Demapping and Decoding", IEEE 69th Vehicular Technology Conference, Barcelona, Spain, Date of Conference April 26-29, 2009, 5 pgs.	
		XIE ET AL., "On the Channel Capacity and Iterative Demapping of Generalized 4PAM over AWGN Channel", 11th IEEE Singapore International Conference on Communication Systems, 2008. ICCS 2008. December 2008, pp. 860-863.	
		YANG ET AL., "A Novel BICM-ID System Approaching Shannon-Limit at High Spectrum Efficiency", IEICE Trans. Commun., Vol. E94-B, No. 3, March 2011, pp. 793-795.	
		YUAN ET AL., "Robust Hierarchical Broadcasting for AWGN and Flat Rayleigh Fading Channels using Multilevel Codes", Proceedings of IEEE Fifth International Symposium on Communication Theory & Applications, ISCTA'99, Ambleside, UK, 1999, S. 3 pgs.	
		ZESONG ET AL., "Shaping Gain by Non-Uniform QAM Constellation with Binary Turbo Coded Modulation", PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS, IEEE, vol. 2. pp. 1863-1867, 7 September 2003.	
		ZHANG ET AL., "A New Constellation Shaping Method and Its Performance Evaluation in BICM-ID", IEEE 70th Vehicular Technology Conference Fall (VTC 2009-Fall). Conference: Sept. 20-23, 2009, 5 pgs.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	Unassigned
	Examiner Name	Unassigned
	Attorney Docket Number	C44-00999.CON15

		ZHENG ET AL., "Shaping Gain of LDPC Coded-QAM Transmitting Systems with Non-Uniform Constellation", IEEE, pp. 6-9, 2007.	

EXAMINER SIGNATURE

Examiner Signature	/TESFALDET BOCURE/	Date Considered	03/28/2020

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3) [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Google Scholar (geometrically shaped constellations) (low density parity check coding) (quadrature amplitude modulation) SIGN IN

Articles About 1,160 results (0.13 sec) My profile My library

Any time
Since 2020
Since 2019
Since 2016
Custom range...

Sort by relevance
Sort by date

include patents
include citations

Create alert

Non-uniform **constellations** for ATSC 3.0 [PDF] ieee.org Full-Text @ USPTO-STIC
NS Loghin, J Zöllner, B Mouhouche ... - IEEE Transactions ..., 2016 - ieeexplore.ieee.org
... shaping, which tackles the symbol probabilities by using a shaping encoder, and **geometrical** shaping by mod ... This paper focuses only on **geometrically shaped** NUCs ... [20] B. Mouhouche, D. Ansorregui, and A. Mourad, "High order non-uniform **constellations** for broadcasting ...
Cited by 95 Related articles All 2 versions

Bandwidth efficient and rate-matched **low-density parity-check coded modulation** [PDF] ieee.org Full-Text @ USPTO-STIC
G Böcherer, F Steiner, P Schulte - IEEE Transactions on ..., 2015 - ieeexplore.ieee.org
... 39] is used in [41] with LDPC codes on **amplitude** and phase-**shift keying** (APSK) **constellations** ... The generated **check** bits are also **shaped** and embedded in the **shaped** data of the next ... 7] using **Geometric** Huffman **Coding** [44] for shaping and LDPC codes from the DVB-S2 ...
Cited by 310 Related articles All 8 versions

Hybrid probabilistic-**geometric** shaping in optical communication systems [PDF] ieee.org Full-Text @ USPTO-STIC
Z Qu, IB Djordjevic - 2018 IEEE Photonics Conference (IPC), 2018 - ieeexplore.ieee.org
... I. INTRODUCTION **Geometric** shaping (GS) and probabilistic **amplitude** shaping (PAS) have been extensively studied to bridge the gap ... REFERENCES [1] Z. Qu and IB Djordjevic, "**Geometrically shaped** 16QAM outperforming probabilistically **shaped** 16QAM," in ...
Cited by 3 Related articles All 3 versions

Universal hybrid probabilistic-**geometric** shaping based on two-dimensional distribution matchers
Z Qu, S Zhang, IB Djordjevic - 2018 Optical Fiber ..., 2018 - ieeexplore.ieee.org
... 1. Introduction Recently both **geometric** shaping (GS) and probabilistic **amplitude** shaping (PAS) have been extensively studied ... 1. the QAM symbols are pulse-**shaped** by the 92 G Sa/s arbitrary waveform ... [2] Z. Qu and IB Djordjevic, "**Geometrically shaped** 16QAM outperforming ...
Cited by 21 Related articles All 4 versions

Beyond 240 Gb/s per wavelength optical transmission using **coded** hybrid subcarrier/**amplitude**/phase/polarization **modulation** [PDF] ieee.org Full-Text @ USPTO-STIC
HQ Bulushev, IB Djordjevic - IEEE Photonics Technology Letters, 2010 - ieeexplore.ieee.org
... Using symmet- rical **geometric** shapes results in closed form numbers for the ... Fig. 2. Signal **constellations** for: (a) 8-HAPP and (b) 20-H-SAPP ... 2. Fig. 2 shows the case for which (a) 8-HAPP, and (b) 20-H- SAPP, where different point color/**shape** represents a different subcarrier ...
Cited by 20 Related articles All 8 versions

Four-dimensional **coded modulation** with **bit**-wise decoders for future optical communications [PDF] osapublishing.org Full View
A Alvarado, E Agrell - Journal of Lightwave Technology, 2015 - osapublishing.org
... Somewhat less power efficient, but easier to implement, are the cubically **shaped constellations** based on ... These **constellations**, however, are not the best choice for **coded** systems based on a ... the transmitted vector Xn is selected with equal probability from a **constellation** S {s1 ,
Cited by 59 Related articles All 17 versions

LDPC **coded modulation** with probabilistic shaping for optical fiber systems [PDF] arxiv.org Full-Text @ USPTO-STIC
T Fehenberger, G Böcherer, ... - 2015 Optical Fiber ..., 2015 - ieeexplore.ieee.org
... The main implementation advantage of probabilistic shaping over **geometric** shaping is that it does not ... Optimized Input Distribution We probabilistically **shape** 16-QAM and 64-QAM by assigning larger ... in optical fiber simulations that 16-QAM and 64-QAM with **shaped** input allow ...
Cited by 70 Related articles All 15 versions

Multi-dimensional space Gray **code** maps for multi-dimensional phase **modulation** as applied to LDPC (**Low Density Parity Check**) **coded modulation** [PDF] googleapis.com

BZ Shen, TK Lee, KB Cameron, HT Tran - US Patent 7,436,902, 2008 - Google Patents

... signal have the same **code** rate and the same **modulation** (the same **constellation shape** whose **constellation** .. 1-dimensional) 3 **bit** basic Gray **code** maps (shown using 8 PSK **shaped constellations**) according to ... That is to say, the **constellation** and/or mapping of the symbols of ...

Cited by 57 Related articles All 4 versions

On probabilistic shaping of **quadrature amplitude modulation** for the nonlinear fiber channel [PDF] ieee.org Full View

T Fehenberger, A Alvarado... - Journal of Lightwave ... , 2016 - ieeexplore.ieee.org

... Probabilistic shaping also gives larger shaping gains than purely **geometric** shaping [25, Fig ... property of (13) and show the 1D PMF P1D for positive PAM **constellations** points x ... This limited number of **shaped** input PMFs potentially allows an easier implementa- tion of shaping ...

Cited by 154 Related articles All 8 versions

Bit-interleaved polar-**coded** OFDM for **low**-latency M2M wireless communications [PDF] ieee.org Full-Text @ USPTO-STIC

T Koike-Akino, Y Wang, SC Draper... - 2017 IEEE ... , 2017 - ieeexplore.ieee.org

... 4. Super-Gaussian distribution for an order of $\gamma = \{0.1, 0.5, 1, 2, 10\}$. **Shaped constellation** is determined by equally-spaced CDF points. B. Super-Gaussian QAM for **Geometric Constellation** Shaping Most typically, regular QAM signal **constellations** are used for communications ...

Cited by 17 Related articles All 5 versions

Google 1 2 3 4 5 6 7 8 9 10 Next

Google Scholar (orthogonality) (hierarchical constellation) (pulse amplitude modulation) SIGN IN

Articles About 4,950 results (0.10 sec) My profile My library

Any time
Since 2020
Since 2019
Since 2016
Custom range...

Sort by relevance
Sort by date

include patents
include citations

Create alert

Performance analysis of DVB-H OFDM **hierarchical modulation** in **impulse** noise environment [PDF] ieee.org
V Tamgnoue, V Moeyaert, S Bette... - 2007 14th IEEE ..., 2007 - ieeexplore.ieee.org
... Keying), equivalent to 4-QAM, and thus specifies the (**Orthogonal** Frequency Division Multiplexing). And for that ... o u#1l. non-uniform **constellation**. **amplitude** ... Alouini. "BER Computation of 4/M-QAM **Hierarchical Constellations**", IEEE Transactions on Broadcasting, Vol ...
Cited by 7 Related articles All 6 versions

Single-carrier frequency domain equalisation with **hierarchical constellations**: an efficient transmission technique for broadcast and multicast systems [PDF] ieee.org Full-Text @ USPTO-STIC
J Silva, R Dinis, N Souto, P Montezuma - IET communications, 2012 - IET
... as such require specialised algorithms at higher layers [4–6]. Since **orthogonal** frequency division ... it is desirable to employ IB-DFE schemes with SC-FDE with **hierarchical constellations** ... **constellations**: For a uniform M-**pulse amplitude modulation** (M-PAM) **constellation** we have ...
Cited by 23 Related articles All 4 versions

Nyquist four-level **pulse amplitude modulation** scheme (PAM-4) based on **hierarchical modulation** in IM/DD-TDM PON with hybrid equalization Full-Text @ USPTO-STIC
N Feng, X Sun - Optics Communications, 2020 - Elsevier
... also been envisioned as a key technology in the power domain non-**orthogonal** multiple access ... between the **constellation** points in the actual PAM-4 (ie, level-2 of **hierarchy**) **constellation** ... Note that the traditional non-**hierarchical** Nyquist **pulsed** PAM-4 (NPAM-4) is considered ...
Cited by 2 Related articles All 2 versions

Orthogonal frequency division multiplexing: A multi-carrier **modulation** scheme [PDF] ieee.org Full-Text @ USPTO-STIC
Y Wu, WY Zou - IEEE Transactions on Consumer Electronics, 1995 - ieeexplore.ieee.org
... longer than the multipath delay, IS1 can be eliminated and **orthogonality** can be ... Wu and Zou: **Orthogonal** Frequency Division Multiplexing: A Multi-Carrier **Modulation** Scheme ... 4 4 **Hierarchical Modulation** to Provide Layered Services Similar to spectrum shaping, since the input ...
Cited by 544 Related articles All 7 versions

Enhanced **hierarchical modulation** with interference cancellation for OFDM systems [PDF] ieee.org Full-Text @ USPTO-STIC
RT Juang, P Ting, KY Lin, HP Lin ... - 2009 IEEE 20th ..., 2009 - ieeexplore.ieee.org
... power ratio of transmitted signal can be reduced, especially for OFDM (**orthogonal** frequency division ... 1 shows the signal **constellation** of a regular square- shaped 16QAM/QPSK **hierarchical modulation** ... while it is a regular square- shaped 16QAM/QPSK **constellation** if 25.0 = λ ...
Cited by 6 Related articles All 2 versions

Iterative Decoding and Channel Estimation of MIMO-OFDM Transmissions with **Hierarchical Constellations** and Implicit Pilots [PDF] ieee.org Full-Text @ USPTO-STIC
N Souto, R Dinis, JC Silva - 2007 IEEE International ..., 2007 - ieeexplore.ieee.org
... Figure 2 we show a transmitter chain that incorporates QAM **hierarchical constellations** into an ... Performance of the LPB of a 16-QAM (kl=0.4) **hierarchical constellation** for different ... Cimini. "Analysis and Simulation of a Digital Mobile Channel using **Orthogonal** Frequency Division ...
Cited by 2 Related articles All 2 versions

Adaptive **hierarchical modulation** over correlated MIMO fading channels [PDF] ieee.org Full-Text @ USPTO-STIC
G Li, Z Zhang, Z Lu - 2008 9th International Conference on ..., 2008 - ieeexplore.ieee.org
... On the other hand, to achieve reliable transmission in a fading channel, **orthogonal** space-time ...

The proposed scheme uses a generalized **hierarchical** M-QAM **constellation**) 2 (zy M + = with Gray ... distances between points in the first, second, th, z " levels of **hierarchy** on the Q ...
Cited by 5 Related articles All 2 versions

Efficient channel estimation for ofdm systems with **hierarchical constellations** [PDF] ieee.org Full-Text @ USPTO-STIC

N Souto, R Dinis, JC Silva - 2007 International Symposium on ..., 2007 - ieeexplore.ieee.org

... I. INTRODUCTION It is widely recognized that OFDM **modulations** (**Orthogonal** Frequency Division Multiplexing ... 2 we show a transmitter chain that incorporates QAM **hierarchical constellations** into an ... stream is encoded, interleaved and mapped into the **constellation** symbols in ...
Cited by 1 Related articles All 9 versions

Optimal alphabets and binary labelings for BICM at low SNR [PDF] ieee.org Full-Text @ USPTO-STIC

E Agrell, A Alvarado - IEEE transactions on information theory, 2011 - ieeexplore.ieee.org

... A general answer to the question of FOO **constellations** for BICM is also given ... if and only if they are labeled by the NBC; if the **constellation** is based on ... capacity, folded binary code, Gray code, Hadamard transform, natural binary code, **pulse amplitude modulation** (PAM), phase ...
Cited by 43 Related articles All 13 versions

[BOOK] Quadrature **amplitude modulation**: From basics to adaptive trellis-coded, turbo-equalised and space-time coded OFDM, CDMA and MC-CDMA systems [PDF] soton.ac.uk

L Hanzo, SX Ng, WT Webb, T Keller - 2004 - eprints.soton.ac.uk

... 267 8.5.2 Signal **Constellations** and Bitmapping ... 320 10.8.2 DistortionoftheStar-**Constellation** ... 749 23.5.4 M-ary **Orthogonal** Signalling ...
Cited by 425 Related articles All 8 versions

Related searches

pulse amplitude modulation **scheme**
hierarchical constellations **channel estimation**
hierarchical constellations **efficient transmission technique**
coded ofdm quadrature amplitude modulation
ofdm communications hierarchical modulations
ofdma based networks hierarchical modulation
adaptive hierarchical modulation **mimo fading channels**
vector rotation hierarchical modulation
multi-hop transmission hierarchical modulation
ofdma pon hierarchical modulation
long term evolution hierarchical modulations
performance analysis hierarchically modulated **signals**
modulation **scheme** orthogonal **frequency division multiplexing**
hierarchical constellations **implicit pilots**
hierarchical constellations **frequency domain equalisation**
hierarchical constellations **mimo ofdm transmissions**

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

BIB DATA SHEET

CONFIRMATION NO. 6350

SERIAL NUMBER	FILING or 371(c) DATE RULE	CLASS	GROUP ART UNIT	ATTORNEY DOCKET NO.
16/728,384	12/27/2019	375	2633	C44-00999.CON15

APPLICANTS

Constellation Designs, LLC, Anaheim, CA;

INVENTORS

Maged F. Barsoum, San Jose, CA;
Christopher R. Jones, Pacific Palisades, CA;

**** CONTINUING DATA ***************************

This application is a CON of 16/206,991 11/30/2018 PAT 10567980
which is a CON of 15/682,475 08/21/2017 PAT 10149179
which is a CON of 15/200,800 07/01/2016 PAT 9743292
which is a CON of 14/491,731 09/19/2014 PAT 9385832
which is a CON of 13/618,630 09/14/2012 PAT 8842761
which is a CON of 13/118,921 05/31/2011 PAT 8270511
which is a CON of 12/156,989 06/05/2008 PAT 7978777
which claims benefit of 60/933,319 06/05/2007

**** FOREIGN APPLICATIONS ***************************

**** IF REQUIRED, FOREIGN FILING LICENSE GRANTED ****

01/28/2020

Foreign Priority claimed ❑ Yes ☑ No 35 USC 119(a-d) conditions met ❑ Yes ☑ No Verified and Acknowledged /TESFALDET BOCURE/ Examiner's Signature	❑ Met after Allowance Initials	STATE OR COUNTRY	SHEETS DRAWINGS	TOTAL CLAIMS	INDEPENDENT CLAIMS
		CA	43	30	3

ADDRESS

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806
UNITED STATES

TITLE

Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

FILING FEE RECEIVED		
2720	FEES: Authority has been given in Paper No._____ to charge/credit DEPOSIT ACCOUNT No._____ for following:	❑ All Fees ❑ 1.16 Fees (Filing) ❑ 1.17 Fees (Processing Ext. of time) ❑ 1.18 Fees (Issue) ❑ Other _____ ❑ Credit

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

APPLICATION NUMBER	FILING OR 371(C) DATE	FIRST NAMED APPLICANT	ATTY. DOCKET NO./TITLE
16/728,384	12/27/2019	Maged F. Barsoum	C44-00999.CON15

CONFIRMATION NO. 6350

71897
KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

PUBLICATION NOTICE

OC000000116833471

Title: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

Publication No.US-2020-0145846-A1
Publication Date:05/07/2020

NOTICE OF PUBLICATION OF APPLICATION

The above-identified application will be electronically published as a patent application publication pursuant to 37 CFR 1.211, et seq. The patent application publication number and publication date are set forth above.

The publication may be accessed through the USPTO's publically available Searchable Databases via the Internet at www.uspto.gov. The direct link to access the publication is currently http://www.uspto.gov/patft/.

The publication process established by the Office does not provide for mailing a copy of the publication to applicant. A copy of the publication may be obtained from the Office upon payment of the appropriate fee set forth in 37 CFR 1.19(a)(1). Orders for copies of patent application publications are handled by the USPTO's Public Records Division. The Public Records Division can be reached by telephone at (571) 272-3150 or (800) 972-6382, by facsimile at (571) 273-3250, by mail addressed to the United States Patent and Trademark Office, Public Records Division, Alexandria, VA 22313-1450 or via the Internet.

In addition, information on the status of the application, including the mailing date of Office actions and the dates of receipt of correspondence filed in the Office, may also be accessed via the Internet through the Patent Electronic Business Center at www.uspto.gov using the public side of the Patent Application Information and Retrieval (PAIR) system. The direct link to access this status information is currently https://portal.uspto.gov/pair/PublicPair. Prior to publication, such status information is confidential and may only be obtained by applicant using the private side of PAIR.

Further assistance in electronically accessing the publication, or about PAIR, is available by calling the Patent Electronic Business Center at 1-866-217-9197.

Office of Data Managment, Application Assistance Unit (571) 272-4000, or (571) 272-4200, or 1-888-786-0101

PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF EFS WEB TRANSMISSION

I hereby certify that this correspondence is being electronically filed with the United States Patent and Trademark Office on July 9, 2020 at or before 11:59 p.m. Pacific Time under the Rules of 37 CFR § 1.8.

/Katrina Lee/
Katrina Lee

Appl No.	: 16/728,384	Confirmation No. 6350
Applicant	: Constellation Designs, LLC	
Filed	: December 27, 2019	
Title	: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	
TC/A.U.	: 2633	
Examiner	: Tesfaldet Bocure	
Docket No.	: C44-00999.CON15	
Customer No.	: 71897	

AMENDMENT

Mail Stop Amendment
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806
July 9, 2020

Commissioner:

In response to the Office action mailed April 9, 2020, please amend the above-identified application as follows:

Amendments to the Specification are on page 2 of this paper.

Amendments to the Claims are reflected in the listing of claims which begins on page 3 of this paper.

Remarks/Arguments begin on page 18 of this paper.

Amendments to the Specification:

Please amend paragraph [0001] as follows:

[0001] This application is a continuation of application number 16/206,991 filed November 30, 2018 and issued on February 18, 2020 as U.S. Patent No. 10,567,980, which application is a continuation of application number 15/682,475 filed August 21, 2017 and issued on December 4, 2018 as U.S. Patent No. 10,149,179, which application is a continuation of application number 15/200,800 filed July 1, 2016 and issued on August 22, 2017 as U.S. Patent No. 9,743,292, which application is a continuation of application number 14/491,731 filed September 19, 2014 and issued on July 5, 2016 as U.S. Patent No. 9,385,832, which application is a continuation of application number 13/618,630 filed September 14, 2012 and issued on September 23, 2014 as U.S. Patent No. 8,842,761, which application is a continuation of application number 13/118,921 filed May 31, 2011 and issued on September 18, 2012 as U.S. Patent No. 8,270,511, which application is a continuation of application number 12/156,989 filed June 5, 2008 and issued on July 12, 2011 as U.S. Patent No. 7,978,777, which application claimed priority to U.S. Provisional Application 60/933,319 filed June 5, 2007, the disclosures of which are incorporated herein by reference.

Amendments to the Claims:

This listing of claims will replace all prior versions, and listings, of claims in the application:

Listing of Claims:

1. (Original) A communication system, comprising:

a receiver capable of receiving signals via a communication channel having a channel signal-to-noise ratio (SNR), wherein the receiver comprises:

a demodulator capable of demodulating a received signal into a demodulated signal;

a demapper, coupled to the demodulator, capable of determining likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels; and

a decoder, coupled to the demapper, capable of using the likelihoods determined by the demapper to provide a sequence of received bits based upon a low density parity check (LDPC) code.

2. (Original) The communication system of claim 1, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations;

the plurality of non-uniform symbol constellations includes a plurality of non-uniform constellations of a first type that comprise multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations, where non-uniform symbol constellations of the first type include at least two constellation points having identical locations and different labels;

the receiver is capable of selecting an LDPC code rate and the non-uniform symbol constellation as a pair from a plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs; and

each of the plurality of non-uniform symbol constellations is only included in one of the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs.

3. (Original) The communication system of claim 2, wherein the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs includes:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 5/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the first type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the first type.

4. (Original) The communication system of claim 3, wherein the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

equal to or less than 5.27dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate that is below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point non-uniform symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellations of the first type.

5. (Currently Amended) The communication system of claim 4, wherein:

the plurality of non-uniform symbol constellations include a plurality of non-uniform constellations of a second type that comprises multiple different sixteen-point symbol constellations, multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations;

no two constellation points within a non-uniform symbol constellation of the second type share an identical location;

the demapper is also capable of determining likelihoods using the demodulated signal and a non-uniform symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs include:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type;

the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

between 3.11dB and 9.25dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

between 9.00dB and 15.93dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

between 18.72dB and 22.13dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

between 24.79dB and 28.20dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type.

6. (Original) The communication system of claim 1, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of non-uniform symbol constellations is capable of providing greater parallel decoding capacity when operated at a symbol constellation operating SNR than the other plurality of non-uniform symbol constellations when operated at the same SNR.

7. (Original) The communication system of claim 6, wherein the symbol constellation operating SNR is a channel SNR where the receiver is capable of using the respective non-uniform symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

8. (Original) The communication system of claim 1, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of constellation points so as to maximize parallel decoding capacity at a symbol constellation operating SNR subject to at least one constraint.

9. (Original) The communication system of claim 1, wherein the non-uniform symbol constellation is an orthogonalized non-uniform Pulse Amplitude Modulated (PAM) constellation.

10. (Original) The communication system of claim 1, further comprising a transmitter capable of transmitting signals via the communication channel, where the transmitter comprises:

a coder capable of receiving bits and outputting encoded bits using a Low Density Parity Check (LDPC) code;

a mapper, coupled to the coder, capable of mapping the encoded bits to symbols in the non-uniform symbol constellation; and

a modulator, coupled to the mapper, capable of generating a signal for transmission via the communication channel based upon symbols selected by the mapper.

11. (Original) A communication system, comprising:

a receiver that receives signals via a communication channel having a channel signal-to-noise ratio (SNR), wherein the receiver comprises:

a demodulator that demodulates a received signal into a demodulated signal;

a demapper that determines likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels; and

a decoder that uses the likelihoods determined by the demapper to provide a sequence of received bits based upon a low density parity check (LDPC) code;

wherein the demapper is interposed between the demodulator and the decoder and the demapper receives information from the demodulator and provides information to the decoder.

12. (Original) The communication system of claim 11, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations;

the plurality of non-uniform symbol constellations includes a plurality of non-uniform constellations of a first type that comprise multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations, where non-uniform symbol constellations of the first type include at least two constellation points having identical locations and different labels;

the receiver selects an LDPC code rate and the non-uniform symbol constellation as a pair from a plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs; and

each of the plurality of non-uniform symbol constellations is only included in one of the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs.

13. (Original) The communication system of claim 12, wherein the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs includes:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 5/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the first type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the first type.

14. (Original) The communication system of claim 13, wherein the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

equal to or less than 5.27dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate that is below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point non-uniform symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellations of the first type.

15. (Currently Amended) The communication system of claim 14, wherein:

the plurality of non-uniform symbol constellations includes a plurality of non-uniform constellations of a second type that comprises multiple different sixteen-point symbol constellations, multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations;

no two constellation points within a non-uniform symbol constellation of the second type share an identical location;

the demapper also determines likelihoods using the demodulated signal and a non-uniform symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs includes:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type;

the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

between 3.11dB and 9.25dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

between 9.00dB and 15.93dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

between 18.72dB and 22.13dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

between 24.79dB and 28.20dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type.

16. (Original) The communication system of claim 11, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of non-uniform symbol constellations provides greater parallel decoding capacity when operated at a symbol constellation operating SNR than the other plurality of non-uniform symbol constellations when operated at the same SNR.

17. (Original) The communication system of claim 16, wherein the symbol constellation operating SNR is a channel SNR where the receiver uses the respective non-uniform symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

18. (Original) The communication system of claim 11, wherein

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of constellation points so as to maximize parallel decoding capacity at a symbol constellation operating SNR subject to at least one constraint.

19. (Original) The communication system of claim 11, wherein the non-uniform symbol constellation is an orthogonalized non-uniform Pulse Amplitude Modulated (PAM) constellation.

20. (Original) The communication system of claim 11, further comprising a transmitter that transmits signals via the communication channel, where the transmitter comprises:

a coder that receives bits and outputs encoded bits using a Low Density Parity Check (LDPC) code;

a mapper that maps the encoded bits to symbols in the selected symbol constellation; and

a modulator that generates a signal for transmission via the communication channel based upon symbols selected by the mapper;

wherein the mapper is interposed between the modulator and the coder and the mapper receives information from the coder and provides information to the modulator.

21. (Original) The communication system of claim 11, further comprising a transmitter that transmits signals via the communication channel, where the transmitter uses the non-uniform symbol constellation to transform encoded bits into the transmitted signals.

22. (Original) A communication system, comprising a receiver that receives signals via a communication channel having a channel signal-to-noise ratio (SNR), wherein the receiver uses a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels to transform the received signals into received bits.

23. (Original) The communication system of claim 22, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations;

the plurality of non-uniform symbol constellations includes a plurality of non-uniform constellations of a first type that comprise multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations, where non-uniform symbol constellations of the first type include at least two constellation points having identical locations and different labels;

the receiver selects an LDPC code rate and the non-uniform symbol constellation as a pair from a plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs; and

each of the plurality of non-uniform symbol constellations is only included in one of the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs.

24. (Original) The communication system of claim 23, wherein the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs includes:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 5/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the first type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the first type.

25. (Original) The communication system of claim 24, wherein the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

equal to or less than 5.27dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate that is below 1/3 and a sixty-four-point non-uniform symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point non-uniform symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and non-uniform symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point non-uniform symbol constellations of the first type.

26. (Currently Amended) The communication system of claim 25, wherein:

the plurality of non-uniform symbol constellations includes a plurality of non-uniform constellations of a second type that comprises multiple different sixteen-point symbol constellations, multiple different sixty-four-point symbol constellations, multiple different two-hundred-fifty-six-point symbol constellations, and multiple different one-thousand-twenty-four-point symbol constellations;

no two constellation points within a non-uniform symbol constellation of the second type share an identical location;

the receiver also transforms received signals into received bits using a non-uniform symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and non-uniform symbol constellation pairs include:

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type;

the receiver is capable of decoding the signals received over the communication channel when the SNR of the communication channel is:

between 3.11dB and 9.25dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/8 and less than or equal to 6/8 and a sixteen-point non-uniform symbol constellation of the second type;

between 9.00dB and 15.93dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 3/6 and less than or equal to 5/6 and a sixty-four-point non-uniform symbol constellation of the second type;

between 18.72dB and 22.13dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 6/8 and less than or equal to 7/8 and a two-hundred-and-fifty-six-point non-uniform symbol constellation of the second type; and

between 24.79dB and 28.20dB while using at least one of the multiple LDPC code rate and non-uniform symbol constellation pairs that each include an LDPC code rate that is equal to or greater than 8/10 and less than or equal to 9/10 and a one-thousand-twenty-four-point non-uniform symbol constellation of the second type.

27. (Original) The communication system of claim 22, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of non-uniform symbol constellations provides greater parallel decoding capacity when operated at a symbol constellation operating SNR than the other plurality of non-uniform symbol constellations when operated at the same SNR.

28. (Original) The communication system of claim 27, wherein the symbol constellation operating SNR is a channel SNR where the receiver uses the respective non-uniform symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

29. (Original) The communication system of claim 22, wherein:

the non-uniform symbol constellation is selected from a plurality of non-uniform symbol constellations; and

each of the plurality of non-uniform symbol constellations is characterized by the assignment of labels and spacing of constellation points so as to maximize parallel decoding capacity at a symbol constellation operating SNR subject to at least one constraint.

30. (Original) The communication system of claim 22, wherein at least one of the plurality of non-uniform symbol constellations is an orthogonalized non-uniform Pulse Amplitude Modulated (PAM) constellation.

REMARKS/ARGUMENTS

SUMMARY OF THE OFFICE ACTION

The Office action dated April 9, 2020 (the "Office action") contains the following objections and rejections:

- the disclosure is objected to because the issuance of U.S. Patent Application No. 16/206,991 occurred on Feb. 18, 2020, and an update with the patent number is required;
- claims 22-30 are rejected under 35 U.S.C. §112 (pre-AIA), second paragraph, as being indefinite;
- claims 1-30 are rejected under 35 U.S.C. §112 (pre-AIA), first paragraph, as failing to comply with the written description requirement;
- claims 1, 10, 11, 20, 21 and 22 are rejected under pre-AIA 35 U.S.C. §102(e) as being anticipated by U.S. Patent Publication No. 2008/0200114 to Eberlein et al. (Eberlein); and
- claims 9, 19, and 30 are rejected under pre-AIA 35 U.S.C. §103(a) in view of Eberlein and U.S. Patent Publication No. 2005/0111581 to Walker et al. (Walker)

SUMMARY OF THE AMENDMENTS TO THE SPECIFICATION

Applicant has amended the specification as requested on page 2 of the Office action to update the priority information with the newly granted U.S. Patent No. 10,567,980. Applicant respectfully requests that the objection be withdrawn in light of the amendment.

SUMMARY OF THE AMENDMENTS TO THE CLAIMS

Claims 1 – 30 are currently pending. Claims 5, 15, and 26 are amended for clarity.

REJECTION OF CLAIMS 22 – 30 UNDER 35 U.S.C. §112

Claims 22-30 are rejected under 35 U.S.C. §112 (pre-AIA), first paragraph, on the basis of claim 22 reciting a single means. Applicant respectfully disagrees and submits

that the rejection is improper. M.P.E.P. § 2164.08(a) sets out the circumstances in which such a rejection is appropriate citing to "*In re Hyatt*, 708 F.2d 712, 714-715, 218 USPQ 195, 197 (Fed. Cir. 1983) (A single means claim which covered every conceivable means for achieving the stated purpose was held nonenabling for the scope of the claim because the specification disclosed at most only those means known to the inventor.)" M.P.E.P. § 2164.08(a). It is important to appreciate, that single means rejection is only appropriate with respect to means-plus-function claims. Furthermore, as the CAFC noted in *In re Hyatt*, "[T]he first sentence of the second paragraph of Sec. 112 is essentially a requirement for precision and definiteness of claim language. If the scope of subject matter embraced by a claim is clear, and if the applicant has not otherwise indicated that he intends the claim to be of a different scope, then the claim does particularly point out and distinctly claim the subject matter which the applicant regards as his invention." *In re Hyatt*, 708 F.2d 712, 714 (Fed. Cir. 1983). The Office action correctly acknowledges that claim 22 is not drafted as a "means-plus-function" claim. *See* MPEP §2181 ("...a claim limitation that does not use the term "means" or "step will trigger a rebuttable presumption that ... pre-AIA 35 U.S.C. 112, sixth paragraph, does not apply."). Applicant agrees that a "receiver" is a structure used in communication systems that is well-known to a POSITA and the scope of the recited "receiver" is clear. Accordingly, claim 22 cannot appropriately be the subject of a "single means claim" rejection and Applicant respectfully requests withdrawal of the rejection of claim 22. For similar reasons, Applicant also requests withdrawal of the single means rejections of claims 23 – 30.

REJECTION OF CLAIMS 1 – 30 UNDER 35 U.S.C. §112

Claims 1-30 are rejected under 35 U.S.C. §112 (pre-AIA), second paragraph on the basis of a lack of written description support. With specific regard to claims 1 and 11, the Office action asserts a lack of written description support for "a demapper, coupled to the demodulator, capable of determining likelihoods using the demodulated signal and a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels". Furthermore, claim 22 is rejected on the basis of a lack of written description support for "[A] communication system, comprising a receiver

that receives signals via a communication channel having a channel signal-to-noise ratio (SNR), wherein the receiver uses a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels to transform the received signals into received bits".

Applicant respectfully disagrees and provides the following explanation of the written description support for the various limitations of the claims.

Claim 1: "a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels"

Support for the use of a non-uniform symbol constellation that includes at least two constellation points having identical locations and different labels can be found throughout the specification and figures of the above referenced application. For example, PAM-4 constellations optimized for parallel decode (PD) capacity are observed in ¶ [0072] to have four points with different labels, but only have two unique point locations at low SNR. The PAM-4 constellations in question are illustrated in Fig. 10d, which is reproduced below including annotations.

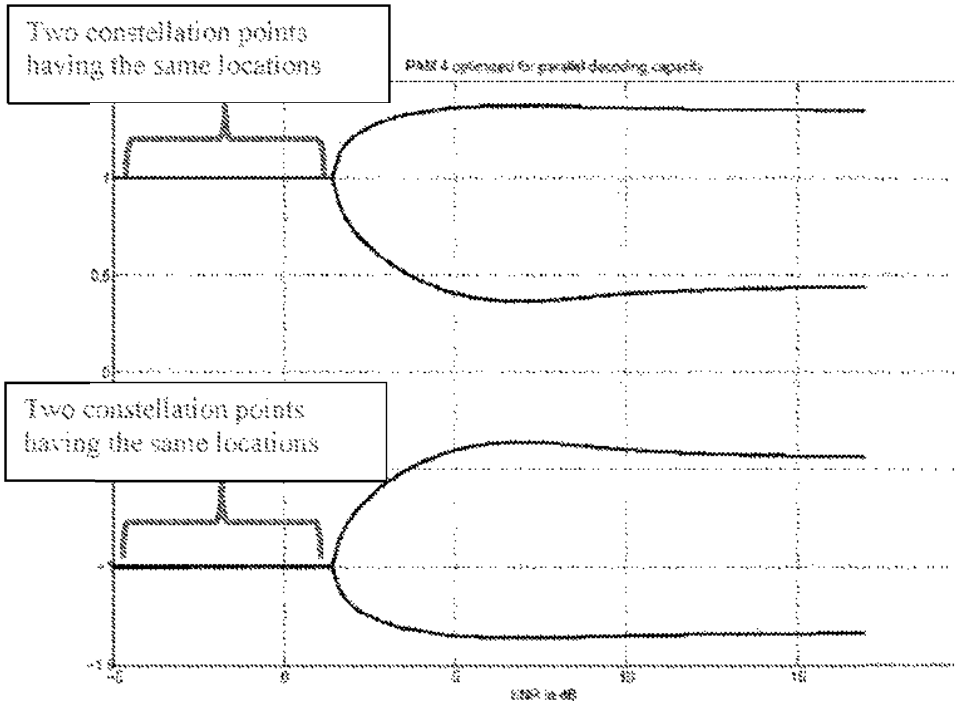


FIG. 10d

As can readily be appreciated, the locations of the four constellation points of a PD capacity optimized PAM-4 constellation are dependent upon SNR. At low SNR, two constellation points with different labels (i.e., x_0 & x_1) occupy the same first location (i.e. amplitude -1) and two constellation points with different labels (i.e., x_2 & x_3) occupy the same second location (i.e. amplitude 1). *See* Fig. 11b. Similar constellations can be observed in Fig. 12d, which is reproduced below with annotations.

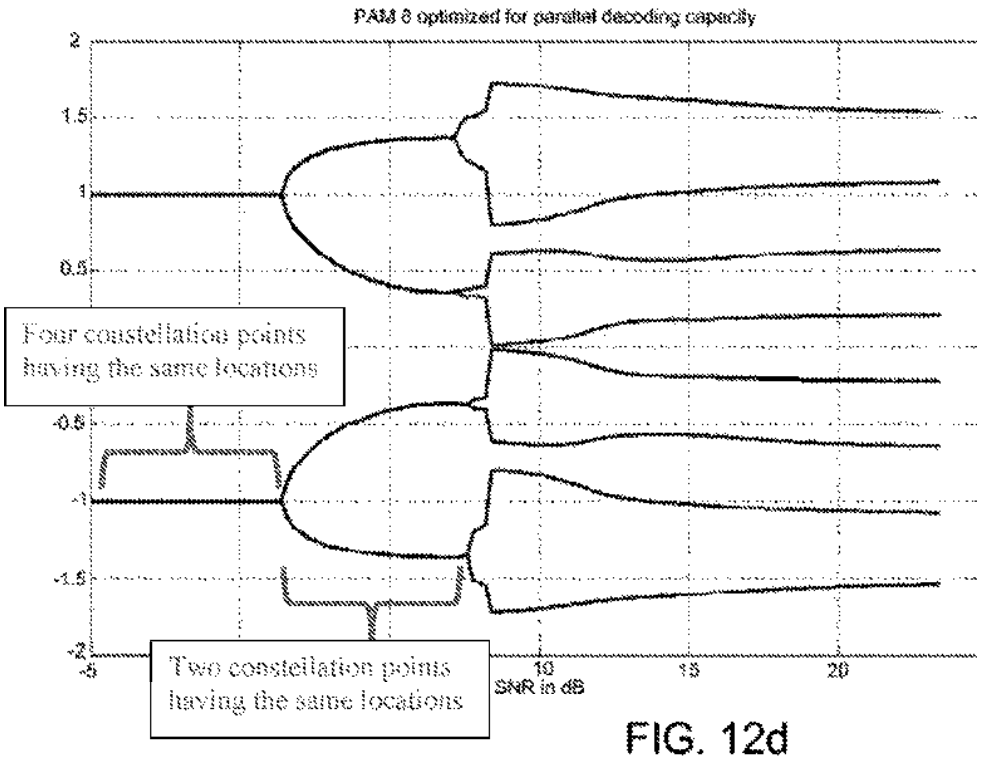


FIG. 12d

As SNR decreases, the locations of the eight constellation points of the PD capacity optimized PAM-8 constellations merge to occupy four locations (i.e. two constellation points occupying each of the four locations) and then to occupy two locations (i.e. four constellation points occupying each of the two locations). Specific amplitudes of PD optimized PAM-8 constellations for specific SNRs are provided in Fig. 13b, which is reproduced below including annotations indicating particular constellation points having identical locations and different labels.

PAM-8 constellations optimized for parallel decoding capacity at different

	(bps)	0.5	1.0	1.5	2.0	2.5
	(SNR)	0.19	5.27	9.00	12.42	15.93
Labels	x_0	-1.00	-1.36	-1.72	-1.64	-1.60
	x_1	-1.00	-1.36	-0.81	-0.97	-1.03
	x_2	-1.00	-0.39	1.72	1.64	-0.19
	x_3	-1.00	-0.39	-0.62	-0.58	-0.58
	x_4	1.00	1.36	0.62	0.58	1.60
	x_5	1.00	1.36	0.02	0.15	1.03
	x_6	1.00	0.39	0.81	0.97	0.19
	x_7	1.00	0.39	-0.02	-0.15	0.58

FIG. 13b

Similar non-uniform symbol constellations in which at least two constellation points have identical locations and different labels can be observed in Figs. 14d, 15b, 16d, 17b, 22b and 23b. Non-uniform symbol constellations that use phase shift keying in which at least two constellation points have identical locations and different labels can be observed in Figs. 19, and 20. Furthermore, the specification of the above referenced application discloses that one-dimensional non-uniform PAM and PSK constellations can be utilized to construct multi-dimensional constellations (e.g. QAMs and APSKs) resulting in multidimensional constellations that include at least two constellation points having identical locations and different labels. (*See* ¶¶ [0008] "... a method for locating points (in one or multiple dimensional space)...", [0055] "The process 50 commences with the selection (52) of an appropriate constellation size *M* and a desired capacity per dimension *η*.", [0080] "Quadrature amplitude modulation (QAM) constellations can be constructed by orthogonalizing PAM constellations into QAM inphase and quadrature components.").

In addition, non-uniform multidimensional constellations are also disclosed that can be obtained through direct optimization in multiple dimensions. (*See* ¶ [0082] "Rather than designing constellations in 1-D (PAM for instance) and then extending to 2-D (QAM), it is possible to take direct advantage in the optimization step of the additional degree of freedom presented by an extra spatial dimension. In general it is possible to design *N*-dimensional constellations and associated labelings.").

Written Description Support for Claim 1

In view of the above, Applicant respectfully submits that claim 1 satisfies all of the requirements of 35 U.S.C. §112 and requests withdrawal of the rejection of claim 1 under 35 U.S.C. §112 (pre-AIA), first paragraph.

Written Description Support for Claims 2-30

Applicant respectfully submits that support for claims 2 – 30 can be found throughout the specification of the above referenced application including the paragraphs and figures referenced above with respect to the discussion of claim 1. Accordingly, Applicant requests withdrawal of the rejection of claims 2 – 30 under 35 U.S.C. §112 (pre-AIA), first paragraph.

REJECTION OF CLAIMS 1, 10, 11, 20, 21 AND 22 UNDER 35 U.S.C. §102(e)

Claims 1, 10, 11, 20, 21, and 22 are rejected under pre-AIA 35 U.S.C. §102(e) as anticipated by Eberlein. Applicant respectfully disagrees. Eberlein discloses the use of a hierarchical modulation scheme involving the use of non-uniform QAM constellations similar to the constellation shown in Fig. 3A of Eberlein, which is reproduced below.

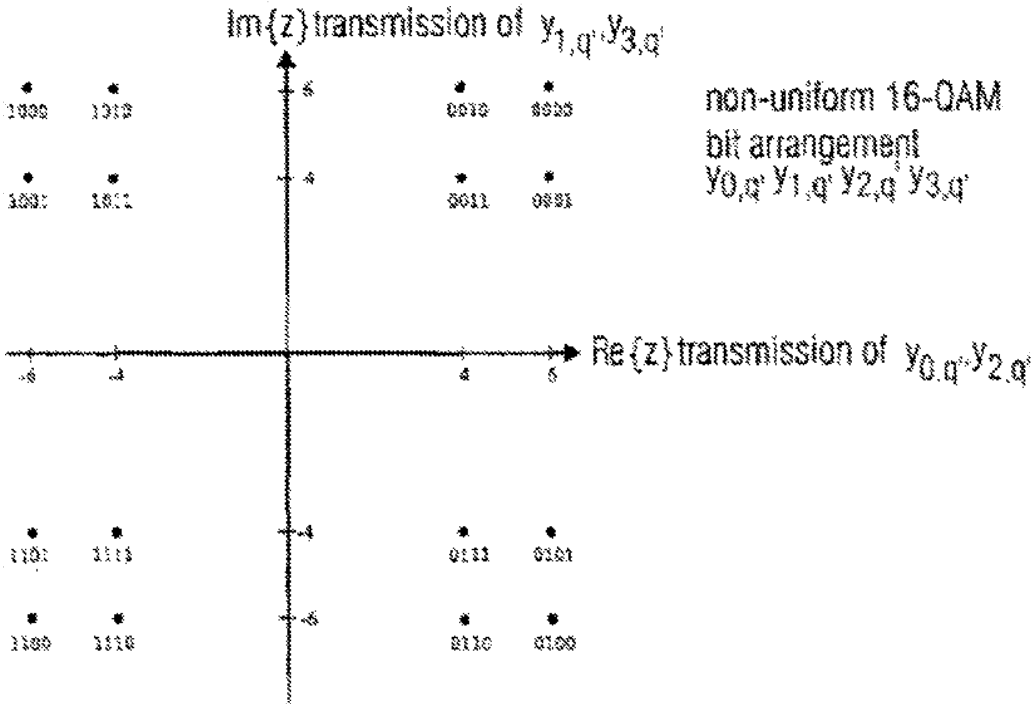


FIGURE 3A

The statement of the rejection asserts that Fig. 3A of Eberlein illustrates a non-uniform constellation that includes at least two constellation points having identical locations and different labels. *See* Office action, 6. The Office action asserts that "signal points located in the first quadrant shown in figure 3A, claimed identical location, having different labelling 0010, 0000, 001, and 0011". *Id* at 6. Applicant respectfully submits that Fig. 3A of Eberlein above clearly shows the four points in the first quadrant as having different locations. The location of each point is expressed as the combination of a real (i.e. Re{z}) and imaginary component (i.e. Im{z}). Eberlein notes in ¶ [0019] "FIGS. 3a and 3b show examples of signal constellations as established in the DVB-T standard" and states in ¶ [0015] that the details are "described in the ETSI standard EN 300744".

Applicant submits herewith via Information Disclosure Statement a copy of the Final Draft ETSI EN 300 744 v1.5.1 dated (2004-06) and entitled "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television" (hereafter "DVB-T modulation and coding standard"). The DVB-T modulation and coding standard describes the real and imaginary components of the two supported non-uniform 16-QAM constellations on p. 21 as follows:

> 4.3.5 Signal constellations and mapping
>
> ...
>
> The exact values of the constellation points are $z \in \{n + j\, m\}$ with values of n, m given below for the various constellations:
>
> ...
>
> **Non-uniform 16-QAM with $\alpha = 2$**
>
> $n \in \{-4, -2, 2, 4\}, m \in \{-4, -2, 2, 4\}$
>
> **Non-uniform 16-QAM with $\alpha = 4$**
>
> $n \in \{-6, -4, 4, 6\}, m \in \{-6, -4, 4, 6\}$

Accordingly, the standard confirms that the non-uniform QAM-16 illustrated in Fig. 3A of Eberlein (corresponding to the 16-QAM α = 4 constellation from the DVB-T modulation and coding standard) is formed by orthogonalizing a PAM-4 {-6, -4, 4, 6} to yield 16 constellation points that each have different locations and different labels (i.e. (n, m) ∈ { (-6, -6), (-6, -4), (-6, 4), (-6, 6), (-4, -6), (-4, -4), (-4, 4), (4, 6), (4, -6), (4, -4), (4, 4), (4, 6), (-6, -6), (6, -4), (6, 4), and (6, 6) }).

Applicant respectfully submits that the M.P.E.P. warns that in applying the broadest reasonable interpretation during examination, "[C]laims are not to be read in a vacuum". M.P.E.P. § 2111. Accordingly, Applicant submits that the assertion in the Office action that points that simply occupy the same quadrant (e.g. have both positive real and imaginary components) have identical locations is inappropriate. Such points clearly do not have identical locations. As such, Applicant respectfully submits that the rejections of

claims 1, 10, 11, 20, 21, and 22 must be withdrawn. Furthermore, Applicant respectfully submits that claims 1, 10, 11, 20, 21, and 22 are allowable over the prior art of record.

REJECTION OF CLAIMS 9, 19, AND 30 ARE REJECTED UNDER 35 U.S.C. §103(a)

Claims 9, 19, and 30 are rejected under pre-AIA 35 U.S.C. §103(a). Claims 9, 19, and 30 depend from claims 1, 11, and 22 respectively. For reasons similar to those outlined above with respect to claims 1, 11, and 22, Applicant respectfully submits that claims 9, 19, and 30 are allowable over the prior art of record. Accordingly, Applicant respectfully requests the withdrawal of the rejection of claims 9, 19, and 30 under 35 U.S.C. §103(a).

CONCLUSION

In view of the foregoing amendments and remarks, Applicant respectfully requests the withdraw of the rejections of claims 1 – 30 and respectfully submits that this application is in condition for allowance. Accordingly, reconsideration of the application and allowance of the claims are respectfully requested. Although the present paper may include a combination of alterations to the application or claims, or characterizations of claim scope or referenced art, Applicant is not conceding that previously pending claims in this application are not patentable over the cited references. Rather, any alterations and/or characterizations are made to strictly facilitate prosecution of this application. As such, reviewers of this or any parent, child, or related prosecution history shall not reasonably infer that Applicant has made any disclaimers or disavowals of any subject matter supported by the present application.

Applicant has submitted arguments believed to be sufficient to overcome all of the outstanding rejections. Consequently, Applicant has not advanced every argument for the allowability of the claims over the references of record. As such, Applicant does not acquiesce to any of the Examiner's statements or characterizations not specifically traversed. If the Examiner should have any remaining questions or objections, a telephone interview to discuss and resolve these issues is respectfully requested.

The Commissioner is authorized to charge any fee under 37 CFR 1.16 and 1.17 that may be required during the **entire pendency** of the application to Deposit Account No. 50-4407. Please show our docket number with any transaction to the Deposit Account.

Respectfully submitted,
KPPB LLP

By /David J. Bailey/
David J. Bailey
Registration No. 59,632
949.852.0000

DJB/kl

Electronic Acknowledgement Receipt	
EFS ID:	39962371
Application Number:	16728384
International Application Number:	
Confirmation Number:	6350
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Customer Number:	71897
Filer:	David J. Bailey/Katrina Lee
Filer Authorized By:	David J. Bailey
Attorney Docket Number:	C44-00999.CON15
Receipt Date:	09-JUL-2020
Filing Date:	27-DEC-2019
Time Stamp:	22:53:08
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1		00999con15amendment.pdf	488655 dd1f16fd6ad32be4126b015d447a72a6b7062efd	yes	28

	Multipart Description/PDF files in .zip description		
	Document Description	**Start**	**End**
	Amendment/Req. Reconsideration-After Non-Final Reject	1	1
	Specification	2	2
	Claims	3	17
	Applicant Arguments/Remarks Made in an Amendment	18	28
Warnings:			
Information:			
	Total Files Size (in bytes):	488655	

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

<u>New Applications Under 35 U.S.C. 111</u>
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.
<u>National Stage of an International Application under 35 U.S.C. 371</u>
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.
<u>New International Application Filed with the USPTO as a Receiving Office</u>
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

PTO/SB/06 (09-11)
Approved for use through 1/31/2014. OMB 0651-0032
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it displays a valid OMB control number.

PATENT APPLICATION FEE DETERMINATION RECORD Substitute for Form PTO-875	Application or Docket Number 16/728,384	Filing Date 12/27/2019	☐ To be Mailed

ENTITY: ☑ LARGE ☐ SMALL ☐ MICRO

APPLICATION AS FILED - PART I

	(Column 1)	(Column 2)			
FOR	NUMBER FILED	NUMBER EXTRA		RATE ($)	FEE ($)
☐ BASIC FEE (37 CFR 1.16(a), (b), or (c))	N/A	N/A		N/A	
☐ SEARCH FEE (37 CFR 1.16(k), (i), or (m))	N/A	N/A		N/A	
☐ EXAMINATION FEE (37 CFR 1.16(o), (p), or (q))	N/A	N/A		N/A	
TOTAL CLAIMS (37 CFR 1.16(i))	minus 20 =	*		x $100 =	
INDEPENDENT CLAIMS (37 CFR 1.16(h))	minus 3 =	*		x $460 =	
☐ APPLICATION SIZE FEE (37 CFR 1.16(s))	If the specification and drawings exceed 100 sheets of paper, the application size fee due is $310 ($155 for small entity) for each additional 50 sheets or fraction thereof. See 35 U.S.C. 41(a)(1)(G) and 37 CFR 1.16(s).				
☑ MULTIPLE DEPENDENT CLAIM PRESENT (37 CFR 1.16(j))					820
* If the difference in column 1 is less than zero, enter "0" in column 2.				TOTAL	820

APPLICATION AS AMENDED - PART II

AMENDMENT		(Column 1)		(Column 2)	(Column 3)			
	07/09/2020	CLAIMS REMAINING AFTER AMENDMENT		HIGHEST NUMBER PREVIOUSLY PAID FOR	PRESENT EXTRA		RATE ($)	ADDITIONAL FEE ($)
	Total (37 CFR 1.16(i))	* 30	Minus	** 30	= 0		x $100 =	0
	Independent (37 CFR 1.16(h))	* 3	Minus	*** 3	= 0		x $460 =	0
	☐ Application Size Fee (37 CFR 1.16(s))							
	☐ FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							
							TOTAL ADD'L FEE	0

AMENDMENT		(Column 1)		(Column 2)	(Column 3)			
		CLAIMS REMAINING AFTER AMENDMENT		HIGHEST NUMBER PREVIOUSLY PAID FOR	PRESENT EXTRA		RATE ($)	ADDITIONAL FEE ($)
	Total (37 CFR 1.16(i))	*	Minus	**	=		x $0 =	
	Independent (37 CFR 1.16(h))	*	Minus	***	=		x $0 =	
	☐ Application Size Fee (37 CFR 1.16(s))							
	☐ FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							
							TOTAL ADD'L FEE	

* If the entry in column 1 is less than the entry in column 2, write "0" in column 3.	SLIE
** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 20, enter "20".	/LISA E FULTON/
*** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 3, enter "3".	
The "Highest Number Previously Paid For" (Total or Independent) is the highest number found in the appropriate box in column 1.	

This collection of information is required by 37 CFR 1.16. The information is required to obtain or retain a benefit by the public which is to file (and by the USPTO to process) an application. Confidentiality is governed by 35 U.S.C. 122 and 37 CFR 1.14. This collection is estimated to take 12 minutes to complete, including gathering, preparing, and submitting the completed application form to the USPTO. Time will vary depending upon the individual case. Any comments on the amount of time you require to complete this form and/or suggestions for reducing this burden, should be sent to the Chief Information Officer, U.S. Patent and Trademark Office, U.S. Department of Commerce, P.O. Box 1450, Alexandria, VA 22313-1450. DO NOT SEND FEES OR COMPLETED FORMS TO THIS ADDRESS. **SEND TO: Commissioner for Patents, P.O. Box 1450, Alexandria, VA 22313-1450.**

If you need assistance in completing the form, call 1-800-PTO-9199 and select option 2.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

U.S.PATENTS

Examiner Initial*	Cite No	Patent Number	Kind Code[1]	Issue Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		10548031	B2	2020-01-28	Barsoum et al.	
		10567980	B2	2020-02-18	Barsoum et al.	
		10693700	B1	2020-06-23	Barsoum et al.	
		10694403	B2	2020-06-23	Barsoum et al.	
		10701570	B2	2020-06-30	Barsoum et al.	
		10708794		2020-07-07	Barsoum et al.	

U.S.PATENT APPLICATION PUBLICATIONS

Examiner Initial*	Cite No	Publication Number	Kind Code[1]	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		20050268206	A1	2005-12-01	Tran et al.	
		20060085720	A1	2006-04-20	Tran et al.	
		20060165190	A1	2006-07-27	Tamaki et al.	
		20070211822	A1	2007-09-13	Olesen et al.	
		20200137595	A1	2020-04-30	Barsoum et al.	
		20200145276	A1	2020- 05-07	Barsoum et al.	
		20200145844	A1	2020- 05-07	Barsoum et al.	
		20200145845	A1	2020- 05-07	Barsoum et al.	
		20200145847	A1	2020- 05-07	Barsoum et al.	
		20200145848	A1	2020- 05-07	Barsoum et al.	
		20200145849	A1	2020- 05-07	Barsoum et al.	
		20200145850	A1	2020- 05-07	Barsoum et al.	
		20200162942	A1	2020- 05-21	Barsoum et al.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

FOREIGN PATENT DOCUMENTS								
Examiner Initial*	Cite No	Foreign Document Number[3]	Country Code[2]	Kind Code[4]	Publication Date	Name of Patentee or Applicant of cited Document	Pages,Columns,Lines where Relevant Passages or Relevant Figures Appear	T[5]

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T[5]
		"Final draft ETSI EN 300 744 V1.4.1", European Standard (Telecommunication series), Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television, European Telecommunications Standards Institute, European Broadcasting Union, August 2000.	
		"Final draft ETSI EN 300 744 V1.5.1", European Standard (Telecommunication series), Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television, European Telecommunications Standards Institute, European Broadcasting Union, June 2004.	
		DE GAUDENZI ET AL., "A New Coded Digital Modulation Scheme for Nonlinear Satellite Channels, with High Power-and Spectral-Efficiency", European Space Agency, STR-242, July 2001, 60 pgs.	
		DE GAUDENZI ET AL., "Analysis and Design of an All-Digital Demodulator for Trellis Coded 16-QAM Transmission over a Nonlinear Satellite Channel", IEEE Transactions on Communications, Vol. 43, Nos. 2, 3, 4, February, March, April 1995, pp. 659-668.	
		In re Gilbert P. Hyatt, United States Court of Appeals Federal Circuit, 708 F.2d 712, June 6, 1983.	

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

EXAMINER SIGNATURE			
Examiner Signature		Date Considered	

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

Electronic Patent Application Fee Transmittal

Application Number:	16728384
Filing Date:	27-Dec-2019
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Filer:	Isaac M. Fine/Katrina Lee
Attorney Docket Number:	C44-00999.CON15

Filed as Large Entity

Filing Fees for Utility under 35 USC 111(a)

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Basic Filing:				
Pages:				
Claims:				
Miscellaneous-Filing:				
Petition:				
Patent-Appeals-and-Interference:				
Post-Allowance-and-Post-Issuance:				
Extension-of-Time:				

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Miscellaneous:				
SUBMISSION- INFORMATION DISCLOSURE STMT	1806	1	240	240
		Total in USD ($)		**240**

Electronic Acknowledgement Receipt	
EFS ID:	39973453
Application Number:	16728384
International Application Number:	
Confirmation Number:	6350
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Customer Number:	71897
Filer:	Isaac M. Fine/Katrina Lee
Filer Authorized By:	Isaac M. Fine
Attorney Docket Number:	C44-00999.CON15
Receipt Date:	10-JUL-2020
Filing Date:	27-DEC-2019
Time Stamp:	20:38:05
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	yes
Payment Type	CARD
Payment was successfully received in RAM	$240
RAM confirmation Number	E202070K38361268
Deposit Account	504407
Authorized User	Katrina Lee
The Director of the USPTO is hereby authorized to charge indicated fees and credit any overpayment as follows: 37 CFR 1.21 (Miscellaneous fees and charges) 37 CFR 1.20 (Post Issuance fees)	

37 CFR 1.19 (Document supply fees)

37 CFR 1.17 (Patent application and reexamination processing fees)

37 CFR 1.16 (National application filing, search, and examination fees)

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Non Patent Literature	etsien300744.pdf	9672135 4a36fdffb234388ebdc5e46ffcd58b6f3de4fa71	no	50
Warnings:					
Information:					
2	Non Patent Literature	etsien300744v5.pdf	12806007 c59f66ef8bb988fa1747d0cb1cf43555422c5e09	no	64
Warnings:					
Information:					
3	Non Patent Literature	degaudenzianewcoded60pgs.pdf	6934716 0285b579d869071125c482bdac9c32a8dd0d8e30	no	60
Warnings:					
Information:					
4	Non Patent Literature	degaudenzietalanalysis.pdf	3577008 06fcee7586f68e4964681b4532cdd12021b6473f	no	10
Warnings:					
Information:					
5	Non Patent Literature	inrehyatt.pdf	1292941 af9bc6551766467faa446a3eeae8d43c00eb2864	no	4
Warnings:					
Information:					
6	Transmittal Letter	00999con15idsxtml.pdf	121271 94ef5d957a8a0ed25159a5e78b3bd689038d6547	no	2
Warnings:					
Information:					

7	Information Disclosure Statement (IDS) Form (SB08)	00999con15ids1449.pdf	116966 b94328fb757bdcaae489827c20644c96ff8fdff8	no	3
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					
8	Fee Worksheet (SB06)	fee-info.pdf	30736 f558e09ae961bad5b15f0c1a1710cf7d7ed094c8	no	2
Warnings:					
Information:					
		Total Files Size (in bytes):	34551780		

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

<u>New Applications Under 35 U.S.C. 111</u>
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.
<u>National Stage of an International Application under 35 U.S.C. 371</u>
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.
<u>New International Application Filed with the USPTO as a Receiving Office</u>
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF EFS WEB TRANSMISSION

I hereby certify that this correspondence is being electronically filed with the United States Patent and Trademark Office on July 10, 2020 at or before 11:59 p.m. Pacific Time under the Rules of 37 C.F.R. 1.8.

/Katrina Lee/
Katrina Lee

Appl No.	: 16/728,384	Confirmation No. 6350
Applicant	: Constellation Designs, LLC	
Filed	: December 27, 2019	
Title	: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	
TC/A.U.	: 2633	
Examiner	: Tesfaldet Bocure	
Docket No.	: C44-00999.CON15	
Customer No.	: 71897	

INFORMATION DISCLOSURE STATEMENT § 1.97(c)

MAIL STOP AMENDMENT
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806
July 10, 2020

Commissioner

In compliance with the duty of disclosure under 37 C.F.R. §§ 1.56, 1.97 and 1.98, and in accordance with the provisions in the MPEP §§ 609 and 707.05(b), enclosed is FORM PTO/SB/08A/B with a listing of references that are known to applicant. Copies of each of the references listed under "Non-Patent Literature Documents" are enclosed.

It is respectfully requested that these references be considered in the examination of this application and identified on the list of references cited on the patent issuing on this application. Applicant also requests that an initialed copy of said FORM PTO/SB/08A/B be entered in the application file, and returned to the Applicant with the next official communication from the Office in accordance with MPEP § 609.

The $240 processing fee will be paid via credit card. If there are any other fees due in connection with this filing, the Commissioner is authorized to charge Deposit Account 50-4407. Please show our docket number with any charge to the Deposit Account.

Respectfully submitted,

KPPB LLP

/Isaac M. Fine/
Isaac M. Fine
Registration No. 74,274
949.852.0000

IMF/kl

Enclosures

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

U.S.PATENTS

Examiner Initial*	Cite No	Patent Number	Kind Code[1]	Issue Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		7123663	B2	2006-10-17	De, Gaudenzi R., et al.	
		7359426	B2	2008-04-15	Ojard, Eric	
		7539261	B2	2009- 05-26	Lu, Ben et al.	
		7599420	B2	2009-10-06	Forenza, Antonio et al.	
		7907641	B2	2011-03-15	Sun, Feng-wen et al.	
		7908541	B2	2011-03-15	Kyung, Gyu-bum et al.	
		8031793	B2	2011-10-04	Ionescu, Dumitru M., et al.	
		8102947	B2	2012-01-24	Eroz, Mustafa et al.	
		8111770	B1	2012-02-07	Moon, Jaekyun et al.	
		8160121	B2	2012-04-17	Forenza, Antonio et al.	
		8171383	B2	2012- 05-01	Landau, Uri M., et al.	
		8199847	B2	2012-06-12	Lee, Lin-nan et al.	
		8428162	B2	2013-04-23	Forenza, Antonio et al.	
		8483145	B2	2013-07-09	Astely, David	

U.S.PATENT APPLICATION PUBLICATIONS

Examiner Initial*	Cite No	Publication Number	Kind Code[1]	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or Relevant Figures Appear
		20050141627	A1	2005-06-30	Walker, Glenn A., et al.	
		20050143004	A1	2005-06-30	Dibiaso, Eric A., et al.	
		20050169400	A1	2005-08-04	Chouly, Antoine et al.	
		20060276145	A1	2006-12-07	Walker, Glenn A., et al.	
		20100077275	A1	2010-03-25	Yu, Qian et al.	
		20200221321	A1	2020-07-09	Barsoum et al.	

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

FOREIGN PATENT DOCUMENTS

Examiner Initial*	Cite No	Foreign Document Number[3]	Country Code[2]	Kind Code[4]	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines where Relevant Passages or Relevant Figures Appear	T[5]
		100630177	KR	B1	2006-09-29	Jeong, Hong S., et al.		
		100634250	KR	B1	2006-10-16	Eroz, Mustafa et al.		
		100723018	KR	B1	2007- 05-30	Lee, In K., et al.		
		100471191	CN	C	2009-03-18	Noll, Barreto A.		
		100584011	CN	C	2010-01-20	Lin, Yang et al.		
		100589469	CN	C	2010-02-10	Ying, Li et al.		
		4554610	JP	B2	2010-07-23	Visoz, Raphaël et al.		
		1670168	EP	B1	2010-07-28	Brunel, Loïc et al.		
		101133558	CN	B	2010-10-06	Yuan, Li et al.		
		2428796	RU	C2	2011-09-10	Kkhan, Faruk et al.		
		4920977	JP	B2	2012-02-10	Wang, Xiaodong et al.		
		5129323	JP	B2	2012-11-09	Malladi, Durga P.		
		101282522	KR	B1	2013-07-04	Kim, Bong H., et al.		
		2491742	RU	C2	2013-08-27	Shtadel, Majer L., et al.		
		101322322	CN	B	2013-11-13	Smee, John E., et al.		
		101346423	KR	B1	2014-01-02	Kim, Bong H., et al.		
		102017445	CN	B	2014- 05-07	Roh, Dong-wook et al.		
		101507782	KR	B1	2015-03-26	LG Electronics Co., Ltd		
		2695799	CA	C	2016-10-11	Forenza, Antonio et al.		
		103501193	CN	B	2017-04-12	Antonio, Forenza et al.		
		2712914	ES	T3	2019- 05-16	Barsoum, Maged F., et al.		
		40011480	HK	A	2020-07-17	Barsoum et al.		

PTO/SB/08a (01-10)
Approved for use through 07/31/2012. OMB 0651-0031
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it contains a valid OMB control number.

INFORMATION DISCLOSURE STATEMENT BY APPLICANT (Not for submission under 37 CFR 1.99)		
	Application Number	16/728,384
	Filing Date	Dec 27, 2019
	First Named Inventor	Maged F. Barsoum
	Art Unit	2633
	Examiner Name	Tesfaldet Bocure
	Attorney Docket Number	C44-00999.CON15

NON-PATENT LITERATURE DOCUMENTS			
Examiner Initials*	Cite No	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T [5]
		"Frame Structure Channel Coding and Modulation for a Second Generation Digital Terrestrial Television Broadcasting System (DVB-T2)", Digital Video Broadcasting, DVB Document A122, June 2008, 158 pgs.	

EXAMINER SIGNATURE			
Examiner Signature		Date Considered	

*EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through a citation if not in conformance and not considered. Include copy of this form with next communication to applicant.

[1] See Kind Codes of USPTO Patent Documents at www.USPTO.GOV or MPEP 901.04. [2] Enter office that issued the document, by the two-letter code (WIPO Standard ST.3). [3] For Japanese patent documents, the indication of the year of the reign of the Emperor must precede the serial number of the patent document.
[4] Kind of document by the appropriate symbols as indicated on the document under WIPO Standard ST.16 if possible. [5] Applicant is to place a check mark here if English language translation is attached.

Electronic Patent Application Fee Transmittal

Application Number:	16728384
Filing Date:	27-Dec-2019
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Filer:	Isaac M. Fine/Katrina Lee
Attorney Docket Number:	C44-00999.CON15

Filed as Large Entity

Filing Fees for Utility under 35 USC 111(a)

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Basic Filing:				
Pages:				
Claims:				
Miscellaneous-Filing:				
Petition:				
Patent-Appeals-and-Interference:				
Post-Allowance-and-Post-Issuance:				
Extension-of-Time:				

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Miscellaneous:				
SUBMISSION- INFORMATION DISCLOSURE STMT	1806	1	240	240
		Total in USD ($)		**240**

Electronic Acknowledgement Receipt	
EFS ID:	40245197
Application Number:	16728384
International Application Number:	
Confirmation Number:	6350
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations
First Named Inventor/Applicant Name:	Maged F. Barsoum
Customer Number:	71897
Filer:	Isaac M. Fine/Katrina Lee
Filer Authorized By:	Isaac M. Fine
Attorney Docket Number:	C44-00999.CON15
Receipt Date:	10-AUG-2020
Filing Date:	27-DEC-2019
Time Stamp:	22:29:48
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	yes
Payment Type	CARD
Payment was successfully received in RAM	$240
RAM confirmation Number	E202080M30166687
Deposit Account	504407
Authorized User	Katrina Lee
The Director of the USPTO is hereby authorized to charge indicated fees and credit any overpayment as follows: 37 CFR 1.16 (National application filing, search, and examination fees) 37 CFR 1.17 (Patent application and reexamination processing fees)	

37 CFR 1.19 (Document supply fees)

37 CFR 1.20 (Post Issuance fees)

37 CFR 1.21 (Miscellaneous fees and charges)

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Foreign Reference	KR100630177.pdf	8281898 8c412c025b4f6a65af60b32c23d09ca5072fd6f7	no	87
Warnings:					
Information:					
2	Foreign Reference	KR100634250.pdf	5004564 39f02557fc6d41eb9b1682cbb3a75340e20a5406	no	53
Warnings:					
Information:					
3	Foreign Reference	KR100723018.pdf	3611716 966a2f885cbf2f79174ce9053779097cb5f339c9	no	38
Warnings:					
Information:					
4	Foreign Reference	CN100471191.pdf	986009 c557ab1d55d4a42f79928cab6a795f8ae26f0db9	no	32
Warnings:					
Information:					
5	Foreign Reference	CN100584011.pdf	3209526 a8b829e0c4c45a0590d6bbba6a18705604aedc7a	no	130
Warnings:					
Information:					
6	Foreign Reference	CN100589469.pdf	1546907 98035f82a00a42023a8d847bb3c0854462db04d9	no	56
Warnings:					
Information:					

7	Foreign Reference	JP4554610.pdf	5136829 602c78fa94499dbfe312930974892b25dfc3fd35	no	49
Warnings:					
Information:					
8	Foreign Reference	EP1670168.pdf	6883289 bd4cc3962ab5edc3aa1fa659116b01bbc541134a	no	21
Warnings:					
Information:					
9	Foreign Reference	CN101133558.pdf	1011813 fc0f9b22c3b1d83f8ab6362eff761bf77623a5e6	no	37
Warnings:					
Information:					
10	Foreign Reference	RU2428796.pdf	2659907 3ec3e629d402afa890568b49ee82f641f054ff56	no	156
Warnings:					
Information:					
11	Foreign Reference	JP4920977.pdf	5575163 9fc8ab50d946a487ce0af217633569ca9a144506	no	53
Warnings:					
Information:					
12	Foreign Reference	JP5129323.pdf	11301793 aaa536f4f267bf1063041a5a442f56be5cdc2269	no	88
Warnings:					
Information:					
13	Foreign Reference	KR101282522.pdf	2555430 b5b2815791737f6b20fc085beda3aab320a27d55	no	35
Warnings:					
Information:					

14	Foreign Reference	RU2491742.pdf	3370929 0ff0f8a31da856eb18c796f4cffc7637bd5e10c7	no	173
Warnings:					
Information:					
15	Foreign Reference	CN101322322.pdf	5702352 859d61385a723a8f6565e5c53dd6ab98362edd0e	no	152
Warnings:					
Information:					
16	Foreign Reference	KR101346423.pdf	3172757 dc8bc17a05a62b2fecd17d20683f8efabf3eabb1	no	36
Warnings:					
Information:					
17	Foreign Reference	CN102017445PART1.pdf	9583775 c83e4d49a9578f943dc13305e2e905dda1536da1	no	121
Warnings:					
Information:					
18	Foreign Reference	CN102017445PART2.pdf	14121599 6175038e46df5bd0f89a62bd137ba44f13ed1732	no	66
Warnings:					
Information:					
19	Foreign Reference	KR101507782B1.pdf	4100231 3114e31ce241a8f2034d5c5789c4c08b1ecb163e	no	52
Warnings:					
Information:					
20	Foreign Reference	CA2695799C.pdf	5610888 91a58cfea04a9c74bc35f9f4db895a9492442db8	no	167
Warnings:					
Information:					

21	Foreign Reference	CN103501193PART1.pdf	10423855 ade1a7e454e6f402478396d1a47b44c746c78816	no	124
Warnings:					
Information:					
22	Foreign Reference	CN103501193PART2.pdf	13059094 2efd140f0c71139502f54da00a69e041073bfc1b	no	61
Warnings:					
Information:					
23	Foreign Reference	ES2712914.pdf	2978143 a0f817af6a3e33650ac12051e723ab9a2a3970aa	no	123
Warnings:					
Information:					
24	Foreign Reference	HK40011480A1.pdf	8153954 9b7f3868807b308524a346724764dc2ac8c97f6d	no	62
Warnings:					
Information:					
25	Non Patent Literature	framestructureDVB.pdf	2256817 e38edb519bfb0e13fffd6a1907d1d4813a15fc57	no	158
Warnings:					
Information:					
26	Transmittal Letter	00999con15idsxtml.pdf	121163 14f7cca624942eb8d5007e5368d93417fde6b13e	no	2
Warnings:					
Information:					
27	Information Disclosure Statement (IDS) Form (SB08)	00999con15ids1449.pdf	121130 51489f700b56406ff4f8a5671f9dfcef6f20b376	no	3
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					

28	Fee Worksheet (SB06)	fee-info.pdf	30737 aaa4a200883b708616ea02e8515254921c7a2b12	no	2

Warnings:	
Information:	
Total Files Size (in bytes):	140572268

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

New Applications Under 35 U.S.C. 111

If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371

If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office

If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

KR100630177B1

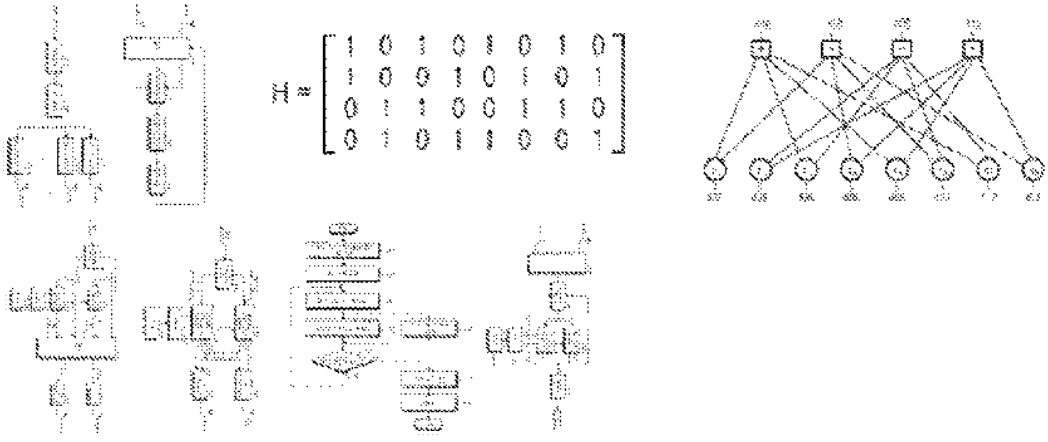
Apparatus and method for encoding/decoding space time low density parity check code with full diversity gain

Abstract

The present invention provides a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas. When information data is received, an even number of a first partial matrix corresponding to the information data is received. A fifth partial matrix comprising a second partial matrix composed of columns and a fourth partial matrix composed of odd-numbered columns of a third partial matrix corresponding to parity, and a sixth partial matrix composed of odd-numbered columns of the first partial matrix And an eighth partial matrix, which is a matrix combining a seventh partial matrix consisting of even-numbered columns of the third partial matrix, is preset on a ninth partial matrix and a binary field in which the first partial matrix and the third partial matrix are exclusively ORed. The information data is encoded according to a rank parity check matrix and generated as a Low Density Parity Check (LDPC) code. The LDPC code is spatially mapped according to a preset spatial mapping method to generate a space-time LDPC code.

Space-Time LDPC Code, Parity Check Matrix, Full Diversity, Full Rank, BPSK, QPSK, 4-QAM

Images (8)



Classifications

H04L1/0057 Block codes
View 6 more classifications
South Korea
Download PDF Find Prior Art Similar
Other languages
Korean
Inventor
김재열
윤석현
정홍실
채찬병

Worldwide applications
2005 ~~KR~~ US

Application KR1020050010836A events

2004-02-06

Priority to KR1020040007978

2004-02-06

Priority to KR20040007978

2005-02-04

Application filed by 삼성전자주식회사

2006-05-12

Publication of KR20060041769A

2006-09-29

Application granted

2006-09-29

Publication of KR100630177B1

Info

Patent citations (1)

Cited by (29)

Legal events

Similar documents

Priority and Related Applications

External links

Espacenet
Global Dossier
Discuss

Description

Apparatus and method for space-time low-density parity check code encoding / decoding having maximum diversity gain {APPARATUS AND METHOD FOR ENCODING / DECODING SPACE TIME LOW DENSITY PARITY CHECK CODE WITH FULL DIVERSITY GAIN}

1 schematically illustrates a transmitter structure of a mobile communication system using a general STC scheme.

2 is a diagram schematically illustrating a receiver structure of a mobile communication system using a general STC scheme.

3 illustrates a parity check matrix of a general (8, 2, 4) LDPC code.

4 is a diagram illustrating a factor graph of the (8, 2, 4) LDPC code of FIG.

5 is a diagram schematically illustrating a structure of a transmitter for performing a function in an embodiment of the present invention when using a BPSK scheme as a signal mapping scheme.

FIG. 6 schematically illustrates a transmitter structure for performing a function in an embodiment of the present invention when using a QPSK scheme and a 4-QAM scheme as a signal mapping scheme. FIG.

7 is a flowchart illustrating a receiver operation process of a mobile communication system according to an embodiment of the present invention.

8 schematically illustrates a receiver structure of a mobile communication system that performs a function in an embodiment of the present invention.

The present invention relates to a mobile communication system, and more particularly, to an apparatus and method for encoding / decoding a space-time low density parity check code for maximizing diversity gain.

The most fundamental problem in communication is how efficiently and reliably data can be transmitted over a channel. In the next generation multimedia mobile communication system, which is being actively researched recently, a channel communication technique suitable for the system is used as a high speed communication system capable of processing and transmitting various information such as video and wireless data beyond the initial voice-oriented service is required. It is essential to increase the efficiency of the system.

However, unlike a wired channel environment, a wireless channel environment exists in a mobile communication system such as multipath interference, shadowing, propagation attenuation, time-varying noise, interference, and fading. Many factors lead to unavoidable errors and loss of information. The loss of information causes severe distortion in the actual transmission signal, thereby degrading the overall performance of the mobile communication system. In general, in order to reduce such information loss, various error-control techniques are used to increase the reliability of the mobile communication system according to the characteristics of the channel. The most basic of these error control techniques is an error correction code. (error-correcting code).

In addition, in the mobile communication system, a diversity scheme is used to remove instability of the communication due to the fading phenomenon, and the space diversity scheme, which is one of such diversity schemes, is a multiple antenna.). Here, the spatial diversity scheme includes a receive antenna diversity scheme using a plurality of receive antennas, a transmit antenna diversity scheme using a plurality of transmit antennas, and a

multiple input using a plurality of receive antennas and a plurality of transmit antennas. It is classified into a multiple input multiple output (MIMO) method. Here, the MIMO scheme is a kind of space-time coding (STC) scheme, and the STC scheme transmits a signal encoded by a predetermined encoding scheme using a plurality of transmit antennas. By extending the coding scheme in the time domain to the space domain, a lower error rate is achieved.

Next, a structure of a mobile communication system transmitter using the STC scheme will be described with reference to FIG. 1.

1 is a diagram schematically illustrating a transmitter structure of a mobile communication system using a general STC scheme.

Referring to FIG. 1, the transmitter includes an encoder 110, a spatial mapper 120, and a plurality of, for example, M signal mappers, that is, a first signal. The signal mapper 130-1 to the M-th signal mapper 130 -M, and a plurality of M transmit antennas Tx.ANT, that is, a first transmission antenna Tx.ANT 1 to It consists of an Mth transmit antenna (Tx.ANT M). First, when information data (information data) is input, the encoder 110 generates the coded symbol by encoding the information data in a predetermined coding scheme according to a preset coding rate (code rate) Then output to the space mapper 120. As an example, when the information data is k bits and the coding rate is k / N, the encoder 110 outputs the k bits of information data as encoded symbols consisting of N bits.

The spatial mapper 120 inputs the coded symbols output from the encoder 110 and performs spatial mapping in a preset spatial mapping method so that the spatial mapper can be transmitted through the M transmission antennas. Control each of the signals to be output to each of the corresponding signal mappers. Here, the spatial mapping method used by the spatial mapper 120 may exist in various ways. In FIG. 1, the encoder 110 outputs the coded symbols output from the encoder 110 by the spatial mapper 120. Let's assume that the space maps in the order printed by. For example, the encoded symbol output from the encoder 110 is

$$C=[c_1,c_2,\cdots,c_N]$$

Assume that, the output of the spatial mapper 120 is the same as Equation 1

$$M \times \frac{N}{M}$$

It becomes a space-time codeword matrix.

$$C = \begin{bmatrix} c_1 & c_{M+1} & \cdots & c_{N-M+1} \\ c_2 & c_{M+2} & \cdots & c_{N-M+2} \\ \cdots & & & \\ c_M & c_{M+M} & \cdots & c_N \end{bmatrix}$$

Since the number of transmit antennas is M, the spatial mapper 120 spatially maps the coded symbols output from the encoder 110 in order according to the output order thereof, and classifies the M symbols into M signals. Each of them is output to each of the first signal mapper 130-1 to the M-th signal mapper 130 -M. Then, each of the first signal mapper 130-1 to the M-th signal mapper 130 -M inputs a signal output from the spatial mapper 120 and maps the signal in a preset manner. Transmit over the air through each of the transmitting antennas, that is, the first to Mth transmitting antennas. Here, each of the first signal mapper 130-1 to the M-th signal mapper 130 -M maps an input signal according to a constellation according to the setting method and transmits the first transmission Output to each of the antenna to the Mth transmit antenna. For example, each of the first signal mapper 130-1 to the M-th signal mapper 130 -M may be a carrier having a constant amplitude and phase when the number of bits n of the input signal is 1 (n = 1). a signal is mapped to the input signal by a binary phase shift keying (BPSK) method in which a phase is converted to correspond to a transmission code by using a carrier, and when the number of bits n of the input signal is 2 (n = 2) The input signal is mapped using the Quadrature Phase Shift Keying (QPSK) scheme, and the input signal is mapped using the Quadrature Amplitude Modulation (QAM) scheme when the number of bits n of the input signal is 3 (n = 3). In FIG. 1, since k-bit information data is transmitted through M transmit antennas, the coding rate is

$$M \times \frac{k}{N}$$

Becomes

In FIG. 1, a transmitter structure of a mobile communication system using a general STC method has been described. Next, a receiver structure of a mobile communication system using a general STC method will be described with reference to FIG. 2.

2 is a diagram schematically illustrating a receiver structure of a mobile communication system using a general STC scheme.

Referring to FIG. 2, the receiver includes a plurality of P reception antennas (Rx.ANT), that is, a first reception antenna (Rx.ANT 1) to a P-th reception antenna (Rx.ANT P), A detector 210, a spatial de-mapper 220, a space time decoder 230, and a spatial mapper 240 are included. In FIG. 2, it is assumed that the number of transmit antennas of the transmitter corresponding to the receiver is different from the number of receive antennas of the receiver. That is, the number of transmit antennas of the transmitter is M as described in FIG. Although the number of receiving antennas of the receiver is assumed to be P, the number of transmitting antennas of the transmitter and the number of receiving antennas of the receiver may be the same.

First, a signal transmitted through a plurality of transmitters, i.e., M transmit antennas, is received through each of the first to Pth receiving antennas. Each of the first to P-th receiving antennas outputs the received signal to the detector 210. The detector 210 detects a transmission signal transmitted from each of the first to Mth antennas by inputting a received signal output from each of the first to Pth antennas, and then detects the space history (220).

The spatial history controller 220 inputs the signal output from the detector 210 and outputs the signal to the space-time decoder 230 by performing a spatial history on the spatial history method corresponding to the spatial mapping method applied by the transmitter. The space-time decoder 230 inputs the signal output from the space inversion unit 220, decodes the signal by a decoding method corresponding to the encoding method applied by the transmitter, and outputs the decoded signal. In this case, the signal output from the space-time decoder 230 is the same as the information data in the transmitter when there is no error due to a radio channel environment, but the error occurs in the actual radio channel environment, and thus iterative decoding for more reliable decoding. (iterative decoding) may be performed.

For the iterative decoding, the space-time decoder 230 outputs the decoded signal to the spatial mapper 240 in a manner corresponding to the coding scheme applied by the transmitter. The spatial mapper 240 inputs the signal output from the space-time decoder 230, performs space-time decoding in a spatial mapping method corresponding to the spatial mapping method applied by the transmitter, and then outputs the same to the detector 210 for repeated decoding. Control to perform. When the iterative decoding is performed as described above, the reliability of the restoration of the information data is increased. When it is decoded as a reliable signal by the space-time decoder 230 through the repeated decoding, the space-time decoder 230 outputs the reliable signal. The signal is output as final information data.

Meanwhile, the space-time code represents a code coded by the STC scheme, and the STC scheme additionally obtains additional information by extending the coding scheme in the time domain to the space domain as described above. That is, the STC scheme not only reduces error rate by providing additional information in the time domain but also increases accuracy of diversity gain in the spatial domain, thereby enabling more accurate signal demodulation. Here, the diversity gain corresponds to the slope of the signal-to-noise ratio (SNR) of the space-time code and the error probability in logarithmic scale. When designing a code, maximizing the diversity gain (hereinafter, referred to as 'full diversity gain') serves as the most important factor in determining the performance of the space-time code.

In order to obtain the maximum diversity gain, in designing the space-time code, the following space-time code design condition 1 to space-time code design condition 3 must be satisfied.

[Space-Time Code Design Condition 1]

In order to obtain the maximum diversity gain, the matrix B (c, e) must have a maximum rank for any space-time codeword c and space-time codeword e. Here, the matrix B (c, e) is a matrix representing the difference between the space-time codeword c and the space-time codeword e.

In addition, when mapping signals using the BPSK scheme, the space-time code design condition 2 must be satisfied.

[Space-Time Code Design Condition 2]

If all nonzero space-time codeword matrices have a maximum rank on the binary field, the space-time codeword has the maximum diversity gain.

In addition, when mapping signals using the QPSK scheme, the space-time code design condition 3 must be satisfied.

[Space-Time Code Design Condition 3]

Spatiotemporal codeword matrix mapped to nonzero space-time codeword matrix c mapped onto binary field

$\Phi(c)$

Has a maximum rank on the binary field, the space-time codeword has the maximum diversity gain.

As described above, the space-time code is designed to have a maximum rank to obtain a maximum diversity gain. That is, the space-time code is designed in such a manner that its generating matrix (G) has a maximum rank.

On the other hand, the next generation communication system has been developed to service high-speed large data having various quality of service (hereinafter referred to as 'QoS'). In this way, the loss of information on transmission is fatal in serving high-speed large data, and therefore the error correction capability of the error correction code acts as an important factor in determining the overall quality of service. Representative codes of the error correction code include a turbo code, a low density parity check (LDPC) code, and the like.

The LDPC code is known to have a superior performance gain in high-speed data transmission, compared to a convolutional code used mainly for error correction. The LDPC code effectively corrects an error due to noise generated in a wireless channel and transmits the data. It has the advantage of increasing the reliability of. In addition, the LDPC code can be decoded using an iterative decoding algorithm based on a sum-product algorithm on a factor (hereinafter, referred to as a 'factor') graph. By using a decoding method using an iterative decoding algorithm based on the sum product algorithm, the decoder of the LDPC code not only has a lower complexity than the decoder of the turbo code but also facilitates the implementation of a parallel processing decoder.

The turbo code has an excellent performance approaching the channel capacity limit of Shannon's channel coding theorem, and the LDPC code having the best performance so far is the block size.

10^{7}

Bit Error Rate (BER)

10^{-5}

Shows the performance of only 0.04 [dB] difference in channel capacity limit of Shannon's channel coding. Here, Shannon's channel coding theory is a theory indicating that reliable communication is possible only at a data rate that does not exceed the capacity of a channel. Random codes with very large block sizes show performance close to the channel capacity limit in Shannon's channel coding theory, but when applied to MAP (maximum a posteriori) or ML (maximum likelihood) decoding, There was a huge load, so the actual implementation was impossible.

However, in the LDPC code, most elements have a zero value, and very few elements other than the elements having the value of zero have a non-zero value. For example, it is defined by a parity check matrix having a value of 1. Hereinafter, for convenience of explanation, the non-zero value is assumed to be a value of 1 and will be described. Therefore, since the parity check matrix of the LDPC code has a small weight, it can be decoded through iterative decoding even in a block code having a relatively long length, and if the block length of the block code is continuously increased, As shown by the turbo code, performance in the form of approaching Shannon's channel capacity limit is shown. Here, the weight represents the number of elements having a non-zero value among the elements constituting the parity check matrix. Therefore, in the next generation communication system, the LDPC code is actively used as the error correction code.

However, in the case of the LDPC code, when the encoding is performed using the generation matrix as in the space-time code, excellent performance cannot be guaranteed. That is, in the case of the LDPC code, as described above, the weight of the parity check matrix is small and the complexity thereof is low when decoding. However, when the parity check matrix is converted into a generation matrix, the weight of the generation matrix is increased. Its complexity increases in coding.

As a result, when designing the space-time code using the LDPC code, it is difficult to apply design conditions in terms of a generation matrix of a general space-time code. Accordingly, there is a need for a method in which a space-time code using the LDPC code can obtain the maximum diversity gain, that is, a parity check matrix capable of obtaining the maximum diversity gain.

Accordingly, an object of the present invention is to provide an apparatus and method for encoding / decoding a space-time LDPC code for obtaining a maximum diversity gain in a mobile communication system.

Another object of the present invention is to provide an apparatus and method for encoding / decoding a space-time LDPC code having a correlation between a plurality of transmit antennas in a mobile communication system using a plurality of transmit antennas.

Another object of the present invention is to provide a method of designing a parity check matrix of a space-time LDPC code for obtaining a maximum diversity gain in a mobile communication system.

The apparatus of the present invention for achieving the above objects; In a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, an encoding apparatus of a space-time low density parity check (LDPC) code, wherein when information data is input, An LDPC encoder for encoding the information data according to a predetermined control to generate an LDPC code, a spatial mapper for spatially mapping the LDPC code according to a predetermined spatial mapping method, and generating the spatiotemporal LDPC code; A fifth partial matrix comprising a second partial matrix composed of even columns of the first partial matrix corresponding to data and a fourth partial matrix composed of odd columns of the third partial matrix corresponding to parity; A sixth partial matrix of odd-numbered columns of the first partial matrix and a seventh column of even-numbered columns of the third partial matrix The information data corresponds to a parity check matrix having a predetermined rank on a binary field and an ninth partial matrix exclusively combining the first partial matrix and the third partial matrix. And a controller for controlling to generate the LDPC code by encoding.

Another apparatus of the present invention for achieving the above objects; In the encoding apparatus of a space-time low density parity check (LDPC)

code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, when information data is input, A space-time LDPC encoder that encodes the information data according to a predetermined control to generate an LDPC code, and spatially maps the LDPC code into a space-time LDPC code corresponding to a preset spatial mapping method; The first partial matrix corresponding to the three subwords, the second partial matrix corresponding to the second and fourth subwords, and the third partial matrix exclusively ORing the first partial and second partial matrices on the binary field. A code for controlling to generate the LDPC code by encoding the information data corresponding to a parity check matrix having a predetermined rank; Wherein the space-time LDPC code is divided into a real part space-time LDPC code and an imaginary part space-time LDPC code, and the real part space-time LDPC code is transmitted through a first transmitting antenna, which is any one of the plurality of transmitting antennas. A first codeword to be transmitted and a second coder to be transmitted through any second transmitting antenna among the plurality of transmitting antennas, and the imaginary part space-time LDPC code is combined with a third coder transmitted through the first transmitting antenna. And a fourth codeword transmitted through the second transmitting antenna.

Another apparatus of the present invention for achieving the above objects is; A decoding apparatus of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, the plurality of receive antennas. A detector for detecting a received signal through the received signal, a spatial mapped node decoder for performing spatial mapped node decoding based on the detected values of the received signals, and a parity with which the spatial mapped node decoded signal is preset LDPC encoder that decodes corresponding to the parity check matrix and the probability values of the signal decoded corresponding to the parity check matrix satisfy the pre-set iterative decoding stop condition, hard-determine the probability values, and output the information data. Does not satisfy the condition of repetitive decoding interruption Otherwise, the controller controls to repeatedly perform the operations of the second process to the third process until the iterative decoding stop condition is satisfied.

The method of the present invention for achieving the above objects; A method of generating a parity check matrix of a space-time low density parity check (LDPC) code in a mobile communication system comprising a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas. Determining a size of a parity check matrix so as to correspond to a coding rate applied when data is encoded into a space-time LDPC code and a codeword length of the space-time LDPC code, and performing a parity check matrix having the determined size corresponding to the information data. Dividing into a partial partition matrix and a second partial matrix corresponding to parity, a third partial matrix including even columns of the first partial matrix, and odd columns of the second partial matrix A fourth partial matrix consisting of: a fifth partial matrix consisting of a third submatrix and a fourth partial matrix; and the first partial matrix A sixth partial matrix composed of odd-numbered columns, a seventh partial matrix composed of even-numbered columns of the second partial matrix, an eighth partial matrix comprising a combination of the sixth and seventh partial matrixes, Generating a ninth partial matrix exclusively combining the one partial matrix and the second partial matrix, and wherein the fifth partial matrix and the eighth partial matrix have a predetermined rank on the ninth partial matrix and the binary field. And generating the parity check matrix.

Another method of the present invention for achieving the above objects is; A method of generating a parity check matrix of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, Determining a size of a parity check matrix to correspond to a coding rate applied when encoding information data into a space-time LDPC code and a codeword length of the space-time LDPC code, and converting the space-time LDPC code into a real part space-time LDPC code and an imaginary part space-time. Splitting the real part space-time LDPC code into a first codeword transmitted through a first transmitting antenna and a second coder transmitting through an arbitrary second transmitting antenna, and dividing the imaginary part space-time LDPC code into the LDPC code. A third codeword transmitted through a first transmitting antenna and transmitted through the second transmitting antenna Dividing the parity check matrix into a first partial matrix corresponding to the first and third codewords, and a second partial matrix corresponding to the second and fourth codewords. And the parity check matrix such that the first partial matrix, the second partial matrix, and the third partial matrix exclusively ORing the first partial matrix and the second partial matrix have a

predetermined rank on a binary field. It characterized in that it comprises a process of generating.

Another method of the present invention for achieving the above objects is; A method of encoding space-time low density parity check (LDPC) codes in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, the method comprising: inputting information data And a fifth partial matrix including a second partial matrix composed of even columns of the first partial matrix corresponding to the information data and a fourth partial matrix composed of odd columns of the third partial matrix corresponding to parity. An eighth partial matrix, which is a combination of a partial matrix and a sixth partial matrix composed of odd-numbered columns of the first partial matrix, and a seventh partial matrix composed of even-numbered columns of the third partial matrix, the first partial matrix and the first partial matrix; The information corresponding to a parity check matrix having a predetermined rank on a binary field and a ninth partial matrix exclusively ORed to the third partial matrix; And encoding the data into an LDPC code, and spatially mapping the LDPC code into a space-time LDPC code according to a predetermined spatial mapping method.

Another method of the present invention for achieving the above objects is; A method of encoding space-time low density parity check (LDPC) codes in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, the method comprising: inputting information data And a first partial matrix corresponding to the first and third subwords, a second partial matrix corresponding to the second and fourth subwords, and an exclusive OR of the first partial matrix and the second partial matrix. Encoding the information data according to a parity check matrix having a predetermined rank on a binary field to generate an LDPC code, and spatially mapping the LDPC code according to a predetermined spatial mapping method. Generating a space-time LDPC code, wherein the space-time LDPC code includes a real part space-time LDPC code and an imaginary part space-time LDPC code. The real part space-time LDPC code is divided into a call and is transmitted through a first codeword transmitted through a first transmit antenna, which is an arbitrary transmit antenna among the plurality of transmit antennas, and an arbitrary second transmit antenna among the plurality of transmit antennas. The imaginary part space-time LDPC code is divided into a third

coder transmitted through the first transmit antenna and a fourth coder transmitted through the second transmit antenna.

Another method of the present invention for achieving the above objects is; A decoding method of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas, the plurality of receive antennas. A first process of detecting a received signal through signals received through the second process, a second process of performing spatial mapping node decoding based on the detected values of the received signals, and a predetermined process of the spatial mapping node decoded signal A third step of decoding corresponding to the parity check matrix; and if the probability values of the signal decoded corresponding to the parity check matrix satisfy a preset repetition decoding interrupt condition, the probabilities are hard-determined and output as information data, and the probability value is output. Repeat repeating if they do not satisfy the condition of Until it meets the stop condition screen characterized in that it comprises a fourth step of performing control so as to repeat the operation of the second process to third process.

Hereinafter, with reference to the accompanying drawings in accordance with the present invention will be described in detail. It should be noted that in the following description, only parts necessary for understanding the operation according to the present invention will be described, and descriptions of other parts will be omitted so as not to distract from the gist of the present invention.

The present invention proposes an encoding / decoding scheme for a space-time code, that is, a space-time LDPC code using a low density parity check (LDPC) code, which is an error-correcting code. Suggest. In particular, the present invention proposes an encoding / decoding scheme for a space-time LDPC code having a maximum diversity gain.

Then, the LDPC code will be described before explaining the present invention.

The LDPC code may be decoded using an iterative decoding algorithm based on a sum-product algorithm on a factor (hereinafter, referred to as a 'factor') graph. By using a decoding method using an iterative decoding algorithm based on the sum product algorithm, the decoder of the LDPC

code not only has a lower complexity than the turbo code decoder, but also is easy to implement a parallel processing decoder. Done. When the LDPC code is expressed as a factor graph, a cycle exists on the factor graph of the LDPC code. It is well known that iterative decoding on the factor graph of the LDPC code in which the cycle exists is suboptimal. It is a known fact. In addition, the LDPC code has been experimentally proved to have excellent performance through iterative decoding.

Also, the LDPC code is a code proposed by Gallager, and most elements have a non-zero value, for example, a value of 1, except for elements having a value of 0. A very small number of elements are defined by a parity check matrix having a non-zero value, for example, a value of 1. Hereinafter, for convenience of explanation, the non-zero value is assumed to be a value of 1 and will be described. Therefore, since the parity check matrix of the LDPC code has a small weight, it can be decoded through iterative decoding even in a block code having a relatively long length, and if the block length of the block code is continuously increased, As shown by the turbo code, performance in the form of approaching Shannon's channel capacity limit is shown. Here, the weight represents the number of elements having a non-zero value among the elements constituting the parity check matrix. For example, the (N, j, k) LDPC code is a linear block code having a block length of N, elements having j values of 1 for each column, and each row. A sparse structure consisting of elements having k values of 1 per (row), and all elements except 0 having values of 1 are elements having a value of 0. It is defined by the parity check matrix of.

Next, a parity check matrix of an (N, j, k) LDPC code, for example, an (8, 2, 4) LDPC code, will be described with reference to FIG. 3.

3 is a diagram illustrating a parity check matrix of a general (8, 2, 4) LDPC code.

Referring to FIG. 3, first, the parity check matrix H of the (8, 2, 4) LDPC code is composed of eight columns and four rows, and the weight of each column is equal to two, and the weight of each row is Uniform to 4 Thus, since the weight of each column and the weight of each row in the parity check matrix are uniform, the (8, 2, 4) LDPC code becomes a regular LDPC code. If the weight of each column and the weight of each row in the parity check matrix are not uniform, the LDPC code becomes an irregular LDPC code.

In FIG. 3, the parity check matrix of the (8, 2, 4) uniform LDPC code has been described. Next, the factor graph of the (8, 2, 4) uniform LDPC code described with reference to FIG. 3 will be described with reference to FIG. 4. Let's explain.

4 is a diagram illustrating a factor graph of the (8, 2, 4) LDPC code of FIG.

Referring to FIG. 4, the factor graph of the (8, 2, 4) LDPC code has eight variable nodes, that is, x_1 (400), x_2 (402), and x_3 (404).), X_4 406, x_5 408, x_6 410, x_7 412, x_8 414, and four check nodes 416, 418, 420, 422 It consists of. Variable node xi and jth when there is an element having a nonzero value, that is, an element having a value of 1, at the point where the i th column and the j th row of the parity check matrix of the (8, 2, 4) LDPC code intersect Branches are created between check nodes.

Since the parity check matrix of the LDPC code has a very small weight, it can be decoded by iterative decoding even in a block code having a relatively long length, and if the block length of the block code is continuously increased, as in the turbo code, It shows performance close to Shannon's channel capacity limit. In addition, MacKay and Neal have already demonstrated that an iterative decoding process of LDPC codes using a flow transfer scheme has a performance close to that of an iterative decoding process of turbo codes.

Accordingly, the present invention proposes a space-time LDPC encoding / decoding method that can obtain a maximum diversity gain when transmitting an LDPC code guaranteeing excellent performance through a plurality of Tx.ANTs.

Next, a transmitter structure of a mobile communication system performing a function in an embodiment of the present invention, in particular, a transmitter structure of a mobile communication system when the BPSK scheme is used as a signal mapping scheme will be described with reference to FIG. 5.

FIG. 5 is a diagram schematically illustrating a transmitter structure for performing a function in an embodiment of the present invention when using a BPSK scheme as a signal mapping scheme.

Referring to FIG. 5, first, the transmitter includes a serial to parallel converter 500, a memory 502, a controller 504, and a first LDPC encoder (LDPC).

encoder 506, a second LDPC encoder 508, a spatial mapper 510, and a plurality of, for example, two signal mappers, i.e., a first signal mapper. 512 to the second signal mapper 514 and a plurality of, for example, the two transmit antennas Tx.ANT, that is, the first transmit antenna Tx.ANT 1 and the second transmit antenna Tx. ANT 2). Here, it is assumed that each of the first signal mapper 512 and the second signal mapper 514 maps signals in a binary phase shift keying (BPSK) scheme.

First, the memory 502 stores a parity check matrix according to the design method of the space-time LDPC code proposed in the present invention. Since the parity check matrix according to the design method of the space-time LDPC code proposed by the present invention will be described in detail below, the detailed description thereof will be omitted. First, when information data s occurs, the information data s is transferred to the serial / parallel converter 500, and the serial / parallel converter 500 converts the information data s into first information data s_1. And the second information data s_2 are converted in parallel and output.

The first information data s_1 is input to the first LDPC encoder 506 and the spatial mapper 510, and the second information data s_2 is input to the second LDPC encoder 508 and the spatial mapper 510. Is entered. The first LDPC encoder 506 encodes the first information data $s_{1\ corresponding}$ to a preset coding rate according to the control of the controller 504, so as to perform an LDPC codeword, that is, LDPC. A coded symbol is generated and output to the spatial mapper 510. Here, the controller 504 controls the first LDPC encoder 506 to correspond to the parity check matrix stored in the memory 502 to generate an LDPC codeword corresponding to the parity check matrix. . For example, assuming that the first information data s_1 is k bits and the coding rate is 1/2, the number N of bits output from the first LDPC encoder 506 is 2k (N = 2k). Here, the LDPC codeword output from the first LDPC encoder 506 becomes a first parity p_1 which is a part of parity based on the entire information data s.

Also, the second LDPC encoder 508 encodes the second information data $s_{2\ according}$ to a preset coding rate under the control of the controller 504 to generate an LDPC codeword, that is, an LDPC coded symbol. The output is then output to the space mapper 510. Here, the controller 504 controls the second LDPC encoder 506 to correspond to the parity check matrix stored in the memory 502 to generate an LDPC codeword corresponding to the parity check matrix. For example, assuming that the second information data

s_2 is k bits and the coding rate is 1/2, the number N of bits output from the second LDPC encoder 506 is 2k (N = 2k). Here, the LDPC codeword output from the second LDPC encoder 508 becomes a second parity p_2 which is a part of parity based on the entire information data s.

Meanwhile, the parity check matrix stored in the memory 502 is expressed as H, and the parity check matrix H is an information part H_s which is a part corresponding to information data and a part corresponding to parity. In parity part (parity part) is divided into $H_{p\,lt}$ can be expressed as Equation 2 below.

$$H=[H_s \;\; H_p]$$

In addition, the size of the parity check matrix H is determined to be a size corresponding to the code rate of the transmitter and the length of the space-time LDPC code which is the final codeword. Meanwhile, a 2k-bit LDPC codeword output from the first LDPC encoder 506 will be referred to as a 'first LDPC codeword c_1 ', and a 2k-bit LDPC output from the second LDPC encoder 506 will be described. The codeword will be referred to as a 'second LDPC codeword c_2 '. In addition, the first LDPC codeword c_1 is composed of k-bit information codeword s_1 and parity codeword p_1 , and the second LDPC codeword c_2 is k-bit information codeword s_2 and parity codeword $p._2$, and thus, the first LDPC codeword c_1 and the second LDPC codeword c_2 may be represented by Equation 3 below.

$$\mathbf{c}_1 = [\mathbf{s}_1 \;\; \mathbf{p}_1]$$

$$\mathbf{c}_2 = [\mathbf{s}_2 \;\; \mathbf{p}_{2]}$$

Here, the first LDPC codeword c_1 and the second LDPC codeword c_2 have a relationship as shown in Equation 4 below with the parity check matrix H.

$$H \cdot c_1 = [H_s \ H_p] \cdot [s_1 \ p_1]^T = 0$$

$$H \cdot c_2 = [H_s \ H_p] \cdot [s_2 \ p_2]^T = 0$$

In addition, the spatial mapper 510 inputs a first LDPC codeword c 1 output from the first LDPC coder 506 and a second LDPC codeword c 2 output from the second LDPC coder 508. Spatially map to a space-time LPDC codeword C by using a preset spatial mapping method, and then output to each of the corresponding signal mappers to be transmitted through the two transmitting antennas, that is, the first transmitting antenna and the second transmitting antenna. To control. There are various ways in which the spatial mapper 510 spatially maps the first LDPC codeword c 1 and the second LDPC codeword c 2. In the present invention, the spatial mapper 510 performs a spatial mapping based on a method as described below. Let's assume it's mapping.

The spatial mapper 510 is an even-numbered information bit constituting the information codeword s 1 of the first LDPC codeword c 1 output from the first LDPC coder 506.

$$s_1^e$$

And even-numbered information bits constituting the information codeword s 2 of the second LDPC codeword c 2 output from the second LDPC coder 508.

$$s_2^e$$

And odd parity bits constituting the parity codeword p 1 of the first LDPC codeword c 1 output from the first LDPC encoder 506.

$$p_1^o$$

And odd parity bits constituting the parity codeword p 2 of the second LDPC codeword c 2 output from the second LDPC coder 508.

$$p_2^o$$

Spatial mapping to transmit through the first transmit antenna. Further, the spatial mapper 510 is an even parity bit constituting a parity codeword p_2 of the second LDPC codeword c_2 output from the second LDPC coder 508.

$$p_2^e$$

And even parity bits constituting the parity codeword p_1 of the first LDPC codeword c_1 output from the first LDPC encoder 506.

$$p_1^e$$

And odd information bits constituting the information codeword s_2 of the second LDPC codeword c_2 output from the second LDPC encoder 508.

$$s_2^o$$

And odd information bits constituting the information codeword s_1 of the first LDPC codeword c_1 output from the first LDPC coder 506.

$$s_1^o$$

Spatial mapping to transmit via the second transmit antenna.

When spatial mapping the first LDPC codeword c_1 and the second LDPC codeword c_2 in the same manner as described above, the output of the spatial mapper 510 is

$$2 \times k$$

A space-time LDPC codeword C, which is a matrix, may be expressed as in Equation 5 below.

$$C = \begin{bmatrix} s_1^e & s_2^e & p_1^o & p_2^o \\ \\ p_2^e & p_1^e & s_2^o & s_1^o \end{bmatrix}$$

However, since each of the first signal mapper 512 and the second signal mapper 514 signal-maps a signal input in a BPSK scheme, the space-time LDPC codeword C is a binary rank design rule. Must satisfy That is, as described in the prior art, when all non-zero space-time codewords have a maximum rank on a binary field, the space-time codewords have a maximum diversity gain. Therefore, in order for the space-time LDPC codeword C to

have the maximum diversity gain, it must have a maximum rank on the binary field, and therefore, the parity check matrix H must be designed to satisfy parity check matrix design condition 1 below.

[Parity Check Matrix Design Condition 1]

$$H_1=[H_s^e\ H_p^o], H_2=[H_s^o\ H_p^e]$$

Wow

$$H_s \oplus H_p$$

Must have a maximum rank on the binary field.

The matrix under the parity check matrix design condition 1

$$H_1=[H_s^e\ H_p^o]$$

Is the matrix

$$H_s$$

Matrix of even-numbered columns in

$$H_s^e$$

And the matrix

$$H_p$$

Matrix of odd columns in

$$H_p^o$$

Is a matrix combining

$$H_2=[H_s^o\ H_p^e]$$

Is the matrix

$$H_s$$

Matrix of odd columns in

$$H_s^o$$

And the matrix

H_p

Matrix of even-numbered columns in

H_p^e

Is a matrix that combines. In this case, when the parity check matrix H is designed to satisfy the parity check matrix condition 1, signals transmitted through each of the different transmit antennas of the transmitter have linear independent characteristics, and thus, the transmit antenna Even if a signal transmitted through any one of the transmission antennas suffers a serious fading phenomenon, it is possible to overcome the error caused by the fading phenomenon at the receiver side.

In this case, the encoding process of the space-time LDPC code described with reference to FIG. 5 will be described below.

First, a parity check matrix H designed to satisfy the parity check matrix design condition 1 is stored in the memory 502, and the controller 504 corresponds to the parity check matrix H stored in the memory 502. Each of the first LDPC encoder 506 and the second LDPC encoder 508 controls to encode the input information data. The spatial mapper 510 generates LDPC codewords c, that is, a first LDPC codeword, generated by the first LDPC coder 506 and the second LDPC coder 508 under the control of the controller 504. The first signal mapper 512 and the second signal are spatially mapped to c_1 and the second LDPC codeword c_2 in the order output by the first LDPC encoder 506 and the second LDPC encoder 508. Output to each of the mapper 514.

Each of the first signal mapper 512 and the second signal mapper 514 maps the signal output from the spatial mapper 510 in the BPSK method, and the first signal mapper 512 The signal mapped signal is transmitted on the air through the first transmission antenna, and the second signal mapper 514 transmits the signal mapped signal on the air through the second transmission antenna. .

5 illustrates a structure of a transmitter that performs a function in an embodiment of the present invention when a BPSK scheme is used as a signal mapping scheme. Next, a quadrature phase shift keying (QPSK) scheme is used as a signal mapping scheme with reference to FIG. 6. And a transmitter structure that performs a function in an embodiment of the

present invention when using a 4-QAM (Quadrature Amplitude Modulation) method.

6 is a diagram schematically illustrating a structure of a transmitter for performing a function in an embodiment of the present invention when using a QPSK scheme and a 4-QAM scheme as a signal mapping scheme.

Referring to FIG. 6, first, the transmitter includes a serial / parallel converter 600 and a memory 602, a controller 604, a real part space time LDPC encoder 606, and an imaginary number. Part time-space LDPC coder 608, a plurality of, for example, two signal mappers, namely, a first signal mapper 610 and a second signal mapper 612, and a plurality, for example the two transmissions Antennas, that is, a first transmitting antenna and a second transmitting antenna. Here, it is assumed that each of the first signal mapper 610 and the second signal mapper 612 maps a signal using any one of a QPSK scheme and a 4-QAM scheme.

First, the memory 602 stores a parity check matrix according to the design method of the space-time LDPC code proposed in the present invention. Since the parity check matrix according to the design method of the space-time LDPC code proposed by the present invention will be described in detail below, the detailed description thereof will be omitted. First, when the information data s is input, the serial-to-parallel converter 600 converts the information data s into real part information data s_I and imaginary part information data s_Q in parallel and outputs the same.

The real part information data s_I is input to the real part space time LDPC encoder 606, and the real part space time LDPC coder 606 presets the real part information data s_I under the control of the controller 604. The first signal mapper 610 and the second signal map are generated by generating an LDPC codeword, that is, an LDPC coded symbol, and performing real-time part-space LDPC encoding. Output to the instrument 612. Herein, the controller 604 controls the real part space-time LDPC coder 606 to generate a space-time LDPC codeword corresponding to the parity check matrix according to the parity check matrix stored in the memory 602. do. For example, assuming that the real part information data s_I is k bits and the coding rate is 1/2, the number N of bits output from the real part space-time LDPC encoder 606 is 2k (N = 2k). As a result, the output data of the real part space-time LDPC encoder 606 becomes a space-time LDPC codeword having a number of 2k bits. Here, the space-time LDPC codeword output

from the real part space-time LDPC coder 606 is expressed as C_I, and the space-time LDPC code word C_I is transmitted through the first transmitting antenna.

$$C_I^1$$

And transmitted through the second transmitting antenna

$$C_I^2$$

Separated by.

Further, according to the control of the imaginary part of information data s_Q is the imaginary part space LDPC is input to the encoder 608, the imaginary part space LDPC encoder 608 is the imaginary part of information data s_Q the controller 604 The first spatial mapper 610 and the second spatial mapper 612 are generated by LDPC codewords, that is, LDPC coded symbols, after performing a fictitious part space-time LDPC encoding. Will output Here, the controller 604 controls the imaginary part space-time LDPC encoder 608 to correspond to the parity check matrix stored in the memory 602 to generate a space-time LDPC codeword corresponding to the parity check matrix. do. For example, assuming that the imaginary part information data s_Q is k bits and the coding rate is 1/2, the number N of bits output from the imaginary part space-time LDPC encoder 608 is 2k (N = 2k). As a result, the output data of the imaginary part space-time LDPC encoder 608 becomes a space-time LDPC codeword having a number of 2k bits. Here, the space-time LDPC codeword output from the imaginary part space-time LDPC coder 608 is expressed as C_Q, and the space-time LDPC code word C_Q is transmitted through a first transmitting antenna.

$$C_Q^1$$

Transmitted via the second and second transmit antennas

$$C_Q^2$$

Separated by.

On the other hand, if a parity check matrix stored in the memory 602 is expressed as H, the parity check matrix H is a part corresponding to a codeword to be transmitted via the first transmission antenna to configure

the space-time LDPC code H_1 And H_2, which is a part corresponding to the codeword transmitted through the second transmitting antenna, may be expressed as Equation 6 below.

$$H=[H_1 \ H_2]$$

In addition, the size of the parity check matrix H is determined to be a size corresponding to the code rate of the transmitter and the length of the space-time LDPC code which is the final codeword. Meanwhile, the space-time LDPC codeword C_I generated by the real part space-time LDPC coder 606 and the space-time LDPC codeword C_Q generated by the imaginary part space-time LDPC coder 608 may be expressed by Equation 7 below.

$$C_I = \begin{bmatrix} \mathbf{c}_I^1 \\ \\ \mathbf{c}_I^2 \end{bmatrix}, \quad C_Q = \begin{bmatrix} \mathbf{c}_Q^1 \\ \\ \mathbf{c}_Q^2 \end{bmatrix}$$

Here, the space-time LDPC codeword C_I and the space-time LDPC codeword C_Q have a relationship as shown in Equation 8 below with the parity check matrix H.

$$H \cdot \mathbf{c}_I = [H_1 \ H_2] \cdot [\mathbf{c}_I^1 \ \mathbf{c}_I^2]^T = 0$$

$$H \cdot \mathbf{c}_Q = [H_1 \ H_2] \cdot [\mathbf{c}_Q^1 \mathbf{c}_Q^2]^T = 0$$

Meanwhile, the first signal mapper 610 outputs the real part space-time LDPC coder 606.

$$\mathbf{C}_{\mathbf{I}}^{\mathbf{1}}$$

And output from the imaginary part space-time LDPC encoder 608.

$$\mathbf{C}_{\mathbf{Q}}^{\mathbf{1}}$$

After inputting signal mapping as shown in Equation (9), it is transmitted through the first transmitting antenna.

$$(\frac{-1}{\sqrt{(2)}})^{C_I^1}+j(\frac{-1}{\sqrt{(2)}})^{C_Q^1}$$

Also, the second signal mapper 612 outputs the real part space-time LDPC coder 606.

$$C_I^2$$

And output from the imaginary part space-time LDPC encoder 608.

$$C_Q^2$$

After inputting the signal mapping as shown in Equation 10 below, and transmits through a second transmitting antenna.

$$(\frac{-1}{\sqrt{(2)}})^{C_I^2}+j(\frac{-1}{\sqrt{(2)}})^{C_Q^2}$$

in this case,

$$2\times k$$

The space-time LDPC codeword C, which is a matrix, may be represented by Equation 11 below.

$$C=\begin{bmatrix} c_I^1+jc_Q^1 \\ \\ c_I^2+jc_Q^2 \end{bmatrix}$$

However, since each of the first signal mapper 610 and the second signal mapper 612 performs signal mapping in one of the QPSK scheme and the 4-QAM scheme, the space-time LDPC codeword C is a QPSK binary. Rank design rules must be met. That is, when all non-zero space-time codewords are mapped on a binary field as described in the prior art, when the mapped space-time codeword has the maximum rank on the binary field, the space-

time codeword has the maximum diversity gain. Accordingly, in order for the space-time LDPC code C to have the maximum diversity gain, it must have a maximum rank on the binary field, and thus the parity check matrix H must satisfy parity check matrix design condition 2 below.

[Parity Check Matrix Design Condition 2]

The real part space-time LDPC code C_I and the imaginary part space-time LDPC code C_Q obtain a maximum diversity gain in a binary field.

$$H_1$$

,

$$H_2, H_1 \oplus H_2$$

Must have a maximum rank on the binary field.

The matrix under the parity check matrix design condition 2

$$H_1$$

Is the matrix

$$H$$

Denotes a parity matrix corresponding to a codeword transmitted through a first transmit antenna in

$$H_2$$

Is the matrix

$$H$$

Denotes a parity matrix corresponding to a codeword transmitted through a second transmit antenna.

In this case, the encoding process of the space-time LDPC code described with reference to FIG. 6 is summarized as follows.

First, a parity check matrix H designed to satisfy the parity check matrix design condition 2 is stored in the memory 602, and the controller 604 corresponds to the parity check matrix H stored in the memory 602. The real part space time LDPC coder 606 and the imaginary part space time LDPC

coder 608 control to encode the input information data. According to the control of the controller 604, the space-time LDPC codeword C_I and the space-time LDPC codeword C_Q generated by each of the real part space-time LDPC coder 606 and the imaginary part space-time LDPC coder 608 are the first signal mapping. Is input to each of the signal 610 and the second signal mapper 612.

Each of the first signal mapper 610 and the second signal mapper 612 is a space-time LDPC codeword C_I and a space-time LDPC codeword in the real part space-time LDPC coder 606 and the imaginary part space-time LDPC coder 608. C_Q is signal mapped in either of the above QPSK and 4-QAM schemes. The first signal mapper 610 transmits the signal mapped signal to the air through the first transmission antenna, and the second signal mapper 612 sends the signal mapped signal to the second signal. Transmit over the air via the transmit antenna.

6 illustrates a structure of a transmitter performing a function in an embodiment of the present invention when using a QPSK scheme and a 4-QAM scheme as a signal mapping scheme, and then, with reference to FIG. The operation of the receiver of the mobile communication system will be described.

7 is a flowchart illustrating an operation of a receiver of a mobile communication system according to an embodiment of the present invention.

Referring to FIG. 7, first, in step 700, the receiver includes a plurality of receivers provided in the receiver, for example, in a transmitter corresponding to the receiver through P receiving antennas, that is, a first receiving antenna to a P receiving antenna. After receiving the transmitted signal, the flow proceeds to step 702. In step 702, the receiver detects the received signal based on the signals received through each of the first to P-th reception antennas, and then proceeds to step 704. Here, of course, there may be various ways of detecting the received signal, and it is assumed in FIG. 7 that the received signal is detected in the manner described below.

In step 704, the receiver performs spatial mapping node decoding based on the detected value of the received signal, and proceeds to steps 706 and 708. Here, the spatial mapping node decryption is a decoding output value for a message and a first LDPC codeword transmitted from a receiving node to a spatial mapping node, that is, decoding of a first LDPC decoder (not

shown) for decoding the first LDPC codeword. The first LDPC decoder at the spatial mapping node based on an output value and a decoding output value of the second LDPC codeword, that is, a decoding output value of a second LDPC decoder (not shown) for decoding the second LDPC codeword; Detecting a message delivered to the second LDPC decoder. In step 706, the receiver performs a decoding process on the first LDPC codeword, that is, a first LDPC decoding process. In step 708, the receiver performs a decoding process on a second LDPC codeword, that is, a second LDPC decoding process. Proceed to step 710.

In step 710, the receiver determines whether a decoding stop condition is satisfied based on a result of performing the first LDPC decoding process and the second LDPC decoding process. In this case, the decoding stop condition may be a predetermined number of iterative decoding or a decoding result to date satisfies the parity check matrix H. That is, the receiver stops the decoding operation when the number of decoding exceeds the number of iterative decoding or when the decoding result up to the present satisfies the parity check matrix H.

If the decoding condition is satisfied in step 710, the receiver proceeds to step 712. In step 712, the receiver calculates a decoding value based on the decoding result thus far, that is, the result value of the first LDPC decoding process and the result value of the second LDPC decoding process, and proceeds to step 714. In step 714, the receiver hard-decisions the information data based on the calculated decoding value and then ends all decoding processes up to now.

If the check result in step 710 does not satisfy the decoding condition, the receiver returns to step 704. That is, the decoding performance is improved by performing the iterative decoding on the basis of the initially received received signal.

In FIG. 7, a receiver operation of a mobile communication system according to an embodiment of the present invention has been described. Next, a receiver structure of a mobile communication system performing a function in an embodiment of the present invention will be described with reference to FIG. 8. Shall be.

8 is a diagram schematically illustrating a receiver structure of a mobile communication system that performs a function in an embodiment of the present invention.

Referring to FIG. 8, the receiver first includes a plurality of receiver antennas, that is, two receiving antennas, that is, a first receiving antenna and a second receiving antenna, a detector 800, and a spatial mapping node decoder. node decoder 802, controller 804, memory 806, first LDPC decoder 808, second LDPC decoder 810, hard decision unit 812. It consists of.

First, each of the first and second receiving antennas receives a signal transmitted from a transmitter corresponding to the receiver and outputs the signal to the detector 800. The detector 800 inputs a received signal output from each of the first and second receiving antennas, and detects a transmission signal transmitted from each of the transmitting antennas of the transmitter, that is, the first transmitting antenna and the second transmitting antenna. After that, it outputs to the spatial node decoder 802. The spatial node decoder 802 inputs a signal output from the detector 800 to detect an estimate detected by the detector 800 for a signal transmitted from each of the first and second transmitting antennas. And output to the first LDPC decoder 808 and the second LDPC decoder 810. In this case, the signal input to the spatial mapping node decoder 802 is initially only the signal output from the detector 800 as described above, but in the subsequent iterative decoding, the signal is output from the detector 800. A signal output from the first LDPC decoder 808 and the second LDPC decoder 810 is included as well as a signal.

The first LDPC decoder 808 inputs a signal output from the spatial mapping node decoder 802 to correspond to a first LDPC encoder based on a parity check matrix designed by the space-time LDPC encoder of the transmitter. Perform decryption. Here, when the transmitter uses the QPSK scheme as described with reference to FIG. 6 as the signal mapping scheme, the first LDPC decoder 808 performs a decoding operation to correspond to the encoding process of the real part space-time LDPC encoder 606. do. That is, the controller 804 controls the first LDPC decoder 808 to perform a decoding process corresponding to the parity check matrix stored in the memory 806 so as to obtain the maximum diversity. It is.

The second LDPC decoder 810 inputs a signal output from the spatial mapping node decoder 802 to a second LDPC encoder based on a parity

check matrix designed by the space-time LDPC encoder of the transmitter. LDPC decoding is performed correspondingly. Here, when the transmitter uses the QPSK scheme as described with reference to FIG. 6 as the signal mapping scheme, the first LDPC decoder 810 performs a decoding operation so as to correspond to an encoding process of the imaginary part space-time LDPC encoder 608. do. That is, the controller 804 controls the first LDPC decoder 810 to perform a decoding process corresponding to the parity check matrix stored in the memory 806 so as to obtain the maximum diversity. It is.

The operation of the detector 800 and the spatial mapping node decoder 802 will now be described in detail.

First, a transmission signal transmitted at an arbitrary time point t is expressed as a transmission signal vector x_t composed of two elements, and the received signal received at the arbitrary time point t is a received signal consisting of one element. It is expressed as a vector y_t. Here, the transmission signal vector x_t is composed of two elements because it is transmitted through two transmission antennas, that is, two transmission antennas of the first transmission antenna and the second transmission antenna. In addition, the channel experienced when the signal is transmitted and received is a fading channel (fading channel),

1×2

Phosphorus

Ω_t

The noise component is expressed as a noise vector n_t composed of one element. Then, the relationship as shown in Equation 12 is established.

$$y_t = \Omega_t x_t + n_t$$

In Equation 12, the transmission signal vector x_t considers a transmission signal transmitted through each of the first transmission antenna and the second transmission antenna at an arbitrary time point t.

$$x_t = (x_t^1, x_t^2)$$

It can be expressed as Where

x_t^1

Is a signal transmitted through the first transmitting antenna at the arbitrary time t, and

x_t^2

Is a signal transmitted through the second transmitting antenna at the arbitrary time t. Therefore, the transmission signal vector

$$x_t=(x_t^1,x_t^2)$$

Probability vector P_t with respect to the received signal vector y_t can be expressed by Equation 13 below. Here, for convenience of description, it is assumed that the signal mapping method applied by the transmitter is a BPSK method, and the QPSK method as well as the BPSK method may be used as the signal mapping method.

$$P_t=(P_t(x_t^1=1,x_t^2=1),P_t(x_t^1=-1,x_t^2=1),P_t(x_t^1=1,x_t^2=-1),P_t(x_t^1=-1,x_t^2=-1))$$

As shown in Equation 13, the transmission signal vector

$$x_t=(x_t^1,x_t^2)$$

For, the probability vector P_t for the received signal vector y_t is composed of four elements. Here, will be the transmitter hayeoteumeuro apply the BPSK scheme in the signal mapping method, consists of the probability vector P_t are four elements, as a result, the message is the received signal vector y_t is detected by the detector (800) To be.

Further, to obtain each of the four elements of the probability vector P_t

$$x_t=(x_t^1,\ x_t^2)=(i,j),(i,j\in -1,1)$$

In the case of Equation 14,

$$P_t(x_t^1=i,x_t^2=j)=Pr(x_t^1=i,x_t^2=j|y_t)$$

In Equation 14, Pr is an operator for obtaining a probability. As described above, the detector 800 detects the probability vector P_t and outputs the detected probability vector P_t to the spatial mapping node decoder 802.

Meanwhile, a log likelihood ratio (LLR) message transmitted from the first transmitting antenna, which is output from the spatial mapping node decoder 802 to the first LDPC decoder 808, may be expressed by Equation 15 below.

$$L_t^{1=}\log\frac{\Pr(x_t^1=+1|y_t)}{\Pr(x_j^t=-1|y_t)} = \log\frac{\sum_{x_t^2\in\{-1,1\}}\Pr(x_t^1=+1,x_t^2|y_t)\Pr(x_t^2|y_t)}{\sum_{x_t^2\in\{-1,1\}}\Pr(x_j^t=-1,x_t^2|y_t)\Pr(x_t^2|y_t)}$$

In Equation 15

$$\Pr(x_k^2|y_k)$$

May be detected from output values of the first LDPC decoder 808 and the second LDPC decoder 810. In addition, the LLR message transmitted from the second transmit antenna may be detected in the same manner as the method of detecting the LLR message transmitted from the first transmit antenna. The LLR message transmitted from the second transmit antenna is input to the second LDPC decoder 810.

In the above description, operations of the detector 800 and the spatial mapping node decoder 802 when the transmitter applies the BPSK scheme as the signal mapping scheme are described. In the following, the transmitter applies the QPSK scheme as the signal mapping scheme. The operation of the detector 800 and the spatial mapping node decoder 802 will now be described.

First, the transmission signal vector transmitted by the transmitter

$$x_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)$$

The probability vector P_t for the received signal vector y_t received by the receiver can be expressed by Equation 16 below.

$$P_t=\left(P_t((x_t^1)^I=1,(x_t^1)^Q=1,(x_t^2)^I=1,(x_t^2)^Q=1),\cdots,\right.$$
$$\left.P_t((x_t^1)^I=-1,(x_t^1)^Q=-1,(x_t^2)^I=-1,(x_t^2)^Q=-1)\right)$$

The transmission signal vector as shown in Equation 16

$$x_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)$$

For, the probability vector P_t for the received signal vector y_t is composed of 16 elements. Here, since the transmitter applies the QPSK scheme as the signal mapping scheme, the probability vector P_t is composed of 16 elements, and as a result, the received signal vector y_t is a message detected from the detector 800. To be.

Here, to obtain each of the sixteen elements of the probability vector P_t

$$x_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)=(i,j,k,l),(i,j,k,l\in -1,1)$$

In the case of, Equation 17 may be represented.

$$P_t((x_t^1)^I=i,(x_t^1)^Q=j,(x_t^2)^I=k,(x_t^2)^Q=l)$$
$$=\Pr((x_t^1)^I=i,(x_t^1)^Q=j,(x_t^2)^I=k,(x_t^2)^Q=l\,|\,y_t)$$

As described above, the detector 800 detects the probability vector P_t and outputs the detected probability vector P_t to the spatial mapping node decoder 802. The LLR message corresponding to the real part transmitted from the first transmit antenna, which is output from the spatial mapping node decoder 802 to the first LDPC decoder 808, may be represented by Equation 18 below.

$$(L_t^1)^I=\log\frac{\Pr((x_t^1)^I=+1|y_t)}{\Pr((x_t^1)^I=-1|y_t)}$$
$$-\log\frac{\sum_{(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q\in\{-1,1\}}\Pr((x_t^1)^I=+1,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)}{\sum_{(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q\in\{-1,1\}}\Pr((x_t^1)^I=-1,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)}$$

In Equation 18

$$\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_k)$$

May be detected from output values of the first LDPC decoder 808 and the second LDPC decoder 810. In addition, the LLR message corresponding to the imaginary part transmitted from the first transmitting antenna and the second transmitting antenna are transmitted in the same manner as the method of detecting the LLR message corresponding to the real part transmitted from the first transmitting antenna. The LLR message corresponding to the real part may be detected and the imaginary part LLR message transmitted from the second transmitting antenna. Here, the LLR message transmitted from the second transmit antenna is input to the second LDPC decoder 810.

The first LDPC decoder 808 and the second LDPC decoder 810 decode the first LDPC codeword and the second LDPC codeword based on the output values of the spatial mapping node decoder 802. The decoded signals are input to the spatial mapping node decoder 802 and the hard determiner 812. By controlling the output values of the first LDPC decoder 808 and the second LDPC decoder 810 to be input back to the spatial mapping node decoder 802, the reliability of the initially estimated signal may be increased. will be. The hard determiner 812 inputs the signals output from the first LDPC decoder 808 and the second LDPC decoder 810 to perform hard decision to restore the final information data.

Meanwhile, in the detailed description of the present invention, specific embodiments have been described, but various modifications are possible without departing from the scope of the present invention. Therefore, the scope of the present invention should not be limited to the described embodiments, but should be defined not only by the scope of the following claims, but also by the equivalents of the claims.

The present invention as described above has the advantage of maximizing the system performance by maximizing error correction capability by newly proposing a parity check matrix capable of encoding / decoding a space-time LDPC code having the maximum diversity gain. In addition, by using the LDPC method as an encoding method of the space-time code, there is an advantage that the space-time code that guarantees excellent performance with a relatively simple hardware structure can be encoded / decoded.

Claims (43)

Hide Dependent

1. A method of generating a parity check matrix of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

   Determining a size of a parity check matrix to correspond to a coding rate applied when encoding information data into a space-time LDPC code and a codeword length of the space-time LDPC code;

   Dividing the parity check matrix of the determined size into a first partial matrix corresponding to the information data and a second partial matrix corresponding to parity;

   A third partial matrix composed of even columns of the first partial matrix, a fourth partial matrix composed of odd columns of the second partial matrix, and the third partial matrix and the fourth partial matrix A fifth partial matrix, which is a matrix of?, A sixth partial matrix composed of odd-numbered columns of the first partial matrix, a seventh partial matrix composed of even-numbered columns of the second partial matrix, and the sixth partial matrix; Generating an eighth partial matrix that is a matrix of seventh partial matrixes, a ninth partial matrix of an exclusive OR of the first partial matrix and the second partial matrix,

   Generating the parity check matrix such that the fifth partial matrix and the eighth partial matrix have a predetermined rank on the ninth partial matrix and a binary field.

2. The method of claim 1,

   The set rank is characterized in that the full rank (full rank).

3. The method of claim 1,

   The number of transmit antennas is two, and the space-time LDPC code is signal mapped using a Binary Phase Shift Keying (BPSK) scheme.

4. A method for generating a parity check matrix of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

   Determining a size of a parity check matrix to correspond to a coding rate applied when encoding information data into a space-time LDPC code and a codeword length of the space-time LDPC code;

   The space time LDPC code is divided into a real part space time LDPC code and an imaginary part space time LDPC code, and the real part space time LDPC code is transmitted through a first coder transmitted through a first transmitting antenna and an arbitrary second transmitting antenna. Dividing the imaginary part space-time LDPC code into a third coder transmitted through the first transmitting antenna and a fourth coder transmitted through the second transmitting antenna;

   Dividing the parity check matrix into a first partial matrix corresponding to the first and third codewords and a second partial matrix corresponding to the second and fourth codewords;

   Generating the parity check matrix such that the first partial matrix, the second partial matrix, and a third partial matrix obtained by exclusively ORing the first partial matrix and the second partial matrix have a predetermined rank on a binary field. The method characterized in that it comprises a.

5. The method of claim 4, wherein

   And generating the parity check matrix such that the real part space time LDPC code and the imaginary part space time LDPC code have a predetermined rank on a binary field.

6. The method of claim 4, wherein

   The set rank is characterized in that the full rank (full rank).

7. The method of claim 6,

The real part space time LDPC code is a codeword corresponding to real part information data of the information data, and the imaginary part space time LDPC code is a code word corresponding to an imaginary part information data of the information data.

8. The method of claim 7, wherein

The number of transmit antennas is two, and the space-time LDPC code is signal mapped using any one of a Quadrature Phase Shift Keying (QPSK) scheme and a Quadrature Amplitude Modulation (4-QAM) scheme.

9. A method of encoding a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

Entering information data,

A fifth part which is a matrix combining a second part matrix composed of even columns of the first part matrix corresponding to the information data and a fourth part matrix composed of odd columns of the third partial matrix corresponding to parity; An eighth partial matrix, which is a matrix combining a sixth partial matrix composed of an odd-numbered column of the matrix and the first partial matrix, and a seventh partial matrix composed of even-numbered columns of the third partial matrix, is formed of the first partial matrix and the first partial matrix. Generating the LDPC code by encoding the information data corresponding to a parity check matrix having a predetermined rank on a binary field and a ninth partial matrix exclusively OR of a three-part matrix;

And mapping the LDPC code into a space-time LDPC code by performing a spatial mapping corresponding to a preset spatial mapping scheme.

10. The method of claim 9,

The set rank is characterized in that the full rank (full rank).

11. The method of claim 10,

The number of the transmitting antennas is two, and further comprising the step of mapping the space-time LDPC code using a Binary Phase Shift Keying (BPSK) scheme.

12. The method of claim 11,

Generating the information data into the space-time LDPC code;

Converting the information data into first information data and second information data in parallel;

Generating a first codeword that is a first parity of the information data by encoding the first information data corresponding to the parity check matrix;

Generating a second codeword which is a second parity of the information data by encoding the second information data corresponding to the parity check matrix;

The even bits forming the first information data, the even bits forming the second information data, the odd bits forming the first codeword, and the second codeword The odd bits are transmitted through one of the two transmit antennas, the even bits forming the second codeword, the even bits forming the first codeword, and the first bit. Generating spatial information by spatial mapping the odd-numbered bits constituting the second information data and the odd-numbered bits constituting the first information data to be transmitted through one of the transmission antennas. The method characterized in that.

13. A method of encoding a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

Entering information data,

A first partial matrix corresponding to the first and third subwords, a second partial matrix corresponding to the second and fourth subwords, and a third partial exclusively OR of the first partial matrix and the second partial matrix Generating a LDPC code by encoding the information data corresponding to a parity check matrix having a predetermined rank on a binary field;

Generating a space-time LDPC code by spatial mapping the LDPC code according to a predetermined spatial mapping method;

The space-time LDPC code is divided into a real part space-time LDPC code and an imaginary part space-time LDPC code, and the real part space-time LDPC code is a first code transmitted through a first transmit antenna, which is an arbitrary transmit antenna among the plurality of transmit antennas. And a second coder transmitted through any second transmit antenna among the plurality of transmit antennas, and the imaginary part space-time LDPC code is divided into a third coder transmitted through the first transmit antenna and the second transmit antenna. The method as claimed in claim 4, wherein the method is partitioned into fourth codewords transmitted through the control unit.

14. The method of claim 13,

The method of claim 4, further comprising the step of signal-thinning the space-time LDPC code using any one of a Quadrature Phase Shift Keying (QPSK) scheme and a Quadrature Amplitude Modulation (4-QAM) scheme.

15. The method of claim 14,

Generating the information data into the space-time LDPC code;

Converting the information data into real part information data and imaginary part information data in parallel;

Generating a first codeword by encoding the real part information data corresponding to the parity check matrix;

Generating a second codeword by encoding the imaginary part information data corresponding to the parity check matrix;

Pre-set bits of the bits constituting the first codeword and preset bits of the bits constituting the second coder are transmitted through the first transmitting antenna, and bits constituting the first coder. Generating a space-time LDPC code by spatial mapping the remaining bits except for the set bits among the set bits and the bits constituting the second codeword to be transmitted through the second transmitting antenna. The method characterized in that it comprises a.

16. The method of claim 15,

Wherein the parity check matrix is generated such that the real part space time LDPC code and the imaginary part space time LDPC code have a predetermined rank on a binary field.

17. The method of claim 16,

The set rank is characterized in that the full rank (full rank).

18. An apparatus for encoding a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

An LDPC encoder which, when the information data is input, encodes the information data according to a predetermined control to generate an LDPC code;

A spatial mapper which spatially maps the LDPC code according to a predetermined spatial mapping method to generate a space-time LDPC code;

A fifth partial matrix that is a combination of a second partial matrix of even columns of the first partial matrix corresponding to the information data and a fourth partial matrix of odd columns of the

third partial matrix corresponding to parity And an eighth partial matrix comprising a sixth partial matrix composed of odd-numbered columns of the first partial matrix and a seventh partial matrix composed of even-numbered columns of the third partial matrix. And a controller configured to generate the LDPC code by encoding the information data corresponding to a parity check matrix having a predetermined rank on a binary field and a ninth partial matrix exclusively OR of a partial matrix. .

19. The method of claim 18,

And said set rank is full rank.

20. The method of claim 19,

And the number of the transmitting antennas is 2, and further comprising a signal mapping unit for mapping the space-time LDPC code using a binary phase shift keying (BPSK) scheme.

21. The method of claim 20,

And the apparatus further comprises a serial / parallel converter for parallel converting the information data into first information data and second information data.

22. The method of claim 21,

The LDPC encoder;

A first LDPC encoder for encoding the first information data corresponding to the parity check matrix to generate a first codeword that is a first parity of the information data;

And a second LDPC encoder for generating a second coder that is a second parity of the information data by encoding the second information data corresponding to the parity check matrix.

23. The method of claim 22,

The spatial mapper includes even-numbered bits constituting the first information data, even-numbered bits constituting the second information data, odd-numbered bits constituting the first codeword, and the second code. Odd-numbered bits constituting the control word are transmitted through one of the two transmit antennas, even-numbered bits constituting the second codeword and even-numbered bits constituting the first codeword. And spatial mapping the odd-numbered bits constituting the second information data and the odd-numbered bits constituting the first information data to be transmitted through the other of the transmit antennas. .

24. An apparatus for encoding a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

When the information data is input, the spatiotemporal LDPC encoder generates the LDPC code by encoding the information data according to a predetermined control, and generates the spatiotemporal LDPC code by spatial mapping the LDPC code according to a preset spatial mapping scheme; ,

A first partial matrix corresponding to the first and third subwords, a second partial matrix corresponding to the second and fourth subwords, and a third partial exclusively OR of the first partial matrix and the second partial matrix A controller for controlling a matrix to generate the LDPC code by encoding the information data corresponding to a parity check matrix having a predetermined rank on a binary field,

The space-time LDPC code is divided into a real part space-time LDPC code and an imaginary part space-time LDPC code, and the real part space-time LDPC code is a first code transmitted through a first transmit antenna, which is an arbitrary transmit antenna among the plurality of transmit antennas. And a second coder transmitted through any second transmit antenna among the plurality of transmit antennas, and the imaginary part space-time LDPC code is divided into a third coder transmitted through the first

transmit antenna and the second transmit antenna. The apparatus is divided into a fourth codeword transmitted through the device.

25. The method of claim 24,

The apparatus further comprises a signal mapping unit for mapping the space-time LDPC code using any one of a quadrature phase shift keying (QPSK) scheme and a quadrature amplitude modulation (4-QAM) scheme.

26. The method of claim 25,

And the apparatus further comprises a serial / parallel converter for parallel converting the information data into real part information data and imaginary part information data.

27. The method of claim 26,

The space-time LDPC encoder;

A real part space-time LDPC encoder for generating a first coder by encoding the real part information data corresponding to the parity check matrix;

An imaginary part space-time LDPC encoder for generating a second coder by encoding the imaginary part information data corresponding to the parity check matrix,

The real part space-time LDPC coder and the imaginary part space-time LDPC coder may include preset bits among the bits constituting the first codeword and preset bits among the bits constituting the second codeword. The remaining bits except the set bits among the bits constituting the first codeword and the bits constituting the second codeword are transmitted through the second transmitting antenna. And generating space-time LDPC codes by spatial mapping to be transmitted.

28. The method of claim 27,

Wherein the parity check matrix is generated such that the real part space-time LDPC code and the imaginary part space-time LDPC code have a predetermined rank on a binary field.

29. The method of claim 28,

And said set rank is full rank.

30. A method of decoding a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

A first process of detecting a received signal through signals received through each of the plurality of receive antennas;

A second step of performing spatial mapping node decoding based on detection values of the received signals;

A third process of decoding the spatial mapped node decoded signal corresponding to a preset parity check matrix;

If the probability values of the signal decoded corresponding to the parity check matrix satisfy a preset repetition decoding interruption condition, the probabilities are hard-determined and output as information data. If the probability values do not satisfy the repetition decoding interruption condition, the iterative decoding is performed. And a fourth step of controlling to repeat the operations of the second to third processes until the interruption condition is satisfied.

31. The method of claim 30,

The parity check matrix is a combination of a second partial matrix composed of even columns of the first partial matrix corresponding to the information data and a fourth partial matrix composed of odd columns of the third partial matrix corresponding to the parity. The eighth partial matrix is a sixth partial matrix consisting of a fifth partial matrix and an eighth partial matrix including odd-numbered columns of the first partial matrix, and a seventh partial matrix including even-numbered columns of the third partial matrix. And a

predetermined rank on a ninth partial matrix and a binary field obtained by exclusively ORing a matrix and the third partial matrix.

32. The method of claim 31, wherein

The set rank is characterized in that the full rank (full rank).

33. 33. The method of claim 32,

The number of the transmitting antennas is two, the space-time LDPC code is characterized in that the signal mapping in the Binary Phase Shift Keying (BPSK) scheme.

34. The method of claim 30,

The parity check matrix is exclusive of a first partial matrix corresponding to a first codeword and a third codeword, a second partial matrix corresponding to a second codeword and a fourth codeword, and the first partial matrix and the second partial matrix. The logical OR of the third partial matrix has a predetermined rank on the binary field,

The space-time LDPC code is divided into a real part space-time LDPC code and an imaginary part space-time LDPC code, and the real part space-time LDPC code is a first code transmitted through a first transmit antenna, which is an arbitrary transmit antenna among the plurality of transmit antennas. And a second coder transmitted through any second transmit antenna among the plurality of transmit antennas, and the imaginary part space-time LDPC code is divided into a third coder transmitted through the first transmit antenna and the second transmit antenna. The method as claimed in claim 4, wherein the method is partitioned into fourth codewords transmitted through the control unit.

35. The method of claim 34, wherein

The set rank is characterized in that the full rank (full rank).

36. 36. The method of claim 35 wherein

The space-time LDPC code is signal mapping according to any one of a Quadrature Phase Shift Keying (QPSK) scheme and a Quadrature Amplitude Modulation (4-QAM) scheme.

37. A decoding apparatus of a space-time low density parity check (LDPC) code in a mobile communication system including a transmitter using a plurality of transmit antennas and a receiver using a plurality of receive antennas,

A detector for detecting a received signal through signals received through each of the plurality of receive antennas;

A spatial mapped node decoder for performing spatial mapped node decoding based on the detected values of the received signals;

An LDPC encoder which decodes the spatial mapped node decoded signal corresponding to a parity check matrix set in advance;

If the probability values of the signal decoded corresponding to the parity check matrix satisfy a preset repetition decoding interruption condition, the probabilities are hard-determined and output as information data. If the probability values do not satisfy the repetition decoding interruption condition, the iterative decoding is performed. And a controller for controlling to repeat the operations of the second to third processes until the stop condition is satisfied.

38. The method of claim 37,

The parity check matrix is a combination of a second partial matrix composed of even columns of the first partial matrix corresponding to the information data and a fourth partial matrix composed of odd columns of the third partial matrix corresponding to the parity. The eighth partial matrix is a sixth partial matrix consisting of a fifth partial matrix and an eighth partial matrix including odd-numbered columns of the first partial matrix, and a seventh partial matrix including even-numbered columns of the third partial matrix. And a predetermined rank on a binary field and a ninth partial matrix obtained by performing an exclusive OR on the matrix and the third partial matrix.

39. The method of claim 38,

And said set rank is full rank.

40. The method of claim 39,

The number of the transmitting antennas is two, the space-time LDPC code is characterized in that the signal mapping in the Binary Phase Shift Keying (BPSK) scheme.

41. The method of claim 37,

The parity check matrix is exclusive of a first partial matrix corresponding to a first codeword and a third codeword, a second partial matrix corresponding to a second codeword and a fourth codeword, and the first partial matrix and the second partial matrix. The logical OR of the third partial matrix has a predetermined rank on the binary field,

The space-time LDPC code is divided into a real part space-time LDPC code and an imaginary part space-time LDPC code, and the real part space-time LDPC code is a first code transmitted through a first transmit antenna, which is an arbitrary transmit antenna among the plurality of transmit antennas. And a second coder transmitted through any second transmit antenna among the plurality of transmit antennas, and the imaginary part space-time LDPC code is divided into a third coder transmitted through the first transmit antenna and the second transmit antenna. The apparatus is divided into a fourth codeword transmitted through the device.

42. The method of claim 41, wherein

And said set rank is full rank.

43. The method of claim 42, wherein

The space-time LDPC code is signal mapped to any one of a Quadrature Phase Shift Keying (QPSK) scheme and a Quadrature Amplitude Modulation (4-QAM) scheme.

Patent Citations (1)

Publication number	Priority date	Publication date	Assignee	Title
Family To Family Citations				
US7013116B2 *	2002-06-26	2006-03-14	Lucent Technologies Inc.	MIMO systems having a channel decoder matched to a MIMO detector

* Cited by examiner, † Cited by third party

Cited By (29)

Publication number	Priority date	Publication date	Assignee	Title
Family To Family Citations				
US9270410B2	2002-04-22	2016-02-23	Texas Instruments Incorporated	MIMO PGRC system and method
FR2841068B1 *	2002-06-14	2004-09-24	Comsis	Method for decoding linear space-time codes in a multi-antenna wireless transmission system, and decoder implementing such a method
JP4224329B2 *	2003-03-25	2009-02-12	パナソニック株式会社	Encoding apparatus and encoding method
KR20050118056A *	2004-05-12	2005-12-15	삼성전자주식회사	Method and apparatus for channel encoding and decoding in mobile communication systems using multi-rate block ldpc codes
US7882413B2 *	2005-01-20	2011-02-01	New Jersey Institute Of Technology	Method and/or system for space-time encoding and/or decoding
JP4212567B2 *	2005-03-31	2009-01-21	株式会社東芝	Receiving apparatus and demodulation method
JP4065283B2	2005-07-06	2008-03-19	松下電器産業株式会社	Sending method
JP2009505560A *	2005-08-19	2009-02-05	ミツビシ・エレクトリック・リサーチ・ラボラトリーズ・インコーポレイテッド	Optimal signaling and selection verification for transmit antenna selection with feedback with errors
US20070126585A1 *	2005-12-06	2007-06-07	Symbol Technologies, Inc.	System integration of RFID and MIMO technologies
US20070183533A1 *	2006-02-08	2007-08-09	Schmidl Timothy M	MIMO system with spatial diversity
KR100950645B1 *	2006-03-03	2010-04-01	삼성전자주식회사	Apparatus and method for transmitting/receiving a signal in a mobile communication system using multiple input multiple output scheme

FR2900294B1 *2006-04-192008-07-04St Microelectronics SaLoading the entry memory of an ldpc decoder with data to decoder
US8060803B2 *2006-05-162011-11-15Nokia CorporationMethod, apparatus and computer program product providing soft iterative recursive least squares (RLS) channel estimator
US8464120B2 *2006-10-182013-06-11Panasonic CorporationMethod and system for data transmission in a multiple input multiple output (MIMO) system including unbalanced lifting of a parity check matrix prior to encoding input data streams
CN102017445B *2007-01-052014-05-07Lg电子株式会社Layer mapping method and data transmission method for MIMO system
WO2008118168A1 *2007-03-232008-10-02Andrew WinklerMethod and system for backup storage of electronic data
KR101435681B1 *2007-11-082014-09-02삼성전자주식회사Method and apparatus for transmitting and receiving data in a communication system using low density parity check code
EP2107707B1 *2008-03-312017-08-23Google Technology Holdings LLCSpatial mapping of an OFDM signal to reduce attenuation from an individual transmit antenna in a mimo transmitter
US20090282316A1 *2008-05-072009-11-12Texas Instruments IncorporatedMemory Access in Low-Density Parity Check Decoders
US20090300461A1 *2008-05-292009-12-03Gadi ShorDevice, method and computer program product for communication
US20110113312A1 *2008-06-092011-05-12Hideki KobayashiCheck matrix generating method, check matrix, decoding apparatus, and decoding method
US8473824B1 *2008-09-082013-06-25Marvell International Ltd.Quasi-cyclic low-density parity-check (QC-LDPC) encoder
US8321752B1 *2008-11-122012-11-27Marvell International Ltd.Integrated 2-level low density parity check (LDPC) codes
JP4898858B2 *2009-03-022012-03-21パナソニック株式会社Encoder, decoder and encoding method
JP5790029B2 *2011-03-012015-10-07ソニー株式会社decoding device, decoding method, and program
KR101772008B1 *2011-03-032017-09-05삼성전자주식회사Method and apparatus for transmitting and receiving in communication/broadcasting system

US8826096B2 *2011-12-292014-09-02Korea Advanced Institute Of Science And TechnologyMethod of decoding LDPC code for producing several different decoders using parity-check matrix of LDPC code and LDPC code system including the same
US10243638B22016-10-042019-03-26At&T Intellectual Property I, L.P.Forward error correction code selection in wireless systems
US10270559B22016-10-042019-04-23At&T Intellectual Property I, L.P.Single encoder and decoder for forward error correction coding
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
US10038576B22018-07-31Method and apparatus for transmitting and receiving data in a communication system
RU2543553C22015-03-10Method and apparatus for channel encoding and decoding in communication system using low-density parity check codes
JP6096922B22017-03-15Check and irregular non-systematic IRA code encoding and decoding system and method
JP6297215B22018-03-20Method for adaptive modulation coding (AMC) and AMC controller
JP6497593B22019-04-10Transmission device, reception device, transmission method, and reception method
KR100739510B12007-07-13Apparatus and method for coding/decoding semi-systematic block low density parity check code
KR100683084B12007-02-15Method and system for providing short block length low density parity checkldpc codes in support of broadband satellite applications
EP2096761B12013-09-11Puncturing of LDPC codes in dependence on the modulation scheme
EP1525664B92015-09-02Method and system for memory management in low density parity check (ldpc) decoders
ES2273177T32007-05-01Method and appliance for coding low density parity check codes (ldpc) in long and short blocks.
JP4109241B22008-07-02Method and apparatus for providing reduced memory low density parity check (LDPC) code
KR100703536B12007-04-03Apparatus and method for encoding/decoding space time block code in a mobile communication system using multiple input multiple output scheme

CN101523757B2013-05-08Link adaptation for retransmission error-control technique transmissions
RU2354045C22009-04-27Device and method for encoding and decoding block codes with low density of parity control with variable coding rate
CN101133558B2010-10-06Method for transmitting data, method for receiving data, transmitter, receiver
JP5964969B22016-08-03Information transmission / reception method and apparatus in broadcasting / communication system
US10277249B22019-04-30Method and apparatus for channel encoding and decoding in a communication system using a low-density parity check code
KR101102396B12012-01-05Method of matching codeword size and transmitter therefor in mobile communications system
US7668224B22010-02-23Encoding for digital communications in a multiple-input, multiple-output environment
KR100856235B12008-09-03Apparatus and method for encoding and decoding block low density parity check codes with a variable coding rate
US7370265B22008-05-06Bandwidth efficient coded modulation scheme based on MLC (multi-level code) signals having multiple maps
US9479290B22016-10-25Method and apparatus for transmitting and receiving information in a broadcasting/communication system
JP4317196B22009-08-19Transmitting apparatus, receiving apparatus, signal transmitting method and signal receiving method in communication system using low density parity check code
KR100574306B12006-04-27Method and system for decoding low density parity checkldpc codes
US8151157B22012-04-03Method of encoding/decoding using low density check code matrix

Priority And Related Applications

Priority Applications (2)

ApplicationPriority dateFiling dateTitle
KR10200400079782004-02-06
KR200400079782004-02-06

Legal Events

DateCodeTitleDescription

2005-02-04A201Request for examination
2006-06-28E902Notification of reason for refusal
2006-08-29E701Decision to grant or registration of patent right
2006-09-22GRNTWritten decision to grant
2009-09-23LAPSLapse due to unpaid annual fee

(19)대한민국특허청(KR)
(12) 등록특허공보(B1)

(51) 。Int. Cl.
H04B 7/02 (2006.01)

(45) 공고일자 2006년09월29일
(11) 등록번호 10-0630177
(24) 등록일자 2006년09월22일

(21) 출원번호 10-2005-0010836
(22) 출원일자 2005년02월04일

(65) 공개번호 10-2006-0041769
(43) 공개일자 2006년05월12일

(30) 우선권주장 1020040007978 2004년02월06일 대한민국(KR)

(73) 특허권자 삼성전자주식회사
경기도 수원시 영통구 매탄동 416

(72) 발명자 정홍실
경기도 수원시 영통구 매탄동 1251-3호

윤석현
경기도 수원시 영통구 영통동 벽적골9단지아파트 944동 1509호

김재열
경기도 수원시 영통구 매탄1동 삼성3차 아파트 3동 408호

채찬병
서울특별시 동대문구 제기2동 벽산아파트 104동 1701호

(74) 대리인 이건주

심사관 : 정해영

(54) 최대 다이버시티 이득을 가지는 시공간 저밀도 패리티검사 부호 부호화/복호화 장치 및 방법

요약

본 발명은 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서, 정보 데이터를 입력받으면 상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정된 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호로 생성하고, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성한다.

대표도

도 5

색인어

시공간 LDPC 부호, 패리티 검사 행렬, 최대 다이버시티(full diversity), 최대 랭크(full rank), BPSK, QPSK, 4-QAM

명세서

도면의 간단한 설명

도 1은 일반적인 STC 방식을 사용하는 이동 통신 시스템의 송신기 구조를 개략적으로 도시한 도면

도 2는 일반적인 STC 방식을 사용하는 이동 통신 시스템의 수신기 구조를 개략적으로 도시한 도면

도 3은 일반적인 (8, 2, 4) LDPC 부호의 패리티 검사 행렬을 도시한 도면

도 4는 도 3의 (8, 2, 4) LDPC 부호의 factor 그래프를 도시한 도면

도 5는 신호 사상 방식으로 BPSK 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조를 개략적으로 도시한 도면

도 6은 신호 사상 방식으로 QPSK 방식 및 4-QAM 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조를 개략적으로 도시한 도면

도 7은 본 발명의 실시예에 따른 이동 통신 시스템의 수신기 동작 과정을 도시한 순서도

도 8은 본 발명의 실시예에서의 기능을 수행하는 이동 통신 시스템의 수신기 구조를 개략적으로 도시한 도면

발명의 상세한 설명

발명의 목적

발명이 속하는 기술 및 그 분야의 종래기술

본 발명은 이동 통신 시스템에 관한 것으로서, 특히 다이버시티 이득을 최대화하는 시공간 저밀도 패리티 검사 부호를 부호화/복호화하는 장치 및 방법에 관한 것이다.

통신에서 가장 근본적인 문제는 채널(channel)을 통하여 얼마나 효율적이고 신뢰성 있게(reliably) 데이터(data)를 전송할 수 있느냐 하는 것이다. 최근에 활발하게 연구되고 있는 차세대 멀티미디어 이동 통신 시스템에서는 초기의 음성 위주의 서비스를 벗어나 영상, 무선 데이터 등의 다양한 정보를 처리하고 전송할 수 있는 고속 통신 시스템이 요구됨에 따라 시스템에 적절한 채널 부호화 기법을 사용하여 시스템의 효율을 높이는 것이 필수적이다.

그런데, 이동 통신 시스템에 존재하는 무선 채널 환경은 유선 채널 환경과는 달리 다중 경로 간섭(multipath interference)과, 쉐도잉(shadowing)과, 전파 감쇠와, 시변 잡음과, 간섭 및 페이딩(fading) 등과 같은 여러 요인들로 인해 불가피한 오류가 발생하여 정보의 손실이 발생한다. 상기 정보 손실은 실제 송신 신호에 심한 왜곡을 발생시켜 상기 이동 통신 시스템 전체 성능을 저하시키는 요인으로 작용하게 된다. 일반적으로 이러한 정보의 손실을 감소시키기 위해 채널의 특성에 따라 다양한 에러 제어 기법(error-control technique)을 이용하여 상기 이동 통신 시스템의 신뢰도를 높이는데, 이러한 에러 제어 기법 중에 가장 기본적인 기법은 에러 정정 부호(error-correcting code)를 사용하는 것이다.

또한, 상기 이동 통신 시스템에서는 상기 페이딩 현상으로 인한 통신의 불안정성을 제거하기 위해 다이버시티(diversity) 방식을 사용하며, 이러한 다이버시티 방식 중 하나인 공간 다이버시티(space diversity) 방식은 다중 안테나(multiple antenna)를 사용한다. 여기서, 상기 공간 다이버시티 방식은 수신 안테나들을 다수개로 사용하는 수신 안테나 다이버시티 방식과, 송신 안테나들을 다수개로 사용하는 송신 안테나 다이버시티 방식 및 다수개의 수신 안테나들과 다수개의 송신 안테나들을 사용하는 다중 입력 다중 출력(MIMO: Multiple Input Multiple Output, 이하 'MIMO'라 칭하기로 한다) 방식으로 분류된다. 여기서, 상기 MIMO 방식은 일종의 시공간 부호화(STC: Space-Time Coding, 이하 'STC'라 칭하기로 한다) 방

이어서, 상기 STC 방식은 미리 설정된 부호화 방식으로 부호화된 신호를 다수개의 송신 안테나들을 사용하여 전송함으로써 시간 영역(time domain)에서의 부호화 방식을 공간 영역(space domain)으로 확장하여 보다 낮은 에러율을 달성하는 방식이다.

그러면 여기서 도 1을 참조하여 상기 STC 방식을 사용하는 이동 통신 시스템 송신기 구조에 대해서 설명하기로 한다.

상기 도 1은 일반적인 STC 방식을 사용하는 이동 통신 시스템의 송신기 구조를 개략적으로 도시한 도면이다.

상기 도 1을 참조하면, 상기 송신기는 부호화기(encoder)(110)와, 공간 사상기(spatial mapper)(120)와, 다수의, 일 예로 M개의 신호 사상기(signal mapper)들, 즉 제1신호 사상기(130-1) 내지 제M신호 사상기(130-M)와, 다수의, 일 예로 상기 M개의 송신 안테나(Tx ANT)들, 즉 제1송신 안테나(Tx ANT 1) 내지 제M송신 안테나(Tx ANT M)로 구성된다. 먼저, 정보 데이터(information data)가 입력되면 상기 부호화기(110)는 미리 설정되어 있는 부호화율(coding rate)에 따라 상기 정보 데이터를 미리 설정되어 있는 부호화 방식으로 부호화하여 부호화 심벌(coded symbol)로 생성한 후 상기 공간 사상기(120)로 출력한다. 일 예로, 상기 정보 데이터가 k비트이고, 상기 부호화율이 k/N일 경우 상기 부호화기(110)는 상기 k비트의 정보 데이터를 N비트로 구성된 부호화 심벌로 출력하는 것이다.

상기 공간 사상기(120)는 상기 부호화기(110)에서 출력한 부호화 심벌을 입력하여 미리 설정되어 있는 공간 사상 방식으로 공간 사상하여 상기 M개의 송신 안테나들을 통해 송신될 수 있도록 제어한 후 상기 공간 사상된 신호들 각각을 해당 신호 사상기들 각각으로 출력하도록 제어한다. 여기서, 상기 공간 사상기(120)가 사용하는 공간 사상 방식은 다양하게 존재할 수 있으며, 상기 도 1에서는 상기 공간 사상기(120)가 상기 부호화기(110)에서 출력한 부호화 심벌을 상기 부호화기(110)에서 출력한 순서대로 공간 사상한다고 가정하기로 한다. 일 예로, 상기 부호화기(110)에서 출력한 부호화 심벌이 $C = [c_1, c_2, \cdots, c_N]$ 라고 가정하면, 상기 공간 사상기(120)의 출력은 하기 수학식 1과 같은 $M \times \frac{N}{M}$ 시공간 부호어 행렬(codeword matrix)이 된다.

수학식 1

$$C = \begin{bmatrix} c_1 & c_{M+1} & \cdots & c_{N-M+1} \\ c_2 & c_{M+2} & \cdots & c_{N-M+2} \\ \cdots & & & \\ c_M & c_{M+M} & \cdots & c_N \end{bmatrix}$$

상기 공간 사상기(120)는 상기 송신 안테나들의 개수가 M개이므로 상기 부호화기(110)에서 출력한 부호화 심벌을 그 출력 순서에 따라 순차적으로 공간 사상하여 M개의 신호들로 분류하고, 상기 M개의 신호들 각각을 제1신호 사상기(130-1) 내지 제M신호 사상기(130-M) 각각으로 출력한다. 그러면 상기 제1신호 사상기(130-1) 내지 제M신호 사상기(130-M) 각각은 상기 공간 사상기(120)에서 출력한 신호를 입력하여 미리 설정되어 있는 방식으로 신호 사상한 후 해당 송신 안테나들, 즉 제1송신 안테나 내지 제M송신 안테나 각각을 통해 에어(air)상으로 송신한다. 여기서, 상기 제1신호 사상기(130-1) 내지 제M신호 사상기(130-M) 각각은 입력되는 신호를 상기 설정 방식에 따른 성상도(constellation)에 따라 신호 사상한 후 상기 제1송신 안테나 내지 제M송신 안테나 각각으로 출력한다. 일 예로, 상기 제1신호 사상기(130-1) 내지 제M신호 사상기(130-M) 각각은 그 입력 신호의 비트수 n이 1일 경우(n = 1) 전폭과 위상이 일정한 캐리어(carrier)를 이용하여 그 위상을 송신 부호에 대응시켜 변환하는 방식인 BPSK(Binary Phase Shift Keying) 방식으로 상기 입력 신호를 신호 사상하며, 상기 입력 신호의 비트수 n이 2일 경우(n = 2) QPSK(Quadrature Phase Shift Keying) 방식으로 상기 입력 신호를 신호 사상하며, 상기 입력 신호의 비트수 n이 3일 경우(n = 3) 8QAM(Quadrature Amplitude Modulation) 방식으로 상기 입력 신호를 신호 사상한다. 상기 도 1에서는 k 비트의 정보 데이터를 M개의 송신 안테나들을 통해 전송하므로 그 부호화율은 $M \times \frac{k}{N}$ 이 된다.

상기 도 1에서는 일반적인 STC 방식을 사용하는 이동 통신 시스템의 송신기 구조에 대해서 설명하였으며, 다음으로 도 2를 참조하여 일반적인 STC 방식을 사용하는 이동 통신 시스템의 수신기 구조에 대해서 설명하기로 한다.

상기 도 2는 일반적인 STC 방식을 사용하는 이동 통신 시스템의 수신기 구조를 개략적으로 도시한 도면이다.

상기 도 2를 참조하면, 상기 수신기는 다수의, 일 예로 P개의 수신 안테나(Rx ANT)들, 즉 제1수신 안테나(Rx ANT 1) 내지 제P수신 안테나(Rx ANT P)와, 검파기(detector)(210)와, 공간 역사상기(spatial de-mapper)(220)와, 시공간 복호화기(space time decoder)(230)와, 공간 사상기(240)로 구성된다. 상기 도 2에서는 상기 수신기에 대응하는 송신기의 송신 안테나들의 개수와 상기 수신기의 수신 안테나들의 개수가 상이할 경우를 가정하였으나, 즉, 상기 송신기의 송신 안테나들의 개수는 상기 도 1에서 설명한 바와 같이 M개이고, 상기 수신기의 수신 안테나들의 개수는 P개인 경우를 가정하였으나, 상기 송신기의 송신 안테나들의 개수와 상기 수신기의 수신 안테나들의 개수는 동일할 수도 있음은 물론이다.

먼저, 상기 송신기에서 다수개, 즉 M개의 송신 안테나들을 통해 송신된 신호는 상기 제1수신 안테나 내지 제P수신 안테나 각각을 통해 수신된다. 상기 제1수신 안테나 내지 제P수신 안테나 각각은 상기 수신한 신호를 상기 검파기(210)로 출력한다. 상기 검파기(210)는 상기 제1수신 안테나 내지 제P수신 안테나 각각에서 출력한 수신 신호를 입력하여 상기 제1송신 안테나 내지 제M송신 안테나 각각에서 송신한 송신 신호를 검파한 후 상기 공간 역사상기(220)로 출력한다.

상기 공간 역사상기(220)는 상기 검파기(210)에서 출력한 신호를 입력하여 상기 송신기에서 적용한 공간 사상 방식에 대응되는 공간 역사상 방식으로 공간 역사상하여 상기 시공간 복호화기(230)로 출력한다. 상기 시공간 복호화기(230)는 상기 공간 역사상기(220)에서 출력한 신호를 입력하여 상기 송신기에서 적용한 부호화 방식에 대응되는 복호화 방식으로 복호화하여 출력한다. 여기서, 상기 시공간 복호화기(230)에서 출력하는 신호는 무선 채널 환경 등에 의한 에러가 존재하지 않을 경우 상기 송신기에서 정보 데이터와 동일하나 실제 무선 채널 환경에서는 에러가 발생하므로 보다 신뢰성있는 복호화를 위해 반복 복호화(iterative decoding)를 수행할 수 있다.

상기 반복 복호화를 위해 상기 시공간 복호화기(230)는 상기 송신기에서 적용한 부호화 방식에 대응되는 방식으로 복호화한 신호를 상기 공간 사상기(240)로 출력한다. 상기 공간 사상기(240)는 상기 시공간 복호화기(230)에서 출력한 신호를 입력하여 상기 송신기에서 적용한 공간 사상 방식에 대응되는 공간 사상 방식으로 시공간 복호화한 후 상기 검파기(210)로 출력하여 반복 복호화를 수행하도록 제어한다. 이렇게 반복 복호화를 수행할 경우 정보 데이터의 복원의 신뢰도가 증가하게 되는 것이며, 상기와 같은 반복 복호화를 통해 상기 시공간 복호화기(230)에서 신뢰성있는 신호로 복호화되면 상기 시공간 복호화기(230)에서 출력하는 신호를 최종 정보 데이터로 출력하게 되는 것이다.

한편, 시공간 부호는 상기 STC 방식으로 부호화된 부호를 나타내며, 상기 STC 방식은 상기에서 설명한 바와 같이 시간 영역에서의 부호화 방식을 공간 영역까지 확장시킴으로써 부가 정보를 추가적으로 획득하게 된다. 즉, 상기 STC 방식은 상기 시간 영역에서 부가적인 정보를 제공하여 에러율(error rate)을 감소시킬 뿐만 아니라, 공간 영역에서의 다이버시티 이득(diversity gain)을 증가시킴으로써 보다 정확한 신호 복조가 가능하도록 한다. 여기서, 상기 다이버시티 이득은 시공간 부호의 신호대 잡음비(SNR: Signal to Noise Ratio, 이하 'SNR'이라 칭하기로 한다)와 에러 확률을 로그가(log scale)로 표현할 경우 그 기울기에 해당하므로, 상기 시공간 부호를 설계할 경우 상기 다이버시티 이득의 극대화(이하 '최대 다이버시티 이득(full diversity gain)'이라 칭하기로 한다)가 상기 시공간 부호의 성능을 좌우하는 가장 중요한 요소로 작용하게 된다.

상기 최대 다이버시티 이득을 얻기 위해서는 상기 시공간 부호를 설계함에 있어 다음과 같은 시공간 부호 설계 조건 1 내지 시공간 부호 설계 조건 3을 만족해야만 한다.

[시공간 부호 설계 조건 1]

최대 다이버시티 이득을 얻기 위해서는 행렬 B(c,e)가 임의의 시공간 부호어 c와 시공간 부호어 e에 대하여 최대 랭크(full rank)를 가져야한다. 여기서, 상기 행렬 B(c,e)은 상기 시공간 부호어 c와 상기 시공간 부호어 e의 차를 나타내는 행렬이다.

또한, BPSK 방식을 사용하여 신호를 사상할 경우, 특히 시공간 부호 설계 조건 2를 만족해야만 한다.

[시공간 부호 설계 조건 2]

모든 0이 아닌 시공간 부호어 행렬 c가 이진 필드(binary field)상에서 최대 랭크를 가지면 시공간 부호어는 최대 다이버시티 이득을 가진다.

또한, QPSK 방식을 사용하여 신호를 사상할 경우, 특히 시공간 부호 설계 조건 3을 만족해야만 한다.

[시공간 부호 설계 조건 3]

모든 0이 아닌 시공간 부호어 행렬 c가 이진 필드상으로 사상되어 사상된 시공간 부호어 행렬 $\Phi(\mathbf{c})$가 이진 필드상에서 최대 랭크를 가지면 시공간 부호어는 최대 다이버시티 이득을 가진다.

상기에서 설명한 바와 같이 상기 시공간 부호는 최대 다이버시티 이득을 획득하기 위해 최대 랭크를 가지는 형태로 설계된다. 즉, 상기 시공간 부호는 그 생성 행렬(G: Generating matrix)이 최대 랭크를 가지는 형태로 설계된다.

한편, 차세대 통신 시스템은 다양한 서비스 품질(Quality of Service; 이하 'QoS' 칭하기로 함)을 가지는 고속의 대용량 데이터를 서비스하는 형태로 발전하고 있다. 이렇게, 고속의 대용량 데이터를 서비스함에 있어서 전송상의 정보 손실은 치명적이며, 따라서 상기 에러 정정 부호의 에러 정정 능력이 전체 서비스 품질을 좌우하는 중요한 요소로 작용하고 있다. 상기 에러 정정 부호의 대표적인 부호들로는 터보 부호(turbo code)와, 저밀도 패리티 검사(LDPC: Low Density Parity Check, 이하 'LDPC'라 칭하기로 함) 부호 등이 있다.

상기 LDPC 부호는 종래 에러 정정을 위해 주로 사용되던 컨벌루셔널 부호(convolutional code)에 비하여 고속 데이터 전송시에 성능 이득이 우수한 것으로 알려져 있으며, 무선 채널에서 발생하는 잡음에 의한 오류를 효과적으로 정정하여 데이터 전송의 신뢰도를 높일 수 있다는 장점을 가진다. 또한, 상기 LDPC 부호는 팩터(factor, 이하 'factor'라 칭하기로 함) 그래프 상에서 합곱(sum-product) 알고리즘(algorithm)에 기반한 반복 복호화 알고리즘을 사용하여 복호화할 수 있다는 장점을 가진다. 상기 합곱 알고리즘에 기반한 반복 복호화 알고리즘을 사용하는 복호화 방법을 사용함으로써 상기 LDPC 부호의 복호화기는 상기 터보 부호의 복호화기에 비해 낮은 복잡도를 가질 뿐만 아니라 병렬 처리 복호화기를 구현함에 있어 용이하게 된다.

상기 터보 부호는 Shannon의 채널 부호화 이론(channel coding theorem)의 채널 용량 한계에 근접하는 우수한 성능을 가지고 있으며, 지금까지 알려진 최고의 성능을 가지는 LDPC 부호는 블록 크기 10^7 을 사용하여 비트 에러율(BER: Bit Error Rate) 10^{-5} 에서 Shannon의 채널 부호화의 채널 용량 한계에서 단지 0.04[dB] 정도의 차이를 가지는 성능을 나타낸다. 여기서, 상기 Shannon의 채널 부호화 이론은 채널의 용량을 초과하지 않는 전송률(data rate)에 한해 신뢰성 있는 통신이 가능함을 나타내는 이론이다. 블록(block) 크기가 굉장히 큰 랜덤(random) 부호는 Shannon의 채널 부호화 이론에서 채널 용량 한계에 근접하는 성능을 보이지만, MAP(maximum a posteriori) 또는 ML(maximum likelihood) 복호화를 적용할 경우 그 계산량에 있어 굉장한 로드(load)가 존재하여 실제 구현이 불가능하였다.

그런데, 상기 LDPC 부호는 대부분의 엘리먼트(element)들이 0의 값(zero value)을 가지며, 상기 0의 값을 가지는 엘리먼트들 이외의 극히 소수의 엘리먼트들이 0이 아닌 값(non-zero value), 일 예로 1의 값을 가지는 패리티 검사 행렬(parity check matrix)에 의해 정의된다. 이하, 설명의 편의상 상기 0이 아닌 값은 1의 값이라고 가정하여 설명하기로 한다. 따라서, 상기 LDPC 부호의 패리티 검사 행렬은 작은 웨이트(weight)를 가지기 때문에, 비교적 긴 길이를 가지는 블록 부호(block code)에서도 반복 복호화를 통해 복호화가 가능하며, 블록 부호의 블록 길이를 계속 증가시켜가면 터보 부호와 같이 Shannon의 채널 용량 한계에 근접하는 형태의 성능을 나타낸다. 여기서, 상기 웨이트라함은 상기 패리티 검사 행렬을 구성하는 엘리먼트들 중 0이 아닌 값을 가지는 엘리먼트들의 개수를 나타낸다. 따라서, 상기 차세대 통신 시스템에서는 상기 에러 정정 부호로서 상기 LDPC 부호를 적극적으로 사용하고 있는 추세에 있다.

그러나, 상기 LDPC 부호의 경우 상기 시공간 부호와 같이 생성 행렬을 이용하여 부호화를 수행할 경우 우수한 성능을 보장할 수가 없다. 즉, 상기 LDPC 부호의 경우 상기에서 설명한 바와 같이 패리티 검사 행렬의 웨이트가 작아 복호화시 그 복잡도가 낮다는 장점을 가지고 있으나, 상기 패리티 검사 행렬을 생성 행렬로 변환할 경우 상기 생성 행렬의 웨이트가 증가하여 부호화시 그 복잡도가 증가하게 된다.

결과적으로, 상기 LDPC 부호를 사용하여 상기 시공간 부호를 설계할 경우 일반적인 시공간 부호의 생성 행렬 관점에의 설계 조건들을 그대로 적용하는 것이 난이하게 된다. 따라서 상기 LDPC 부호를 사용하는 시공간 부호가 상기 최대 다이버시티 이득을 획득할 수 있는 방법, 즉 상기 최대 다이버시티 이득을 획득할 수 있는 패리티 검사 행렬에 대한 필요성이 대두되고 있다.

발명이 이루고자 하는 기술적 과제

따라서, 본 발명의 목적은 이동 통신 시스템에서 최대 다이버시티 이득을 획득하는 시공간 LDPC 부호를 부호화/복호화하는 장치 및 방법을 제공함에 있다.

본 발명의 다른 목적은 다수개의 송신 안테나들을 사용하는 이동 통신 시스템에서 다수개의 송신 안테나들간에 상관 관계를 가지는 시공간 LDPC 부호를 부호화/복호화하는 장치 및 방법을 제공함에 있다.

본 발명의 다른 목적은 이동 통신 시스템에서 최대 다이버시티 이득을 획득하는 시공간 LDPC 부호의 패리티 검사 행렬의 설계 방법을 제공함에 있다.

상기한 목적들을 달성하기 위한 본 발명의 장치는; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 장치에 있어서, 정보 데이터가 입력되면, 소정 제어에 따라 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 LDPC 부호화부와, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 공간 사상기와, 상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제1부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 패리적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하도록 제어하는 제어기를 포함함을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 다른 장치는; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 장치에 있어서, 정보 데이터가 입력되면, 소정 제어에 따라 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하고, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 시공간 LDPC 부호화부와, 제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 패리적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하도록 제어하는 제어기를 포함하며, 상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기 다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할되고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 또 다른 장치는; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 복호화 장치에 있어서, 상기 다수의 수신 안테나를 각각을 통해 수신되는 신호들을 수신 신호 결합하는 결합기와, 상기 수신 신호들의 결합값을 기반으로 하여 공간 사상 노드 복호화를 수행하는 공간 사상 노드 복호화기와, 상기 공간 사상 노드 복호화된 신호를 미리 설정되어 있는 패리티 검사 행렬에 상응하게 복호화하는 LDPC 부호화부와, 상기 패리티 검사 행렬에 상응하게 복호화한 신호의 확률값들이 미리 미리 설정한 반복 복호화 중단 조건을 만족하면 상기 확률값들을 정합정하여 정보 데이터로 출력하고, 상기 확률값들이 상기 반복 복호화 중단 조건을 만족하지 않으면 상기 반복 복호화 중단 조건을 만족할 때까지 상기 제2과정 내지 제3과정의 동작을 반복하여 수행하도록 제어하는 제어기를 포함함을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 방법은; 다수개의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 패리티 검사 행렬을 생성하는 방법에 있어서, 정보 데이터를 시공간 LDPC 부호로 부호화시 적용되는 부호화율과 상기 시공간 LDPC 부호의 부호어 길이에 대응하도록 패리티 검사 행렬의 크기를 결정하는 과정과, 상기 결정된 크기의 패리티 검사 행렬을 상기 정보 데이터에 대응되는 제1부분 행렬과 패리티에 대응되는 제2부분 행렬로 분할하는 과정과, 상기 제1부분 행렬의 짝수번째(even) 열들로 구성된 제3부분 행렬과, 상기 제2부분 행렬의 홀수번째(odd) 열들로 구성된 제4 부분 행렬과, 상기 제3 부분행렬과 제4 부분행렬을 결합한 행렬인 제5부분 행렬과, 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과, 상기 제2부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬과, 상기 제6 부분 행렬과 제7 부분 행렬을

결합한 행렬인 제8부분 행렬과, 상기 제1부분 행렬과 상기 제2부분 행렬을 배타적 논리합한 제9부분 행렬을 생성하는 과정과, 상기 제5 부분 행렬 및 제8 부분 행렬이 상기 제9부분 행렬과 이전 컬럼상에서 미리 설정한 랭크를 가지도록 상기 패리티 검사 행렬을 생성하는 과정을 포함함을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 다른 방법은; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 패리티 검사 행렬을 생성하는 방법에 있어서, 정보 데이터를 시공간 LDPC 부호로 부호화시 적용되는 부호화율과 상기 시공간 LDPC 부호의 부호어 길이에 대응하도록 패리티 검사 행렬의 크기를 결정하는 과정과, 상기 시공간 LDPC 부호를 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할하고, 상기 실수 파트 시공간 LDPC 부호를 제1송신 안테나를 통해 송신되는 제1부호어와 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할하고, 상기 허수 파트 시공간 LDPC 부호를 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할하는 과정과, 상기 패리티 검사 행렬을 상기 제1부호어와 제2부호어에 대응되는 제1부분 행렬과 상기 제2부호어와 제4부호어에 대응되는 제2부분 행렬로 분할하는 과정과, 상기 제1부분 행렬과, 상기 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 배타적 논리합한 제3부분 행렬이 이전 컬럼상에서 미리 설정한 랭크를 가지도록 상기 패리티 검사 행렬을 생성하는 과정을 포함함을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 또 다른 방법은; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 방법에 있어서, 정보 데이터를 입력하는 과정과, 상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이전 컬럼상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 과정과, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 과정을 포함함을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 또 다른 방법은; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 방법에 있어서, 정보 데이터를 입력하는 과정과, 제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 배타적 논리합한 제3부분 행렬이 이전 컬럼상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 과정과, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 과정을 포함하며, 상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기 다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제3부호어로 분할되고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 한다.

상기한 목적들을 달성하기 위한 본 발명의 또 다른 방법은; 다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 복호화 방법에 있어서, 상기 다수의 수신 안테나들 각각을 통해 수신되는 신호들을 수신 신호 검파하는 제1과정과, 상기 수신 신호들의 검파값을 기반으로 하여 공간 사상 노프 복호화를 수행하는 제2과정과, 상기 공간 사상 노프 복호화된 신호를 미리 설정되어 있는 패리티 검사 행렬에 상응하게 복호화하는 제3과정과, 상기 패리티 검사 행렬에 상응하게 복호화한 신호의 확률값들이 미리 미리 설정한 반복 복호화 중단 조건을 만족하면 상기 확률값들을 경판정하여 정보 데이터로 출력하고, 상기 확률값들이 상기 반복 복호화 중단 조건을 만족하지 않으면 상기 반복 복호화 중단 조건을 만족할 때까지 상기 제2과정 내지 제3과정의 동작을 반복하여 수행하도록 제어하는 제4과정을 포함함을 특징으로 한다.

발명의 구성 및 작용

이하, 본 발명에 따른 첨부한 도면을 참조하여 상세히 설명한다. 하기의 설명에서는 본 발명에 따른 동작을 이해하는데 필요한 부분만이 설명되며 그 이외 부분의 설명은 본 발명의 요지를 흐트리지 않도록 생략될 것이라는 것을 유의하여야 한다.

본 발명은 오류 정정 부호(error-correcting code)인 저밀도 패리티 검사(LDPC: Low Density Parity Check, 이하 'LDPC'라 칭하기로 한다) 부호를 사용하는 시공간 부호, 즉 시공간 LDPC 부호에 대한 부호화/복호화 방안을 제안한다. 특히, 본 발명은 최대 다이버시티 이득(full diversity gain)을 가지는 시공간 LDPC 부호에 대한 부호화/복호화 방안을 제안한다.

그러면, 본 발명을 설명하기에 앞서 상기 LDPC 부호에 대해서 설명하기로 한다.

상기 LDPC 부호는 팩터(factor, 이하 'factor'라 칭하기로 한다) 그래프 상에서 합곱(sum-product) 알고리즘(algorithm)에 기반한 반복 복호화(iterative decoding) 알고리즘을 사용하여 복호화할 수 있다. 상기 합곱 알고리즘에 기반한 반복 복호화 알고리즘을 사용하는 복호화 방법을 사용함으로써 상기 LDPC 부호의 복호화기(decoder)는 터보 부호(turbo code)의 복호화기에 비해 낮은 복잡도를 가질 뿐만 아니라 병렬 처리 복호화기를 구현함에 있어 용이하게 된다. 상기 LDPC 부호를 factor 그래프로 표현하면 상기 LDPC 부호의 factor 그래프상에 사이클(cycle)이 존재하게 되는데, 사이클이 존재하는 상기 LDPC 부호의 factor 그래프 상에서의 반복 복호화는 준최적(suboptimal)이라는 것은 이미 잘 알려져 있는 사실이다. 또한 상기 LDPC 부호는 반복 복호화를 통해 우수한 성능을 가진다는 것 역시 실험적으로 입증된 바 있다.

또한, 상기 LDPC 부호는 Gallager에 의해 제안된 부호이며, 대부분의 엘리먼트(element)들이 0의 값(non-zero value), 일 예로 1의 값을 가지며, 상기 0의 값을 가지는 엘리먼트들 이외의 극히 소수의 엘리먼트들이 0이 아닌 값(non-zero value), 일 예로 1의 값을 가지는 패리티 검사 행렬(parity check matrix)에 의해 정의된다. 이하, 설명의 편의상 상기 0이 아닌 값은 1의 값이라고 가정하여 설명하기로 한다. 따라서, 상기 LDPC 부호의 패리티 검사 행렬은 작은 웨이트(weight)를 가지기 때문에, 비교적 긴 길이를 가지는 블록 부호(block code)에서도 반복 복호화를 통해 복호화가 가능하며, 블록 부호의 블록 길이를 계속 증가시키면 터보 부호와 같이 Shannon의 채널 용량 한계에 근접하는 성능을 나타낸다. 여기서, 상기 웨이트라함은 상기 패리티 검사 행렬을 구성하는 엘리먼트들 중 0이 아닌 값을 가지는 엘리먼트들의 개수를 나타낸다. 일 예로, (N, j, k) LDPC 부호는 블록(block) 길이가 N인 선형 블록 부호(linear block code)로, 각 열(column)마다 j개의 1의 값을 가지는 엘리먼트들과, 각 행(row)마다 k개의 1의 값을 가지는 엘리먼트들을 가지고, 상기 1의 값을 가지는 엘리먼트들을 제외한 엘리먼트들은 모두 0의 값을 가지는 엘리먼트들로 구성된 성긴(sparse, 이하 'sparse'라 칭하기로 한다) 구조의 패리티 검사 행렬에 의해 정의된다.

그러면 여기서 도 3을 참조하여 (N, j, k) LDPC 부호, 일 예로 (8, 2, 4) LDPC 부호의 패리티 검사 행렬에 대해서 설명하기로 한다.

상기 도 3은 일반적인 (8, 2, 4) LDPC 부호의 패리티 검사 행렬을 도시한 도면이다.

상기 도 3을 참조하면, 먼저 상기 (8, 2, 4) LDPC 부호의 패리티 검사 행렬 H는 8개의 열들과 4개의 행들로 구성되어 있으며, 각 열의 웨이트는 2로 균일하며, 각 행의 웨이트는 4로 균일하다. 이렇게, 상기 패리티 검사 행렬내 각 열의 웨이트와 각 행의 웨이트가 균일하므로 상기 (8, 2, 4) LDPC 부호는 균일 LDPC 부호(regular LDPC code)가 되는 것이다. 만약, 상기 패리티 검사 행렬내 각 열의 웨이트와 각 행의 웨이트가 균일하지 않을 경우 그 LDPC 부호는 불균일 LDPC 부호(irregular LDPC code)가 되는 것이다.

상기 도 3에서는 (8, 2, 4) 균일 LDPC 부호의 패리티 검사 행렬에 대해서 설명하였으며, 다음으로 도 4를 참조하여 상기 도 3에서 설명한 (8, 2, 4) 균일 LDPC 부호의 factor 그래프에 대해서 설명하기로 한다.

상기 도 4는 도 3의 (8, 2, 4) LDPC 부호의 factor 그래프를 도시한 도면이다.

상기 도 4를 참조하면, 상기 (8, 2, 4) LDPC 부호의 factor 그래프는 8개의 변수 노드(variable node)들, 즉 x_1(400)과, x_2(402)과, x_3(404)과, x_4(406)과, x_5(408)과, x_6(410)과, x_7(412)과, x_8(414)와, 4개의 검사 노드(check node)들(416,418,420,422)로 구성된다. 상기 (8, 2, 4) LDPC 부호의 패리티 검사 행렬의 i번째 열과 j번째 행이 교차하는 지점에 0이 아닌 값을 가지는 엘리먼트, 즉 1의 값을 가지는 엘리먼트가 존재할 경우 변수 노드 xi와 j번째 검사 노드 사이에 브랜치(branch)가 생성된다.

상기 LDPC 부호의 패리티 검사 행렬은 매우 작은 웨이트를 가지기 때문에, 비교적 긴 길이를 가지는 블록 부호(block code)에서도 반복 복호화를 통해 복호화가 가능하며, 블록 부호의 블록 길이를 계속 증가시키면 터보 부호와 같이

Shannon의 채널 용량 한계에 근접하는 형태의 성능을 나타낸다. 또한, MacKay와 Neal은 호퉁 전달 방식(flooding transfer scheme)을 사용하는 LDPC 부호의 반복 복호화 과정이 터보 부호의 반복 복호화 과정에 거의 근접하는 성능을 가진다는 것을 이미 증명한 바가 있다.

따라서, 본 발명에서는 우수한 성능을 보장하는 LDPC 부호를 다수의 송신 안테나(Tx.ANT)들을 통해 송신할 경우 최대 다이버시티 이득을 얻을 수 있는 시공간 LDPC 부호화/복호화 방안을 제안한다.

그러면 여기서 도 5를 참조하여 본 발명의 실시예에서의 기능을 수행하는 이동 통신 시스템의 송신기 구조, 특히 신호 사상 방식으로 BPSK 방식을 사용할 경우의 이동 통신 시스템의 송신기 구조에 대해서 설명하기로 한다.

상기 도 5는 신호 사상 방식으로 BPSK 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조를 개략적으로 도시한 도면이다.

상기 도 5를 참조하면, 먼저 상기 송신기는 직렬/병렬 변환기(serial to parallel converter)(500)와, 메모리(memory)(502)와, 제어기(controller)(504)와, 제 1 LDPC 부호화기(LDPC encoder)(506)와, 제 2 LDPC 부호화기(508)와, 공간 사상기(spatial mapper)(510)와, 다수의, 일 예로 2개의 신호 사상기(signal mapper)들, 즉 제1 신호 사상기(512) 내지 제2신호 사상기(514)와, 다수의, 일 예로 상기 2개의 송신 안테나(Tx.ANT)들, 즉 제1송신 안테나(Tx.ANT 1)와 제2송신 안테나(Tx.ANT 2)로 구성된다. 여기서, 상기 제1신호 사상기(512)와, 제2신호 사상기(514) 각각은 BPSK(Binary Phase Shift Keying) 방식으로 신호를 사상한다고 가정하기로 한다.

먼저 상기 메모리(502)는 본 발명에서 제안하는 시공간 LDPC 부호의 설계 방법에 따른 패리티 검사 행렬을 저장한다. 본 발명에서 제안하는 시공간 LDPC 부호의 설계 방법에 따른 패리티 검사 행렬은 하기에서 구체적으로 설명할 것이므로 여기서는 그 상세한 설명을 생략하기로 한다. 먼저, 정보 데이터(information data) s가 발생하면, 상기 정보 데이터 s는 상기 직렬/병렬 변환기(500)로 입력되고, 상기 직렬/병렬 변환기(500)는 상기 정보 데이터 s를 제1정보 데이터 s_1과 제2정보 데이터 s_2로 병렬 변환하여 출력한다.

상기 제1정보 데이터 s_1는 상기 제 1 LDPC 부호화기(506)와 공간 사상기(510)로 입력되고, 상기 제2정보 데이터 s_2는 제 2 LDPC 부호화기(508)와 공간 사상기(510)로 입력된다. 상기 제 1 LDPC 부호화기(506)는 상기 제어기(504)의 제어에 따라 상기 제1정보 데이터 s_1를 미리 설정되어 있는 부호화율(coding rate)에 상응하게 부호화하여 LDPC 부호어(codeword), 즉 LDPC 부호화 심벌(coded symbol)로 생성한 후 상기 공간 사상기(510)로 출력한다. 여기서, 상기 제어기(504)는 상기 메모리(502)에 저장되어 있는 패리티 검사 행렬에 상응하게 상기 제 1 LDPC 부호화기(506)를 제어하여 상기 패리티 검사 행렬에 상응하는 LDPC 부호어를 생성하도록 제어한다. 일 예로, 상기 제1정보 데이터 s_1가 k비트이고, 상기 부호화율이 1/2라고 가정하면 상기 제 1 LDPC 부호화기(506)에서 출력하는 비트수 N은 2k가 된다(N = 2k). 여기서, 상기 제 1 LDPC 부호화기(506)에서 출력하는 LDPC 부호어는 전체 정보 데이터 s를 기준으로한 패리티(parity)의 일부분인 제1패리티(p_1)가 되는 것이다.

또한, 상기 제 2 LDPC 부호화기(508)는 상기 제어기(504)의 제어에 따라 상기 제2정보 데이터 s_2를 미리 설정되어 있는 부호화율에 상응하게 부호화하여 LDPC 부호어, 즉 LDPC 부호화 심벌로 생성한 후 상기 공간 사상기(510)로 출력한다. 여기서, 상기 제어기(504)는 상기 메모리(502)에 저장되어 있는 패리티 검사 행렬에 상응하게 상기 제 2 LDPC 부호화기(506)를 제어하여 상기 패리티 검사 행렬에 상응하는 LDPC 부호어를 생성하도록 제어한다. 일 예로, 상기 제2정보 데이터 s_2가 k비트이고, 상기 부호화율이 1/2라고 가정하면 상기 제 2 LDPC 부호화기(506)에서 출력하는 비트수 N은 2k가 된다(N = 2k). 여기서, 상기 제 2 LDPC 부호화기(508)에서 출력하는 LDPC 부호어는 전체 정보 데이터 s를 기준으로한 패리티의 일부분인 제2패리티(p_2)가 되는 것이다.

한편, 상기 메모리(502)에 저장되어 있는 상기 패리티 검사 행렬을 H라고 표현하기로 하며, 상기 패리티 검사 행렬 H는 정보 데이터에 대응되는 파트인 정보 파트(information part) H_s와 패리티에 대응되는 파트인 패리티 파트(parity part) H_p로 구분되며 하기 수학식 2와 같이 나타낼 수 있다.

수학식 2

$$H=[H_s \;\; H_p]$$

또한, 상기 패리티 검사 행렬 H의 크기는 상기 송신기의 부호화율 및 최종 부호어인 시공간 LDPC 부호의 길이에 상응하는 크기로 결정됨은 물론이다. 한편, 상기 제 1 LDPC 부호화기(506)에서 출력하는 2k 비트의 LDPC 부호어를 '제 1 LDPC 부호어 c_1'라고 표현하기로 하며, 상기 제 2 LDPC 부호화기(506)에서 출력하는 2k 비트의 LDPC 부호어를 '제 2 LDPC 부호어 c_2'라고 표현하기로 한다. 또한 상기 제 1 LDPC 부호어 c_1는 k 비트의 정보 부호어 s_1와 패리티 부호어 p_1로 구성되며, 상기 제 2 LDPC 부호어 c_2는 k 비트의 정보 부호어 s_2와 패리티 부호어 p_2로 구성되며, 따라서 상기 제 1 LDPC 부호어 c_1와 제 2 LDPC 부호어 c_2는 하기 수학식 3과 같이 나타낼 수 있다.

수학식 3

$$\mathbf{c}_1 = [\mathbf{s}_1 \ \ \mathbf{p}_1]$$

$$\mathbf{c}_2 = [\mathbf{s}_2 \ \ \mathbf{p}_2]$$

여기서, 상기 제1 LDPC 부호어 c_1과 제2 LDPC 부호어 c_2는 상기 패리티 검사 행렬 H와 하기 수학식 4와 같은 관계를 가진다.

수학식 4

$$\mathbf{H} \cdot \mathbf{c}_1 = [\mathbf{H}_s \ \ \mathbf{H}_p] \cdot [\mathbf{s}_1 \ \mathbf{p}_1]^T = 0$$

$$\mathbf{H} \cdot \mathbf{c}_2 = [\mathbf{H}_s \ \ \mathbf{H}_p] \cdot [\mathbf{s}_2 \ \mathbf{p}_2]^T = 0$$

또한, 상기 공간 사상기(510)는 상기 제 1 LDPC 부호화기(506)에서 출력한 제1 LDPC 부호어 c_1과 상기 제 2 LDPC 부호화기(508)에서 출력한 제 2 LDPC 부호어 c_2를 입력하여 미리 설정되어 있는 공간 사상 방식으로 공간 사상하여 시공간 LPDC 부호어 C로 생성한 후 상기 2개의 송신 안테나들, 즉 제1송신 안테나와 제2송신 안테나를 통해 송신되도록 해당 신호 사상기를 각각으로 출력하도록 제어한다. 상기 공간 사상기(510)가 상기 제1 LDPC 부호어 c_1과 제 2 LDPC 부호어 c_2를 공간 사상하는 방식은 다양하게 존재하며, 본 발명에서는 하기에서 설명하는 바와 같은 방식을 기반으로 하여 공간 사상한다고 가정하기로 한다.

상기 공간 사상기(510)는 상기 제 1 LDPC 부호화기(506)에서 출력한 제1 LDPC 부호어 c_1의 정보 부호어 s_1를 구성하는 짝수번째(even) 정보 비트 $\mathbf{s}^{e}_{1}$와, 제 2 LDPC 부호화기(508)에서 출력한 제2 LDPC 부호어 c_2의 정보 부호어 s_2를 구성하는 짝수번째 정보 비트 $\mathbf{s}^{e}_{2}$와, 제 1 LDPC 부호화기(506)에서 출력한 제1LDPC 부호어 c_1의 패리티 부호어 p_1를 구성하는 홀수번째(odd) 패리티 비트 $\mathbf{p}^{o}_{1}$와 제 2 LDPC 부호화기(508)에서 출력한 제2LDPC 부호어 c_2의 패리티 부호어 p_2를 구성하는 홀수번째 패리티 비트 $\mathbf{p}^{o}_{2}$를 상기 제 1 송신 안테나를 통해 송신하도록 공간 사상한다. 또한, 상기 공간 사상기(510)는 상기 제 2 LDPC 부호화기(508)에서 출력한 제2 LDPC 부호어 c_2의 패리티 부호어 p_2를 구성하는 짝수번째 패리티 비트 $\mathbf{p}^{e}_{2}$와, 제 1 LDPC 부호화기(506)에서 출력한 제1 LDPC 부호어 c_1의 패리티 부호어 p_1를 구성하는

짝수번째 패리티 비트 $\mathbf{p}_1^e$와, 제2 LDPC 부호화기(508)에서 출력한 제2 LDPC 부호어 c_2의 정보 부호어 s_2를 구성하는 홀수번째 정보 비트 $\mathbf{s}_2^o$와 제1 LDPC 부호화기(506)에서 출력한 제1 LDPC 부호어 c_1의 정보 부호어 s_1를 구성하는 홀수번째 정보 비트 $\mathbf{s}_1^o$를 제2송신 안테나를 통해 송신하도록 공간 사상한다.

상기와 같은 방식으로 상기 제1 LDPC 부호어 c_1과 제 2 LDPC 부호어 c_2를 공간 사상할 경우 상기 공간 사상기(510)의 출력은 $2\times k$행렬인 시공간 LDPC 부호어 C로, 하기 수학식 5와 같이 나타낼 수 있다.

수학식 5

$$\mathbf{C}=\begin{bmatrix} \mathbf{s}_1^e & \mathbf{s}_2^e & \mathbf{p}_1^o & \mathbf{p}_2^o \\ \mathbf{p}_2^e & \mathbf{p}_1^e & \mathbf{s}_2^o & \mathbf{s}_1^o \end{bmatrix}$$

그런데, 상기 제1신호 사상기(512)와, 제2신호 사상기(514) 각각은 BPSK 방식으로 입력되는 신호를 신호 사상하므로 상기 시공간 LDPC 부호어 C는 이진(binary) 랭크(rank) 설계 규칙을 만족해야만 한다. 즉, 상기 종래 기술 부분에서 설명한 바와 같이 모든 0이 아닌 시공간 부호어는 이진 필드(binary field)상에서 최대 랭크(full rank)를 가져야 그 시공간 부호어가 최대 다이버시티 이득을 가지게 된다. 따라서, 상기 시공간 LDPC 부호어 C가 최대 다이버시티 이득을 가지기 위해서는 상기 이진 필드상에서 최대 랭크를 가져야만 하며, 따라서 상기 패리티 검사 행렬 H는 하기 패리티 검사 행렬 설계 조건 1을 만족하도록 설계되어야만 한다.

[패리티 검사 행렬 설계 조건 1]

$\mathbf{H}_1=[\mathbf{H}_s^e\ \mathbf{H}_p^o], \mathbf{H}_2=[\mathbf{H}_s^o\ \mathbf{H}_p^e]$ 와 $\mathbf{H}_s \oplus \mathbf{H}_p$가 이진 필드상에서 최대 랭크를 가져야 한다.

상기 패리티 검사 행렬 설계 조건 1에서 상기 행렬 $\mathbf{H}_1=[\mathbf{H}_s^e\ \mathbf{H}_p^o]$은 상기 행렬 $\mathbf{H}_s$에서 짝수번째 열들로 구성된 행렬 $\mathbf{H}_s^e$와 상기 행렬 $\mathbf{H}_p$에서 홀수번째 열들로 구성된 행렬 $\mathbf{H}_p^o$를 결합한 행렬이며, 상기 $\mathbf{H}_2=[\mathbf{H}_s^o\ \mathbf{H}_p^e]$은 상기 행렬 $\mathbf{H}_s$에서 홀수번째 열들로 구성된 행렬 $\mathbf{H}_s^o$와 상기 행렬 $\mathbf{H}_p$에서 짝수번째 열들로 구성된 행렬 $\mathbf{H}_p^e$를 결합한 행렬이다. 이렇게, 상기 패리티 검사 행렬 조건 1을 만족하도록 패리티 검사 행렬 H를 설계할 경우에는 상기 송신기의 서로 다른 송신 안테나를 각각을 통해 송신되는 신호들이 선형적 독립(linearly independent) 특성을 가지게 되므로, 상기 송신 안테나들중 어느 한 송신 안테나를 통해 송신되는 신호가 심각한 페이딩(fading) 현상을 겪게 되어도 수신기측에서는 상기 페이딩 현상으로 인한 오류를 극복하는 것이 가능하게 된다.

그러면 여기서 상기 도 5에서 설명한 시공간 LDPC 부호의 부호화 과정에 대해서 정리하면 다음과 같다.

먼저, 상기 패리티 검사 행렬 설계 조건 1을 만족하도록 설계된 패리티 검사 행렬 H가 상기 메모리(502)에 저장되며, 상기 제어기(504)는 상기 메모리(502)에 저장되어 있는 패리티 검사 행렬 H에 상응하게 상기 제 1 LDPC 부호화기(506)와 상기 제 2 LDPC 부호화기(508) 각각이 그 입력되는 정보 데이터를 부호화하도록 제어한다. 그리고, 상기 공간 사상기(510)는 상기 제어기(504)의 제어에 따라 상기 제 1 LDPC 부호화기(506)와 제 2 LDPC 부호화기(508)에서 생성한 LDPC 부호어 c들, 즉 제1 LDPC 부호어 c_1과 제2 LDPC 부호어 c_2를 상기 제 1 LDPC 부호화기(506)와 제 2 LDPC 부호화기(508)에서 출력한 순서에 상응하게 공간 사상하여 상기 제1신호 사상기(512)와 제2신호 사상기(514) 각각으로 출력하는 것이다.

상기 제1신호 사상기(512)와 제2신호 사상기(514) 각각은 상기 공간 사상기(510)에서 출력한 신호를 상기 BPSK 방식으로 신호 사상하고, 상기 제1신호 사상기(512)는 상기 신호 사상한 신호를 상기 제1송신 안테나를 통해 에어(air)상으로 송신하고, 상기 제2신호 사상기(514)는 상기 신호 사상한 신호를 상기 제2송신 안테나를 통해 에어상으로 송신한다.

상기 도 5에서는 신호 사상 방식으로 BPSK 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조에 대해서 설명하였으며, 다음으로 도 6을 참조하여 신호 사상 방식으로 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature AmpRude Modulation) 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조에 대해서 설명하기로 한다.

상기 도 6은 신호 사상 방식으로 QPSK 방식 및 4-QAM 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조를 개략적으로 도시한 도면이다.

상기 도 6을 참조하면, 먼저 상기 송신기는 직렬/병렬 변환기(600)와 메모리(602)와, 제어기(604)와, 실수 파트 시공간 LDPC 부호화기(real part space time LDPC encoder)(606)와, 허수 파트 시공간 LDPC 부호화기(608)와, 다수의, 일 예로 2개의 신호 사상기들, 즉 제1신호 사상기(610)와 제2신호 사상기(612)와, 다수의, 일 예로 상기 2개의 송신 안테나들, 즉 제1송신 안테나와 제2송신 안테나로 구성된다. 여기서, 상기 제1신호 사상기(610)와, 제2신호 사상기(612) 각각은 QPSK 방식 및 4-QAM 방식 중 어느 한 방식으로 신호를 사상한다고 가정하기로 한다.

먼저, 상기 메모리(602)는 본 발명에서 제안하는 시공간 LDPC 부호의 설계 방법에 따른 패리티 검사 행렬을 저장한다. 본 발명에서 제안하는 시공간 LDPC 부호의 설계 방법에 따른 패리티 검사 행렬은 하기에서 구체적으로 설명할 것이므로 여기서는 그 상세한 설명을 생략하기로 한다. 먼저, 정보 데이터 s가 입력되면 상기 직렬/병렬 변환기(600)는 상기 정보 데이터 s를 실수 파트 정보 데이터 s_I과 허수 파트 정보 데이터 s_Q로 병렬 변환하여 출력한다.

상기 실수 파트 정보 데이터 s_I는 상기 실수 파트 시공간 LDPC 부호화기(606)로 입력되고, 상기 실수 파트 시공간 LDPC 부호화기(606)는 상기 제어기(604)의 제어에 따라 상기 실수 파트 정보 데이터 s_I를 미리 설정되어 있는 부호화율에 상응하게 부호화하여 LDPC 부호어, 즉 LDPC 부호화 심벌로 생성한 후 실수 파트 시공간 LDPC 부호화를 수행하여 시공간 LDPC 부호어로 생성한 후 상기 제1신호 사상기(610)와 제 2 신호 사상기(612)로 출력한다. 여기서, 상기 제어기(604)는 상기 메모리(602)에 저장되어 있는 패리티 검사 행렬에 상응하게 상기 실수 파트 시공간 LDPC 부호화기(606)를 제어하여 상기 패리티 검사 행렬에 상응하는 시공간 LDPC 부호어를 생성하도록 제어한다. 일 예로, 상기 실수 파트 정보 데이터 s_I가 k비트이고, 상기 부호화율이 1/2라고 가정하면 상기 실수 파트 시공간 LDPC 부호화기(606)에서 출력하는 비트수 N은 2k가 되며(N = 2k), 결과적으로 상기 실수 파트 시공간 LDPC 부호화기(606)의 출력 데이터는 비트수가 2k인 시공간 LDPC 부호어가 된다. 여기서, 상기 실수 파트 시공간 LDPC 부호화기(606)에서 출력하는 시공간 LDPC 부호어를 C_I라고 표현하기로 하며, 상기 시공간 LDPC 부호어 C_I는 상기 제1송신 안테나를 통해 송신되는 C_I^1과 상기 제2송신 안테나를 통해 송신되는 C_I^2로 구분된다.

또한, 상기 허수 파트 정보 데이터 s_Q는 상기 허수 파트 시공간 LDPC 부호화기(608)로 입력되며, 상기 허수 파트 시공간 LDPC 부호화기(608)는 상기 허수 파트 정보 데이터 s_Q를 상기 제어기(604)의 제어에 따라 미리 설정되어 있는 부호화율에 상응하게 부호화하여 LDPC 부호어, 즉 LDPC 부호화 심벌로 생성한 후 허수 파트 시공간 LDPC 부호화를 수행한 후 상기 제1 공간 사상기(610)와 제 2 공간 사상기(612)로 출력한다. 여기서, 상기 제어기(604)는 상기 메모리(602)에 저장되어 있는 패리티 검사 행렬에 상응하게 상기 허수 파트 시공간 LDPC 부호화기(608)를 제어하여 상기 패리티 검사 행렬에 상응하는 시공간 LDPC 부호어를 생성하도록 제어한다. 일 예로, 상기 허수 파트 정보 데이터 s_Q가 k비트이고, 상기 부호화율이 1/2라고 가정하면 상기 허수 파트 시공간 LDPC 부호화기(608)에서 출력하는 비트수 N은 2k가 되며(N = 2k), 결과적으로 상기 허수 파트 시공간 LDPC 부호화기(608)의 출력 데이터는 비트수가 2k인 시공간 LDPC 부호어가 된다. 여기서, 상기 허수 파트 시공간 LDPC 부호화기(608)에서 출력하는 시공간 LDPC 부호어를 C_Q라고 표현하기로 하며, 상기 시공간 LDPC 부호어 C_Q는 제1송신 안테나를 통해 송신되는 C_Q^1과 제2송신 안테나를 통해 송신되는 C_Q^2로 구분된다.

한편, 상기 메모리(602)에 저장되어 있는 패리티 검사 행렬을 H라고 표현하면, 상기 패리티 검사 행렬 H는 시공간 LDPC 부호를 구성할 때 제1송신 안테나를 통해 송신되는 부호어에 대응되는 파트인 H_1와 제2송신 안테나를 통해 송신되는 부호어에 대응되는 파트인 H_2로 구분되며, 이는 하기 수학식 6과 같이 나타낼 수 있다.

수학식 6

$$H=[H_1 \ H_2]$$

또한, 상기 패리티 검사 행렬 H의 크기는 상기 송신기의 부호화율 및 최종 부호어인 시공간 LDPC 부호의 길이에 상응하는 크기로 결정됨은 물론이다. 한편, 상기 실수 파트 시공간 LDPC 부호화기(606)에서 생성하는 시공간 LDPC 부호어 C_I와 허수 파트 시공간 LDPC 부호화기(608)에서 생성하는 시공간 LDPC 부호어 C_Q는 하기 수학식 7과 같이 나타낼 수 있다.

수학식 7

$$C_I=\begin{bmatrix} c_I^1 \\ \\ c_I^2 \end{bmatrix}, \quad C_Q=\begin{bmatrix} c_Q^1 \\ \\ c_Q^2 \end{bmatrix}$$

여기서, 상기 시공간 LDPC 부호어 C_I과 시공간 LDPC 부호어 C_Q는 상기 패리티 검사 행렬 H와 하기 수학식 8과 같은 관계를 가진다.

수학식 8

$$H \cdot c_I=[H_1 \ H_2] \cdot [c_I^1 \ c_I^2]^T=0$$

$$H \cdot c_Q=[H_1 \ H_2] \cdot [c_Q^1 \ c_Q^2]^T=0$$

한편, 상기 제1신호 사상기(610)는 상기 실수 파트 시공간 LDPC 부호화기(606)에서 출력한 C_I^1와 상기 허수 파트 시공간 LDPC 부호화기(608)에서 출력한 C_Q^1를 입력하여 하기 수학식 9와 같이 신호 사상한 후 제1송신 안테나를 통해 송신한다.

수학식 9

$$\left(\frac{-1}{\sqrt{(2)}}\right) C_I^1+j\left(\frac{-1}{\sqrt{(2)}}\right) C_Q^1$$

또한, 상기 제2신호 사상기(612)는 상기 실수 파트 시공간 LDPC 부호화기(606)에서 출력한 C_I^2와 상기 허수 파트 시공간 LDPC 부호화기(608)에서 출력한 C_Q^2를 입력하여 하기 수학식 10과 같이 신호 사상한 후 제2송신 안테나를 통해 송신한다.

수학식 10

$$\left(\frac{-1}{\sqrt{(2)}}\right) C_I^2+j\left(\frac{-1}{\sqrt{(2)}}\right) C_Q^2$$

이 경우, $2 \times k$ 행렬인 시공간 LDPC 부호어 C는 하기 수학식 11과 같이 나타낼 수 있다.

수학식 11

$$C = \begin{bmatrix} c_I^1 + jc_Q^1 \\ c_I^2 + jc_Q^2 \end{bmatrix}$$

그런데, 상기 제1신호 사상기(610)와, 제2신호 사상기(612) 각각은 상기 QPSK 방식 및 4-QAM 방식중 어느 한 방식으로 신호 사상을 수행하므로 상기 시공간 LDPC 부호어 C는 QPSK 이진 랭크 설계 규칙을 만족해야만 한다. 즉, 상기 종래 기술 부분에서 설명한 바와 같이 모든 0이 아닌 시공간 부호어를 이진 필드 상에서 사상시켰을 때 사상된 시공간 부호어가 이진 필드상에서 최대 랭크를 가지면 그 시공간 부호어가 최대 다이버시티 이득을 가지게 된다. 따라서, 상기 시공간 LDPC 부호어 C가 최대 다이버시티 이득을 가지기 위해서는 상기 이진 필드상에서 최대 랭크를 가져야만 하며, 따라서 상기 패리티 검사 행렬 H는 하기 패리티 검사 행렬 설계 조건 2를 만족해야만 한다.

[패리티 검사 행렬 설계 조건 2]

상기 실수 파트 시공간 LDPC 부호 C_I 및 허수 파트 시공간 LDPC 부호 C_Q가 이진 필드에서 최대 다이버시티 이득을 얻으며 H_1, $H_2 H_1 \oplus H_2$ 가 이진 필드상에서 최대 랭크를 가져야 한다.

상기, 패리티 검사 행렬 설계 조건 2에서 상기 행렬 H_1은 상기 행렬 H에서 제1송신 안테나를 통해 송신되는 부호어에 해당하는 패리티 행렬을 나타내며, 행렬 H_2은 상기 행렬 H에서 제2송신 안테나를 통해 송신되는 부호어에 해당하는 패리티 행렬을 나타낸다.

그러면 여기서 상기 도 6에서 설명한 시공간 LDPC 부호의 부호화 과정에 대해서 설명하면 다음과 같다.

먼저, 상기 패리티 검사 행렬 설계 조건 2를 만족하는 형태로 설계된 패리티 검사 행렬 H가 상기 메모리(602)에 저장되며, 상기 제어기(604)는 상기 메모리(602)에 저장되어 있는 패리티 검사 행렬 H에 상응하게 상기 실수 파트 시공간 LDPC 부호화기(606)와 상기 허수 파트 시공간 LDPC 부호화기(608)가 그 입력되는 정보 데이터를 부호화하도록 제어한다. 상기 제어기(604)의 제어에 따라 상기 실수 파트 시공간 LDPC 부호화기(606)와 허수 파트 시공간 LDPC 부호화기(608) 각각이 생성한 시공간 LDPC 부호어 C_I와 시공간 LDPC 부호어 C_Q는 상기 제1신호 사상기(610)와 제2신호 사상기(612) 각각으로 입력된다.

상기 제1신호 사상기(610)와 제2신호 사상기(612) 각각은 상기 실수 파트 시공간 LDPC 부호기(606) 및 허수 파트 시공간 LDPC 부호기(608)에서 시공간 LDPC 부호어 C_I와 시공간 LDPC 부호어 C_Q를 상기 QPSK 및 4-QAM 방식중 어느 한 방식으로 신호 사상한다. 그리고, 상기 제1신호 사상기(610)는 상기 신호 사상한 신호를 상기 제1송신 안테나를 통해 에어상으로 송신하고, 상기 제2신호 사상기(612)는 상기 신호 사상한 신호를 상기 제2송신 안테나를 통해 에어상으로 송신한다.

상기 도 6에서는 신호 사상 방식으로 QPSK 방식 및 4-QAM 방식을 사용할 경우 본 발명의 실시예에서의 기능을 수행하는 송신기 구조에 대해서 설명하였으며, 다음으로 도 7을 참조하여 본 발명의 실시예에 따른 이동 통신 시스템의 수신기의 동작에 대해서 설명하기로 한다.

상기 도 7은 본 발명의 실시예에 따른 이동 통신 시스템의 수신기 동작 과정을 도시한 흐름도이다.

상기 도 7을 참조하면, 먼저 700단계에서 상기 수신기는 상기 수신기에 구비되어 있는 다수의, 일 예로 2개의 수신 안테나들, 즉 제1수신 안테나 내지 제P수신 안테나들을 통해 상기 수신기에 대응하는 송신기에서 송신한 신호를 수신한 후 702 단계로 진행한다. 상기 702단계에서 상기 수신기는 상기 제1수신 안테나 내지 제P수신 안테나 각각을 통해 수신한 신호들을 기반으로 하여 수신된 신호를 검파한 후 704단계로 진행한다. 여기서, 상기 수신 신호를 검파하는 방식은 다양하게 존재할 수 있음은 물론이며, 상기 도 7에서는 상기 수신 신호 신호를 하기에서 설명할 바와 같은 방식으로 검파한다고 가정하기로 한다.

상기 704단계에서 상기 수신기는 상기 수신 신호를 검파한 값을 기반으로 하여 공간 사상 노드 복호화를 수행하고 706 단계 및 708 단계로 진행한다. 여기서, 상기 공간 사상 노드 복호화는 수신 노드에서 공간 사상 노드로 전달되는 메시지와 제1 LDPC 부호어에 대한 복호화 출력값, 즉 상기 제1LDPC 부호어를 복호화하기 위한 제1LDPC 복호화기(도시하지 않음)의 복호화 출력값과 제2 LDPC 부호어에 대한 복호화 출력값, 즉 상기 제2LDPC 부호어를 복호화하기 위한 제2LDPC 복호화기(도시하지 않음)의 복호화 출력값을 기반으로 하여 상기 공간 사상 노드에서 상기 제1 LDPC 복호화기 및 제2 LDPC 복호화기로 전달되는 메시지(message)를 결정하는 것을 나타낸다. 상기 706 단계에서는 상기 수신기는 상기 제1LDPC 부호어에 대한 복호화 과정, 즉 제1 LDPC 복호화 과정을 수행하며, 상기 708단계에서는 제2LDPC 부호어에 대한 복호화 과정, 즉 제2 LDPC 복호화 과정을 수행한 후 710단계로 진행한다.

상기 710단계에서 상기 수신기는 상기 제1 LDPC 복호화 과정 및 제2 LDPC 복호화 과정을 수행한 결과를 가지고 복호화 중단 조건을 만족하는지를 검사한다. 여기서, 상기 복호화 중단 조건은 미리 설정한 반복 복호화 횟수 혹은 현재까지의 복호화 결과가 상기 패리티 검사 행렬 H를 만족하는지 등이 될 수 있다. 즉, 상기 수신기는 상기 복호화 회수가 상기 반복 복호화 회수를 초과하거나 혹은 현재까지의 복호화 결과가 상기 패리티 검사 행렬 H를 만족할 경우 상기 복호화 동작을 중단하게 된다.

상기 710단계에서 검사 결과 상기 복호화 조건을 만족할 경우 상기 수신기는 712단계로 진행한다. 상기 712단계에서 상기 수신기는 현재까지의 복호화 결과, 즉 제1 LDPC 복호화 과정의 결과값 및 제2 LDPC 복호화 과정의 결과값을 기반으로 하여 복호화값을 계산하고 714단계로 진행한다. 상기 714단계에서 상기 수신기는 상기 계산한 복호화값을 기반으로 하여 정보 데이터를 경판정(hard-decision)한 후 현재까지의 모든 복호화 과정을 종료한다.

만약, 상기 710단계에서 검사 결과 상기 복호화 조건을 만족하지 않을 경우 상기 수신기는 704단계로 되돌아간다. 즉, 상기 시공간 LDPC 부호는 초기 수신된 수신 신호를 기반으로 하여 반복 복호화를 수행함으로써 그 복호화 성능이 향상된다.

상기 도 7에서는 본 발명의 실시예에 따른 이동 통신 시스템의 수신기 동작에 대해서 설명하였으며, 다음으로 도 8을 참조하여 본 발명의 실시예에서의 기능을 수행하는 이동 통신 시스템의 수신기 구조에 대해서 설명하기로 한다.

상기 도 8은 본 발명의 실시예에서의 기능을 수행하는 이동 통신 시스템의 수신기 구조를 개략적으로 도시한 도면이다.

상기 도 8을 참조하면, 먼저 상기 수신기는 다수개의, 일 예로 2개의 수신 안테나들, 즉 제1수신 안테나와 제2수신 안테나와, 검파기(detector)(800)와, 공간 사상 노드 복호화기(spatial node decoder)(802)와, 제어기(804)와, 메모리(806)와, 제1 LDPC 복호기(LDPC decoder)(808), 제 2 LDPC 복호기(810), 경판정기(hard decision unit)(812)로 구성된다.

먼저, 상기 제1수신 안테나와 제2수신 안테나 각각은 상기 수신기에 대응하는 송신기에서 송신한 신호를 수신하여 상기 검파기(800)로 출력한다. 상기 검파기(800)는 상기 제1수신 안테나와 제2수신 안테나 각각에서 출력한 수신 신호를 입력하여 상기 송신기의 송신 안테나들, 즉 제1송신 안테나 내지 제2송신 안테나 각각에서 송신한 송신 신호를 검파한 후 상기 공간 노드 복호화기(802)로 출력한다. 상기 공간 노드 복호화기(802)는 상기 검파기(800)에서 출력한 신호를 입력하여 상기 제1송신 안테나 및 제2송신 안테나 각각에서 송신한 신호에 대해 상기 검파기(800)에서 검파한 추정치를 공간 역사상하여 상기 제1 LDPC 복호기(808) 및 제2 LDPC 복호기(810)로 출력한다. 여기서, 상기 공간 사상 노드 복호화기(802)로 입력되는 신호는 최초에는 상기에서 설명한 바와 같이 상기 검파기(800)에서 출력한 신호뿐이지만, 이후 반복 복호화를 수행함에 있어서는 상기 검파기(800)에서 출력한 신호뿐만 아니라 상기 제1 LDPC 복호기(808) 및 제2 LDPC 복호기(810)에서 출력한 신호까지 포함된다.

상기 제1 LDPC 복호기(808)는 상기 공간 사상 노드 복호화기(802)에서 출력한 신호를 입력하여 상기 송신기의 상기 시공간 LDPC 부호화기에서 설계한 패리티 검사 행렬을 기반으로 하는 제1 LDPC 부호화기에 상응하도록 LDPC 복호화를 수행한다. 여기서, 상기 송신기가 신호 사상 방식으로 상기 도 6에서 설명한 바와 같이 QPSK 방식을 사용할 경우 상기 제

제 1 LDPC 복호화기(808)는 실수 파트 시공간 LDPC 부호화기(606)의 부호화 과정에 대응하도록 복호화 동작을 수행하게 된다. 즉, 상기 제어기(804)는 상기 메모리(806)에 저장되어 있는, 상기 최대 다이버시티를 얻을 수 있도록 설계한 패리티 검사 행렬에 상응하게 상기 제 1 LDPC 복호화기(808)에서 복호화 과정을 수행하도록 제어하는 것이다.

또한, 상기 제 2 LDPC 복호화기(810)는 상기 공간 사상 노드 복호화기(802)에서 출력한 신호를 입력하여 상기 송신기의 상기 시공간 LDPC 부호화기에서 설계한 패리티 검사 행렬을 기반으로 하는 제 2 LDPC 부호화기에 상응하도록 LDPC 복호화를 수행한다. 여기서, 상기 송신기가 신호 사상 방식으로 상기 도 6에서 설명한 바와 같이 QPSK 방식을 사용할 경우 상기 제 1 LDPC 복호화기(810)는 허수 파트 시공간 LDPC 부호화기(608)의 부호화 과정에 대응하도록 복호화 동작을 수행하게 된다. 즉, 상기 제어기(804)는 상기 메모리(806)에 저장되어 있는, 상기 최대 다이버시티를 얻을 수 있도록 설계한 패리티 검사 행렬에 상응하게 상기 제 1 LDPC 복호화기(810)에서 복호화 과정을 수행하도록 제어하는 것이다.

그러면 여기서 상기 검파기(800) 및 공간 사상 노드 복호화기(802)의 동작을 구체적으로 설명하기로 한다.

먼저, 임의의 시점 t에서 송신되는 송신 신호를 2개의 엘리먼트들로 구성된 송신 신호 벡터(vector) x_t로 표현하기로 하며, 상기 임의의 시점 t에서 수신되는 수신 신호를 1개의 엘리먼트로 구성된 수신 신호 벡터 y_t로 표현하기로 한다. 여기서, 상기 송신 신호 벡터 x_t는 2개의 송신 안테나들, 즉 제1송신 안테나와 제2송신 안테나의 2개의 송신 안테나들을 통해 송신되므로 2개의 엘리먼트들로 구성되는 것이다. 또한, 상기 신호가 송수신단에서 겪는 채널은 페이딩 채널(fading channel)이며, 상기 페이딩 채널을 1×2 인 행렬 Ω_t로 표현하기로 하며, 잡음(noise) 성분을 1개의 엘리먼트로 구성된 잡음 벡터 n_t라고 표현하기로 한다. 그러면 하기 수학식 12와 같은 관계가 성립한다.

수학식 12

$$y_t = \Omega_t x_t + n_t$$

상기 수학식 12에서 상기 송신 신호 벡터 x_t는 임의의 시점 t에서 상기 제1송신 안테나와 제2송신 안테나 각각을 통해 송신되는 송신 신호를 고려하여 $x_t = (x_t^1, x_t^2)$ 와 같이 표현할 수 있다. 여기서, 상기 x_t^1는 상기 임의의 시점 t에서 상기 제1송신 안테나를 통해서 송신되는 신호이며, 상기 x_t^2는 상기 임의의 시점 t에서 상기 제2송신 안테나를 통해서 송신되는 신호이다. 그러므로, 상기 송신 신호 벡터 $x_t = (x_t^1, x_t^2)$에 대해 상기 수신 신호 벡터 y_t에 대한 확률(probability) 벡터 P_t는 하기 수학식 13과 같이 나타낼 수 있다. 여기서, 설명의 편의상 송신기가 적용하는 신호 사상 방식을 BPSK 방식이라고 가정하기로 하며, 상기 신호 사상 방식으로 BPSK 방식 뿐만 아니라 QPSK 방식이 사용될 수 있음은 물론이다.

수학식 13

$$P_t = (P_t(x_t^1 = 1, x_t^2 = 1), P_t(x_t^1 = -1, x_t^2 = 1), P_t(x_t^1 = 1, x_t^2 = -1), P_t(x_t^1 = -1, x_t^2 = -1))$$

상기 수학식 13에 나타낸 바와 같이 상기 송신 신호 벡터 $x_t = (x_t^1, x_t^2)$에 대해 상기 수신 신호 벡터 y_t에 대한 확률 벡터 P_t는 4개의 엘리먼트들로 구성된다. 여기서, 상기 송신기가 그 신호 사상 방식으로 상기 BPSK 방식을 적용하였으므로, 상기 확률 벡터 P_t는 4개의 엘리먼트들로 구성되는 것이며, 결과적으로 상기 수신 신호 벡터 y_t가 상기 검파기(800)로부터 검출되는 메시지가 되는 것이다.

또한, 상기 확률 벡터 P_t의 4개의 엘리먼트들 각각을 구하기 위해 $x_t = (x_t^1, x_t^2) = (i,j), (i,j \in -1,1)$인 경우 하기 수학식 14와 같이 나타낼 수 있다.

수학식 14

$$P_t(x_t^1=i,x_t^2=j)=Pr(x_t^1=i,x_t^2=j|y_t)$$

상기 수학식 14에서 Pr은 확률을 구하는 연산자이다. 상기에서 설명한 바와 같이 상기 검파기(800)는 상기 확률 벡터 P_t를 검출하고, 상기 검출한 확률 벡터 P_t를 상기 공간 사상 노드 복호기(802)로 출력한다.

한편, 상기 공간 사상 노드 복호기(802)에서 제 1 LDPC 복호화기(808)로 출력되는, 상기 제1송신 안테나에서 송신되는 LLR(log likelihood ratio) 메시지는 하기 수학식 15와 같이 나타낼 수 있다.

수학식 15

$$L_t^1=\log\frac{Pr(x_t^1=+1|y_t)}{Pr(x_j^1=-1|y_t)}=\log\frac{\sum_{x_t^2\in\{-1,1\}}Pr(x_t^1=+1,x_t^2|y_t)Pr(x_t^2|y_t)}{\sum_{x_j^2\in\{-1,1\}}Pr(x_j^1=-1,x_t^2|y_t)Pr(x_t^2|y_t)}$$

상기 수학식 15에서 $Pr(x_k^2|y_k)$는 상기 제 1 LDPC 복호화기(808) 및 제 2 LDPC 복호기(810)의 출력 값으로부터 검출할 수 있다. 또한, 상기 제1송신 안테나에서 송신되는 LLR 메시지를 검출하는 방식과 동일한 방식으로 상기 제2송신 안테나에서 송신되는 LLR 메시지를 검출할 수 있다. 상기 제 2 송신 안테나에서 송신 되는 LLR 메시지는 상기 제 2 LDPC 복호화기(810)로 입력된다.

상기에서는 송신기가 신호 사상 방식으로 BPSK 방식을 적용한 경우의 상기 검파기(800) 및 공간 사상 노드 복호화기(802)의 동작에 대해서 설명하였으며, 다음으로 상기 송신기가 신호 사상 방식으로 QPSK 방식을 적용한 경우의 상기 검파기(800) 및 공간 사상 노드 복호화기(802)의 동작에 대해서 설명하기로 한다.

먼저, 상기 송신기가 송신하는 송신 신호 벡터 $x_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)$에 대해 상기 수신기가 수신하는 수신 신호 벡터 y_t에 대한 확률 벡터 P_t는 하기 수학식 16과 같이 나타낼 수 있다.

수학식 16

$$P_t=(P_t((x_t^1)^I=1,(x_t^1)^Q=1,(x_t^2)^I=1,(x_t^2)^Q=1),\cdots,P_t((x_t^1)^I=-1,(x_t^1)^Q=-1,(x_t^2)^I=-1,(x_t^2)^Q=-1))$$

상기 수학식 16에 나타낸 바와 같이 상기 송신 신호 벡터 $x_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)$에 대해 상기 수신 신호 벡터 y_t에 대한 확률 벡터 P_t는 16개의 엘리먼트들로 구성된다. 여기서, 상기 송신기가 그 신호 사상 방식으로 상기 QPSK 방식을 적용하였으므로, 상기 확률 벡터 P_t는 16개의 엘리먼트들로 구성되는 것이며, 결과적으로 상기 수신 신호 벡터 y_t가 상기 검파기(800)로부터 검출되는 메시지가 되는 것이다.

여기서, 상기 확률 벡터 P_t의 16개의 엘리먼트들 각각을 구하기 위해 $\mathbf{x}_t=((x_t^1)^I,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q)=(i,j,k,l),(i,j,k,l\in-1,1)$인 경우 하기 수학식 17과 같이 나타낼 수 있다.

수학식 17

$$P_t((x_t^1)^I=i,(x_t^1)^Q=j,(x_t^2)^I=k,(x_t^2)^Q=l)$$

$$=\Pr((x_t^1)^I=i,(x_t^1)^Q=j,(x_t^2)^I=k,(x_t^2)^Q=l\,|\,y_t)$$

상기에서 설명한 바와 같이 상기 결합기(800)는 상기 확률 벡터 P_t를 검출하고, 상기 검출한 확률 벡터 P_t를 상기 공간 사상 노드 복호기(802)로 출력한다. 상기 공간 사상 노드 복호기(802)에서 제 1 LDPC 복호화기(808)로 출력하는, 제1송신 안테나에서 송신되는 실수 파트에 해당하는 LLR 메시지는 하기 수학식 18과 같이 나타낼 수 있다.

수학식 18

$$(L_t^1)^I=\log\frac{\Pr((x_t^1)^I=+1|y_t)}{\Pr((x_t^1)^I=-1|y_t)}$$
$$=\log\frac{\sum_{(x)^Q,(x)^I,(x)^Q=(-1,1)}\Pr((x_t^1)^I=+1,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)}{\sum_{(x)^I,(x)^Q,(x)^I,(x)^Q=(-1,1)}\Pr((x_t^1)^I=-1,(x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_t)}$$

상기 수학식 18에서 $\Pr((x_t^1)^Q,(x_t^2)^I,(x_t^2)^Q|y_k)$는 상기 제 1 LDPC 복호화기(808) 및 제 2 LDPC 복호화기(810)의 출력 값으로부터 검출할 수 있다. 또한, 상기 제1송신 안테나에서 송신되는 실수 파트에 해당하는 LLR 메시지를 검출하는 방식과 동일한 방식으로 상기 제1송신 안테나에서 송신되는 허수 파트에 해당하는 LLR 메시지와, 상기 제2송신 안테나에서 송신되는 실수 파트에 해당하는 LLR 메시지 및 상기 제2송신 안테나에서 송신되는 허수 파트 LLR 메시지를 검출할 수 있다. 여기서, 상기 제 2 송신 안테나에서 송신되는 LLR 메시지는 상기 제 2 LDPC 복호화기(810)로 입력된다.

상기 제 1 LDPC 복호화기(808) 및 제 2 LDPC 복호기(810)는 상기 공간 사상 노드 복호화기(802)의 출력 값을 기반으로 하여 제 1 LDPC 부호어 및 제 2 LDPC 부호어를 복호화하며, 그 복호화된 신호들은 다시 상기 공간 사상 노드 복호기(802)와 정판정기(812)로 입력된다. 상기 제 1 LDPC 복호화기(808) 및 상기 제 2 LDPC 복호화기(810)의 출력 값을 상기 공간 사상 노드 복호화기(802)에 다시 입력되도록 해서함으로써 초기에 추정한 신호의 신뢰도를 증가시킬 수 있는 것이다. 상기 정판정기(812)는 상기 제 1 LDPC 복호화기(808) 및 제 2 LDPC 복호화기(810)에서 출력한 신호를 입력하여 정판정(hard decision)하여 최종 정보 데이터로 복원한다.

한편 본 발명의 상세한 설명에서는 구체적인 실시예에 관해 설명하였으나, 본 발명의 범위에서 벗어나지 않는 한도내에서 여러 가지 변형이 가능함은 물론이다. 그러므로 본 발명의 범위는 설명된 실시예에 국한되어 정해져서는 안되며 후술하는 특허청구의 범위뿐만 아니라 이 특허청구의 범위와 균등한 것들에 의해 정해져야 한다.

발명의 효과

상술한 바와 같은 본 발명은, 최대 다이버시티 이득을 가지는 시공간 LDPC 부호를 부호화/복호화할 수 있는 패리티 검사 행렬을 생성해 제안함으로써 여러 정정 능력을 최대화하여 시스템 전체 성능을 극대화시킨다는 이점을 가진다. 또한, 시공간 부호의 부호화 방식으로 상기 LDPC 방식을 사용함으로써 비교적 간단한 하드웨어 구조로 우수한 성능을 보장하는 시공간 부호를 부호화/복호화할 수 있다는 이점을 가진다.

(57) 청구의 범위

청구항 1.

단수개의 충진 안테나를을 사용하는 송신기에, 다수의 수신 안테나를을 사용하는 수신기를 포함하는 이동 통신 시스템에서 사전신 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 패리티 검사 행렬을 생성하는 방법에 있어서,

정보 데이터를 사전인 LDPC 부호로 부호화시 적용되는 부호화율과 상기 사전인 LDPC 부호의 부호의 길이에 대응하도록 패리티 검사 행렬의 크기를 결정하는 과정과,

상기 결정된 크기의 패리티 검사 행렬을 상기 길로 데이터에 대응되는 제1부분 행렬과 패리티에 대응되는 제2부분 행렬로 분할하는 과정과,

상기 제1부분 행렬의 복수개(ovon) 열들로 구성된 제3부분 행렬과, 상기 제2부분 행렬의 출수열개(odd) 열들로 구성된 제4 부분행렬과, 상기 제3 부분행렬과 제4 부분행렬을 결합한 행렬인 제5부분 행렬과, 상기 제1부분 행렬의 출수열에 상기 제7 부분 구성된 제0부분 행렬과, 상기 제2부분 행렬인 제8부분 열들로 구성된 제7부분 부분 행렬을 결합한 행렬인 제8부분 행렬과, 상기 제1부분 행렬과 상기 제2부분 행렬을 배타적 논리합한 제9부분 행렬을 생성하는 과정과,

상기 제5 부분 행렬 및 제8 부분 행렬이 상기 제0부분 행렬과 이전 패드상에서 비어 있경된 행크를 꺼적드록 상기 패턴의 검사 행렬을 생성하는 과정을 프랍합을 특정으로 하는 상기 방법.

청구항 2.

제1항에 있어서,

상기 송정 행크는 최대 행크(full rank)임을 특징으로 하는 상기 방법.

청구항 3.

제1항에 있어서,

상기 송신 안테나를의 개수는 2개이며, 상기 사전인 LDPC 부호는 BPSK(Binary Phase Shift Keying) 방식을 사용하여 신호 사상됨을 특징으로 하는 상기 방법.

청구항 4.

낮수의 송신 안테나를을 사용하는 송신기와, 다수의 수신 안테나를을 사용하는 수신기를 포함하는 이동 통신 시스템에서 사전인 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 패리티 검사 행렬을 생성하는 방법에 있어서,

정보 데이터를 사전인 LDPC 부호로 부호화시 적용되는 부호화율과 상기 사전인 LDPC 부호의 부호의 길이에 대응하도록 패리티 검사 행렬의 크기를 결정하는 과정과,

상기 사전인 LDPC 부호를 길수 씌는 사전인 LDPC 부호의 행수 차밀 사전인 LDPC 부호로 분할하고, 상기 길수 씌는 사전인 LDPC 부호를 제1송신 안테나를 통해 송신되는 제1부분 행렬과 상기 제1부분 행렬을 제외한 제2부분 행렬로 분할하고, 상기 하수 개의 사전 사전인 LDPC 부호를 상기 제1송신 안테나를 통해 송신되는 제3부분과 상기 제2송신 안테나를을 통해 송신되는 제4부분에로 분할하는 과정과,

상기 패리티 검사 행렬을 상기 제1부분과 제3부분에 대응되는 제1부분 행렬과 상기 제2부분과 제4부분에 대응되는 제2부분 행렬로 분할하는 과정과,

상기 제1부분 행렬과, 상기 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 배타적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지도록 상기 패리티 검사 행렬을 생성하는 과정을 포함함을 특징으로 하는 상기 방법.

청구항 5.

제4항에 있어서,

상기 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호는 이진 필드상에서 미리 설정한 랭크를 가지도록 상기 패리티 검사 행렬을 생성하는 과정을 포함함을 특징으로 하는 상기 방법.

청구항 6.

제4항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 방법.

청구항 7.

제6항에 있어서,

상기 실수 파트 시공간 LDPC 부호는 상기 정보 데이터의 실수 파트 정보 데이터에 대응되는 부호어이며, 상기 허수 파트 시공간 LDPC 부호는 상기 정보 데이터의 허수 파트 정보 데이터에 대응되는 부호어임을 특징으로 하는 상기 방법.

청구항 8.

제7항에 있어서,

상기 송신 안테나들의 개수는 2개이며, 상기 시공간 LDPC 부호는 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature Amplitude Modulation) 방식 중 어느 한 방식을 사용하여 신호 사상됨을 특징으로 하는 상기 방법.

청구항 9.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 방법에 있어서,

정보 데이터를 입력하는 과정과,

상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 과정과,

상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 과정을 포함함을 특징으로 하는 상기 방법.

청구항 10.

제9항에 있어서,

상기 설정 행코는 최대 랭크(full rank)임을 특징으로 하는 상기 방법.

청구항 11.

제10항에 있어서,

상기 송신 안테나들의 개수는 2개이며, 상기 시공간 LDPC 부호를 BPSK(Binary Phase Shift Keying) 방식을 사용하여 신호 사상하는 과정을 더 포함함을 특징으로 하는 상기 방법.

청구항 12.

제11항에 있어서,

상기 정보 데이터를 상기 시공간 LDPC 부호로 생성하는 과정은;

상기 정보 데이터를 제1정보 데이터와 제2정보 데이터로 병렬 변환하는 과정과,

상기 제1정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 상기 정보 데이터의 제1패리티인 제1부호어를 생성하는 과정과,

상기 제2정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 상기 정보 데이터의 제2패리티인 제2부호어를 생성하는 과정과,

상기 제1정보데이터를 구성하는 짝수번째 비트들과, 상기 제2정보 데이터를 구성하는 짝수번째 비트들과, 상기 제1부호어를 구성하는 홀수번째 비트들과, 상기 제2부호어를 구성하는 홀수번째 비트들이 상기 2개의 송신 안테나들중 어느 한 송신 안테나를 통해 송신되고, 상기 제2부호어를 구성하는 짝수번째 비트들과, 상기 제1부호어를 구성하는 짝수번째 비트들과, 상기 제2정보데이터를 구성하는 홀수번째 비트들과, 상기 제1정보데이터를 구성하는 홀수번째 비트들을 상기 송신 안테나들중 나머지 한 송신 안테나를 통해 송신되도록 공간 사상하여 상기 시공간 LDPC 부호로 생성하는 과정을 포함함을 특징으로 하는 상기 방법.

청구항 13.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 방법에 있어서,

정보 데이터를 입력하는 과정과,

제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 대각적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 과정과,

상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 과정을 포함하며,

상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기

다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할하고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 하는 상기 방법.

청구항 14.

제13항에 있어서,

상기 시공간 LDPC 부호를 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature Amplitude Modulation) 방식 중 어느 한 방식을 사용하여 신호 사상하는 과정을 더 포함함을 특징으로 하는 상기 방법.

청구항 15.

제14항에 있어서,

상기 정보 데이터를 상기 시공간 LDPC 부호로 생성하는 과정은;

상기 정보 데이터를 실수 파트 정보 데이터와 허수 파트 정보 데이터로 병렬 변환하는 과정과,

상기 실수 파트 정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 제1부호어를 생성하는 과정과,

상기 허수 파트 정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 제2부호어를 생성하는 과정과,

상기 제1부호어를 구성하는 비트들중 미리 설정된 비트들과 상기 제2부호어를 구성하는 비트들중 미리 설정된 비트들을 상기 제1송신 안테나를 통해 송신되고, 상기 제1부호어를 구성하는 비트들중 상기 설정 비트들을 제외한 나머지 비트들과 상기 제2부호어를 구성하는 비트들중 상기 설정 비트들을 제외한 나머지 비트들을 상기 제2송신 안테나를 통해 송신되도록 공간 사상하여 상기 시공간 LDPC 부호로 생성하는 과정을 포함함을 특징으로 하는 상기 방법.

청구항 16.

제15항에 있어서,

상기 패리티 검사 행렬은 상기 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호가 이진 필드상에서 미리 설정한 랭크를 가지도록 생성됨을 특징으로 하는 상기 방법.

청구항 17.

제16항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 방법.

청구항 18.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 장치에 있어서,

정보 데이터가 입력되면, 소정 제어에 따라 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하는 LDPC 부호화부와,

상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 공간 사상기와,

상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 상기 LDPC 부호로 생성하도록 제어하는 제어기를 포함함을 특징으로 하는 상기 장치.

청구항 19.

제18항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 장치.

청구항 20.

제19항에 있어서,

상기 송신 안테나들의 개수는 2개이며, 상기 시공간 LDPC 부호를 BPSK(Binary Phase Shift Keying) 방식을 사용하여 신호 사상하는 신호 사상부를 더 포함함을 특징으로 하는 상기 장치.

청구항 21.

제20항에 있어서,

상기 장치는 상기 정보 데이터를 제1정보 데이터와 제2정보 데이터로 병렬 변환하는 직렬/병렬 변환기를 더 포함함을 특징으로 하는 상기 장치.

청구항 22.

제21항에 있어서,

상기 LDPC 부호화부는:

상기 제1정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 상기 정보 데이터의 제1패리티인 제1부호어를 생성하는 제1LDPC 부호화기와,

상기 제2정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 상기 정보 데이터의 제2패리티인 제2부호어를 생성하는 제2LDPC 부호화기를 포함함을 특징으로 하는 상기 장치.

청구항 23.

제22항에 있어서,

상기 공간 사상기는 상기 제1정보 데이터를 구성하는 짝수번째 비트들과, 상기 제2정보 데이터를 구성하는 짝수번째 비트들과, 상기 제1부호어를 구성하는 홀수번째 비트들과, 상기 제2부호어를 구성하는 홀수번째 비트들이 상기 2개의 송신 안테나들중 어느 한 송신 안테나를 통해 송신되고, 상기 제2부호어를 구성하는 짝수번째 비트들과, 상기 제1부호어를 구성하는 짝수번째 비트들과, 상기 제2정보 데이터를 구성하는 홀수번째 비트들과, 상기 제1정보 데이터를 구성하는 홀수번째 비트들을 상기 송신 안테나들중 나머지 한 송신 안테나를 통해 송신되도록 공간 사상함을 특징으로 하는 상기 장치.

청구항 24.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 부호화 장치에 있어서,

정보 데이터가 입력되면, 소정 제어에 따라 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하고, 상기 LDPC 부호를 미리 설정되어 있는 공간 사상 방식에 상응하게 공간 사상하여 시공간 LDPC 부호로 생성하는 시공간 LDPC 부호화부와,

제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 대각적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지는 패리티 검사 행렬에 상응하게 상기 정보 데이터를 부호화하여 LDPC 부호로 생성하도록 제어하는 제어기를 포함하며,

상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기 다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할되고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 하는 상기 장치.

청구항 25.

제24항에 있어서,

상기 장치는 상기 시공간 LDPC 부호를 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature Amplitude Modulation) 방식 중 어느 한 방식을 사용하여 신호 사상하는 신호 사상부를 더 포함함을 특징으로 하는 상기 장치.

청구항 26.

제25항에 있어서,

상기 장치는 상기 정보 데이터를 실수 파트 정보 데이터와 허수 파트 정보 데이터로 병렬 변환하는 직렬/병렬 변환기를 더 포함함을 특징으로 하는 상기 장치.

청구항 27.

제26항에 있어서,

상기 시공간 LDPC 부호화부는:

상기 실수 파트 정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 제1부호어를 생성하는 실수 파트 시공간 LDPC 부호화기와,

상기 허수 파트 정보 데이터를 상기 패리티 검사 행렬에 상응하게 부호화하여 제2부호어를 생성하는 허수 파트 시공간 LDPC 부호화기를 포함하며,

상기 실수 파트 시공간 LDPC 부호화기와 허수 파트 시공간 LDPC 부호화기는 상기 제1부호어를 구성하는 비트들중 미리 설정된 비트들과 상기 제2부호어를 구성하는 비트들중 미리 설정된 비트들을 상기 제1송신 안테나를 통해 송신되고, 상기 제1부호어를 구성하는 비트들중 상기 설정 비트들을 제외한 나머지 비트들과 상기 제2부호어를 구성하는 비트들중 상기 설정 비트들을 제외한 나머지 비트들을 상기 제2송신 안테나를 통해 송신되도록 공간 사상하여 상기 시공간 LDPC 부호로 생성함을 특징으로 하는 상기 장치.

청구항 28.

제27항에 있어서,

상기 패리티 검사 행렬은 상기 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호가 이진 필드상에서 미리 설정한 랭크를 가지도록 생성됨을 특징으로 하는 상기 장치.

청구항 29.

제28항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 장치.

청구항 30.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 복호화 방법에 있어서,

상기 다수의 수신 안테나를 각각을 통해 수신되는 신호들을 수신 신호 검파하는 제1과정과,

상기 수신 신호들의 검파값을 기반으로 하여 공간 사상 노드 복호화를 수행하는 제2과정과,

상기 공간 사상 노드 복호화된 신호를 미리 설정되어 있는 패리티 검사 행렬에 상응하게 복호화하는 제3과정과,

상기 패리티 검사 행렬에 상응하게 복호화한 신호의 확률값들이 미리 미리 설정한 반복 복호화 중단 조건을 만족하면 상기 확률값들을 경판정하여 정보 데이터로 출력하고, 상기 확률값들이 상기 반복 복호화 중단 조건을 만족하지 않으면 상기 반복 복호화 중단 조건을 만족할 때까지 상기 제2과정 내지 제3과정의 동작을 반복하여 수행하도록 제어하는 제4과정을 포함함을 특징으로 하는 상기 방법.

청구항 31.

제30항에 있어서,

상기 패리티 검사 행렬은 상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정한 랭크를 가짐을 특징으로 하는 상기 방법.

청구항 32.

제31항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 방법.

청구항 33.

제32항에 있어서,

상기 송신 안테나들의 개수는 2개이며, 상기 시공간 LDPC 부호는 BPSK(Binary Phase Shift Keying) 방식으로 신호 사상됨을 특징으로 하는 상기 방법.

청구항 34.

제30항에 있어서,

상기 패리티 검사 행렬은 제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 배타적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지며,

상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기 다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할되고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 하는 상기 방법.

청구항 35.

제34항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 방법.

청구항 36.

제35항에 있어서,

상기 시공간 LDPC 부호는 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature Amplitude Modulation) 방식 중 어느 한 방식으로 신호 사상됨을 특징으로 하는 상기 방법.

청구항 37.

다수의 송신 안테나들을 사용하는 송신기와, 다수의 수신 안테나들을 사용하는 수신기를 포함하는 이동 통신 시스템에서 시공간 저밀도 패리티 검사(LDPC: Low Density Parity Check) 부호의 복호화 장치에 있어서,

상기 다수의 수신 안테나들 각각을 통해 수신되는 신호들을 수신 신호 검파하는 검파기와,

상기 수신 신호들의 검파값을 기반으로 하여 공간 사상 노드 복호화를 수행하는 공간 사상 노드 복호화기와,

상기 공간 사상 노드 복호화된 신호를 미리 설정되어 있는 패리티 검사 행렬에 상응하게 복호화하는 LDPC 부호화부와,

상기 패리티 검사 행렬에 상응하게 복호화한 신호의 확률값들이 미리 미리 설정한 반복 복호화 중단 조건을 만족하면 상기 확률값들을 경판정하여 정보 데이터로 출력하고, 상기 확률값들이 상기 반복 복호화 중단 조건을 만족하지 않으면 상기 반복 복호화 중단 조건을 만족할 때까지 상기 제2과정 내지 제3과정의 동작을 반복하여 수행하도록 제어하는 제어기를 포함함을 특징으로 하는 상기 장치.

청구항 38.

제37항에 있어서,

상기 패리티 검사 행렬은 상기 정보 데이터에 대응되는 제1부분 행렬의 짝수번째(even) 열들로 구성된 제2부분 행렬과 패리티에 대응되는 제3부분 행렬의 홀수번째 열들로 구성된 제4부분 행렬을 결합한 행렬인 제5부분 행렬 및 상기 제1부분 행렬의 홀수번째 열들로 구성된 제6부분 행렬과 상기 제3부분 행렬의 짝수번째 열들로 구성된 제7부분 행렬을 결합한 행렬인 제8부분 행렬이 상기 제1부분 행렬과 상기 제3부분 행렬을 배타적 논리합한 제9부분 행렬과 이진 필드상에서 미리 설정한 랭크를 가짐을 특징으로 하는 상기 장치.

청구항 39.

제38항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 장치.

청구항 40.

제39항에 있어서,

상기 송신 안테나들의 개수는 2개이며, 상기 시공간 LDPC 부호는 BPSK(Binary Phase Shift Keying) 방식으로 신호 사상됨을 특징으로 하는 상기 장치.

청구항 41.

제37항에 있어서,

상기 패리티 검사 행렬은 제1부호어와 제3부호어에 대응되는 제1부분 행렬과, 제2부호어와 제4부호어에 대응되는 제2부분 행렬 및 상기 제1부분 행렬과 제2부분 행렬을 배타적 논리합한 제3부분 행렬이 이진 필드상에서 미리 설정한 랭크를 가지며,

상기 시공간 LDPC 부호는 실수 파트 시공간 LDPC 부호와 허수 파트 시공간 LDPC 부호로 분할되고, 상기 실수 파트 시공간 LDPC 부호는 상기 다수의 송신 안테나들중 임의의 송신 안테나인 제1송신 안테나를 통해 송신되는 제1부호어와 상기 다수의 송신 안테나들중 임의의 제2송신 안테나를 통해 송신되는 제2부호어로 분할되고, 상기 허수 파트 시공간 LDPC 부호는 상기 제1송신 안테나를 통해 송신되는 제3부호어와 상기 제2송신 안테나를 통해 송신되는 제4부호어로 분할됨을 특징으로 하는 상기 장치.

청구항 42.

제41항에 있어서,

상기 설정 랭크는 최대 랭크(full rank)임을 특징으로 하는 상기 장치.

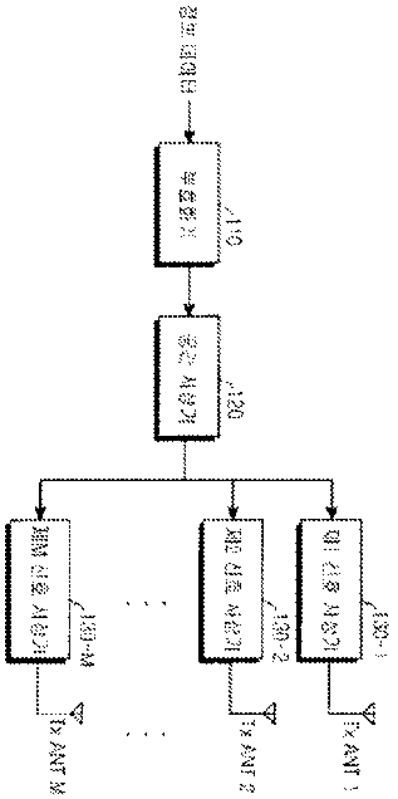
청구항 43.

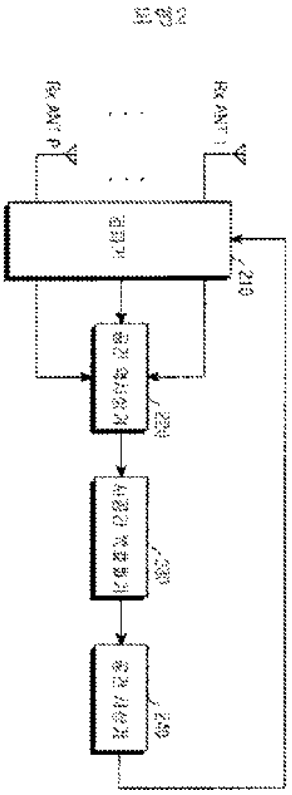
제42항에 있어서,

상기 사용한 LDPC 부호는 QPSK(Quadrature Phase Shift Keying) 방식 및 4-QAM(Quadrature Amplitude Modulation) 방식 중 어느 한 방식으로 신호 사상됨을 특징으로 하는 상기 장치.

도면

도면1

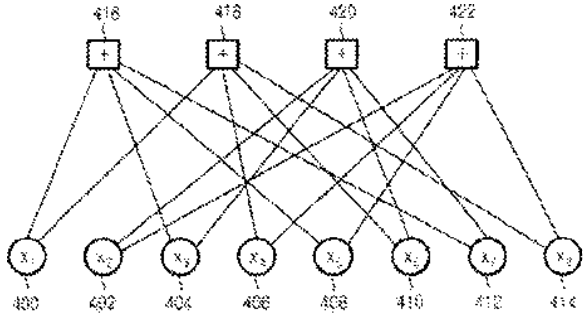




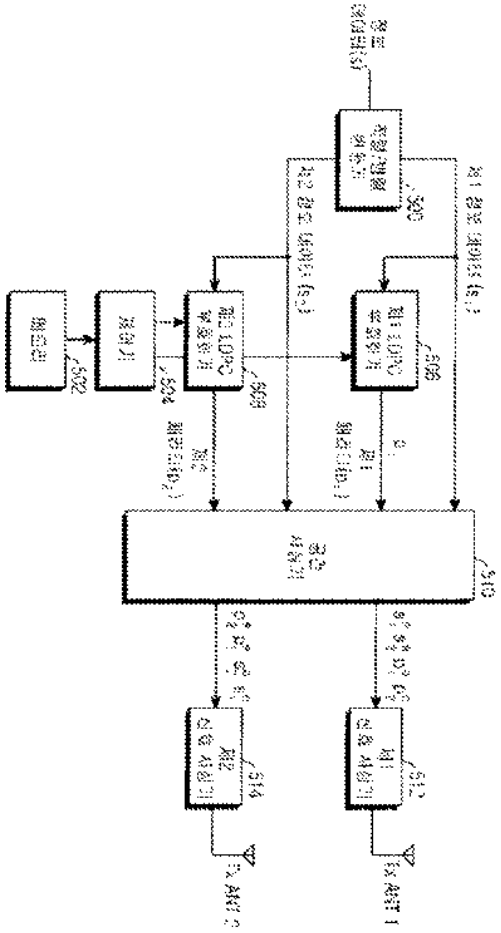
도면2

$$H = \begin{bmatrix} 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 \\ 1 & 0 & 0 & 1 & 0 & 1 & 0 & 1 \\ 0 & 1 & 1 & 0 & 0 & 1 & 1 & 0 \\ 0 & 1 & 0 & 1 & 1 & 0 & 0 & 1 \end{bmatrix}$$

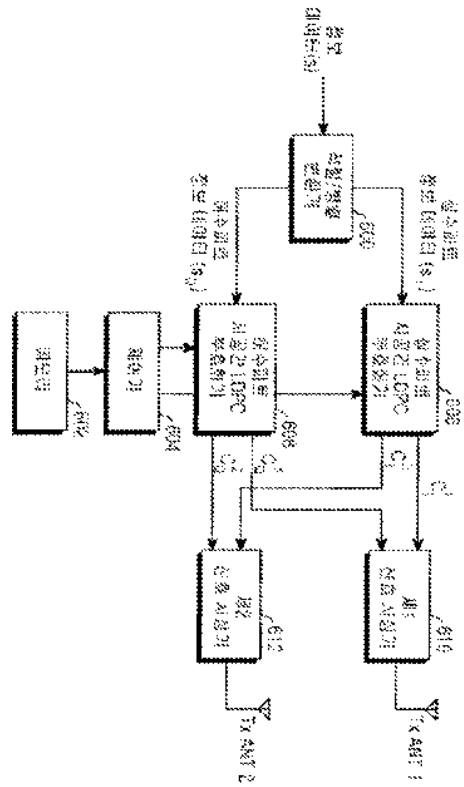
도면4


416
418
420
422
400
402
404
406
408
410
412
414

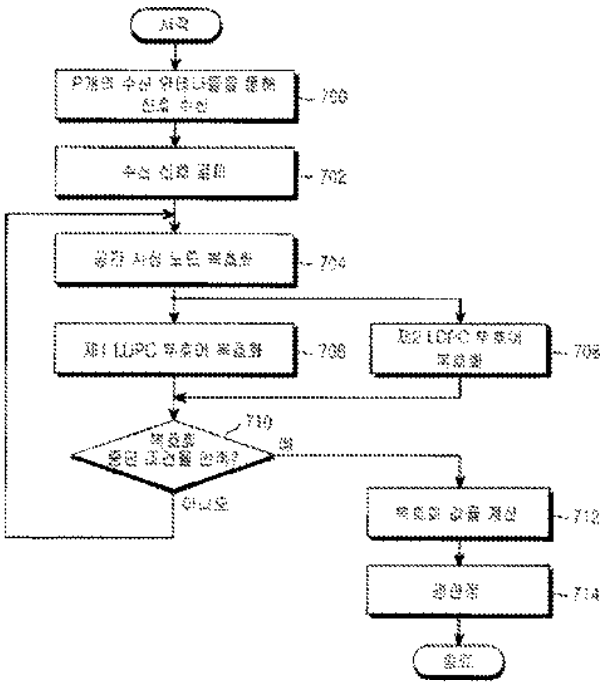
도면5


502
504
506
508
510
512
514
516

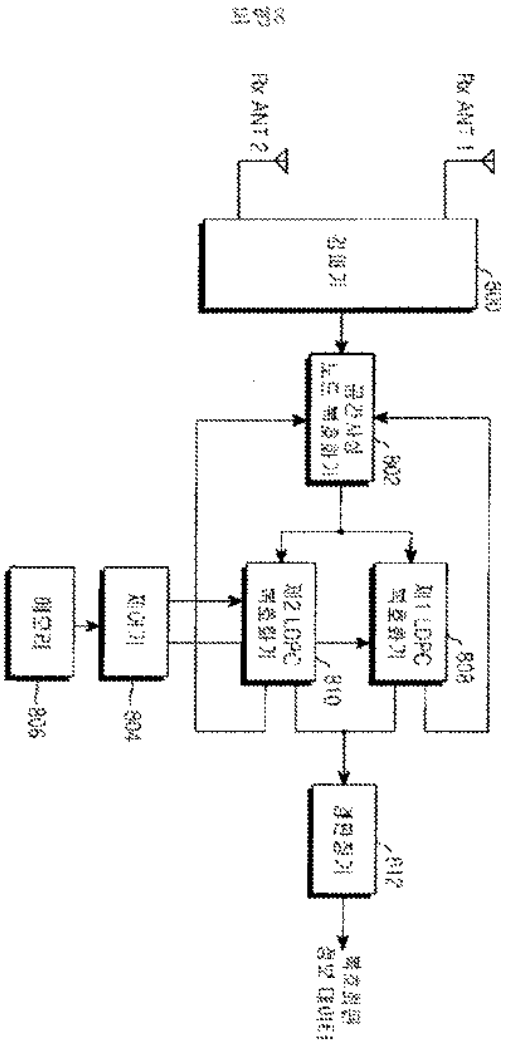
도면6



도면7



도면8


Rx ANT 2
Rx ANT 1
검파기
800
공간사상 노드 복호화기
802
제1 LDPC 복호화기
808
제2 LDPC 복호화기
810
제어기
804
메모리
806
결정기
812
복호화된 정보 데이터

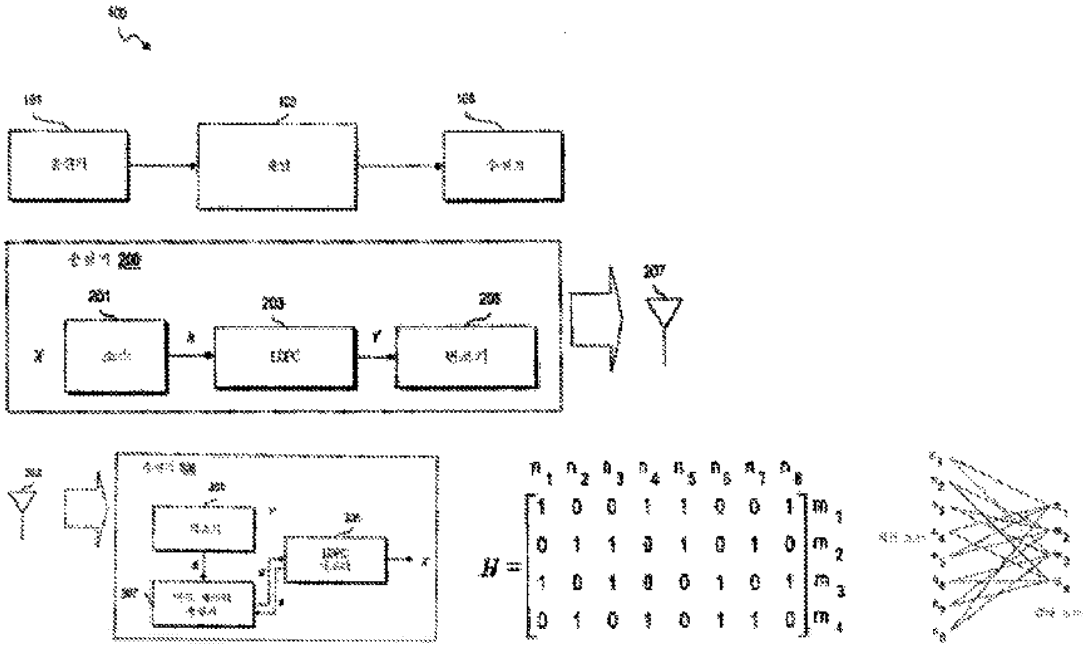
KR100634250B1

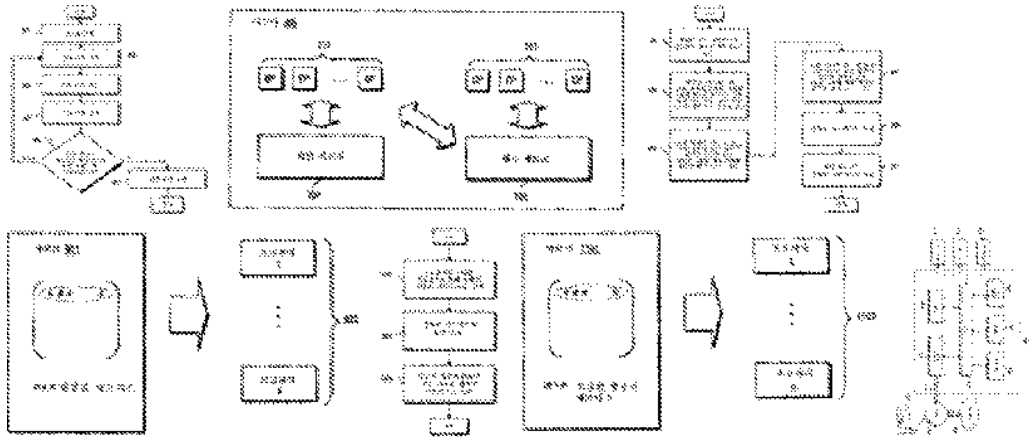
Method and system for generating parallel decodable low density parity check ldpc codes

Abstract

A method of efficiently decoding low density parity check (LDPC) codes is provided. LDPC decoder 305 includes memories 901 and 1101 that store mapped matrices that satisfy a plurality of parallel decodable conditions that allow package memory structures 901 and 1101. Decoder 305 also includes parallel processors 903 and 1103 that access edge values from a stored mapped matrix that decodes LDPC codes. The method has special applicability for satellite broadcast systems.
Low density parity check (LDPC) code, mapped matrices, parallel processors,

Images (12)





Classifications

H03M13/1168 Quasi-cyclic LDPC [QC-LDPC] codes, i.e. the parity-check matrix being composed of permutation or circulant sub-matrices wherein the sub-matrices have column and row weights greater than one, e.g. multi-diagonal sub-matrices

View 3 more classifications

South Korea

Download PDF Find Prior Art Similar

Other languages

Korean

Inventor

린-난 리

펭-웬 선

무스타파 에로즈

Worldwide applications

2004 US JP KR CA EP CN 2005 *HK*

Application KR20040051588A events

2003-07-03

Priority to US48497403P

2003-07-03

Priority to US60/484,974

2004-07-02

Application filed by 더 디렉티브 그룹, 인크.

2005-01-12

Publication of KR20050004118A

2006-10-16

Application granted

2006-10-16

Publication of KR100634250B1

Info

Patent citations (7)

Cited by (49)

Legal events

Similar documents

Priority and Related Applications

External links

Espacenet

Global Dossier

Discuss

Description

METHOD AND SYSTEM FOR GENERATING PARALLEL DECODABLE LOW DENSITY PARITY CHECK (LDPC) CODES}

1 is a communication system configured to use a low density parity check (LDPC) code in accordance with an embodiment of the present invention.

2 is an exemplary transmitter of the system of FIG.

3 is an exemplary receiver of the system of FIG.

4 is a low density parity check matrix in accordance with an embodiment of the present invention.

5 is a bipartite graph of the LDPC code of the matrix of FIG.

6 is an operational flowchart of the LDPC decoder of FIG. 3 in accordance with an embodiment of the present invention.

7 illustrates components of an LDPC decoder in accordance with an embodiment of the present invention.

8 is a flow diagram of a process for generating a mapped matrix for efficient parallel decoding in accordance with various embodiments of the present invention.

9 is an LDPC decoder with a memory that stores a mapped matrix of edge values for supporting parallel decoding in accordance with an embodiment of the present invention.

10 is a flow chart of a process for generating a derived mapped matrix to accommodate parallel decoding in accordance with various embodiments of the present invention.

11 is an LDPC decoder with a memory that stores a derived mapped matrix of edge values for supporting parallel decoding in accordance with an embodiment of the present invention.

12 is a computer system capable of performing a process of encoding and decoding LDPC codes in accordance with embodiments of the present invention.

<Explanation of symbols for the main parts of the drawings>

101: transmitter

103: channel

105: receiver

200: transmitter

201: source

203: LDPC encoder

205: modulator

300: receiver

301: demodulator

305: LDPC decoder

307: Bit Metric Generator

<Related application>

This application is related to U.S. Provisional Application No. 60 / 484,974, filed July 3, 2003, entitled "General Parallel Decoding LDPC Code." It claims its priority date under section 119 (e), the entire contents of which are incorporated herein.

The present invention relates to a communication system, and more particularly to a coded system.

Communication systems use coding to ensure reliable communication over noisy communication channels. The communication channel has a fixed capacity that can be expressed in bits per symbol at a given signal-to-noise ratio (SNR), which defines a theoretical upper limit known as the Shannon limit. As a result, coding designs have aimed to achieve speeds near the Shannon limit. One type of code near the Shannon limit is the low density parity check (LDPC) code.

Traditionally, LDPC code has not been widely available due to many shortcomings. One of the drawbacks is that the LDPC encoding technique is very complex. To encode an LDPC code using its constructor matrix, you need to store a very large non-sparse matrix. In addition, the LDPC code

requires a large block to be valid, and as a result, there is a problem in storing the matrix even when the parity check matrix of the LDPC code is sparse.

There are also many challenges to implementation. For example, storage is an important reason why LDPC codes are not really widely used. One of the major issues in LDPC decoding is the organization of memory. Recognizing that the larger the memory size, the lower the cost per bit, there is a need for a study of the code architecture that allows an efficient lumped memory structure for large edge values in the LDPC decoder.

In addition, a major challenge in LDPC code implementation is how to achieve a connection network between various processing engines (nodes) in a decoder. In addition, the computational load in the decoding process, in particular the check node operation, causes problems.

In general, it is recognized that code corresponding to a set number of parallel engines cannot be reconfigured to accommodate other numbers of parallel engines in order to maintain an efficient memory architecture. This inflexibility presents a serious drawback because different applications require different decoding rates. This limitation also hinders the ability to accommodate and exploit improvements in semiconductor performance. For example, as the processing power increases, the number of parallel engines required for a given decoding speed should decrease, but the design of a fixed number of engines does not allow a direct reduction in the number of processors used.

Thus, there is a need for an LDPC communication system that utilizes an efficient decoding process. There is also a need to minimize storage requirements for implementing LDPC coding. There is also a need for a decoding scheme that can easily accommodate future technological advances.

The present invention seeks to solve this problem, and provides a solution for decoding a structured low density parity check (LDPC) code. In order to generate a mapped matrix corresponding to the parity check matrix, a criterion is set for the code capable of parallel decoding. The first criterion is that neither of any other two entries in the same row of the mapped matrix are connected to the same bit nodes or the same check nodes. Second, if the group of bit nodes represented by BG_i (i = 0,1, ..., m-1) represents all the bit nodes connected to the edges of the i th row, then for $i \neq j$, BG It is

certain that $_{i}$ and BG $_{j}$ have no common node or are the same. This criterion also applies to inspection nodes, whereby a group of inspection nodes represented by CG $_{i}$ (i = 0,1, ..., m-1) represents all inspection nodes connected to the edges of the i th row. . For i ≠ j, it is verified that CG $_{i}$ and CG $_{j}$ have no or the same common node. This criterion derives LDPC codes that utilize the parallel decoding structure of LDPC codes and provide an efficient memory architecture. Block matrix A, an n × n density balanced matrix, to maintain three conditions for efficient parallelism for fewer processors than n processors (i.e., m processors, where m is an integer fraction of n) $_{The}$ parity check matrix is constructed using $_{ij}$. The block matrix is decomposed into matrices with B $_{ij}$ entries which are m × m density balanced matrices by matrix exchange. In the bipartite graph, the derived code can be obtained by deleting certain edges in the mapped matrix. In this configuration, m parallel engines can be used, thus increasing the flexibility of the implementation.

According to one aspect of an embodiment of the present invention, a method for supporting decoding of a low density parity check (LDPC) code is disclosed. The method includes constructing a mapped matrix based on the parity check matrix associated with the LDPC code. The mapped matrix meets a number of parallel decodable conditions to allow package memory structures accessible by multiple processors operating in parallel. The method also includes storing, by the processors, a matrix mapped to the memory for decoding the LDPC code.

According to another aspect of an embodiment of the present invention, a decoding apparatus for decoding a low density parity check code is disclosed. The apparatus includes a memory configured to store a mapped matrix that satisfies a number of parallel decodable conditions to allow a package memory structure. The apparatus also includes a plurality of processes operating in parallel, which access the memory to decode the LDPC code.

According to yet another aspect of an embodiment of the present invention, a method for supporting decoding of a low density parity check code is disclosed. The method includes initializing the mapped matrix based on a parity check matrix corresponding to the LDPC code. The method also includes determining that none of any other two entries in the same row of the mapped matrix are connected to the same bit nodes or the same check nodes. If the bit node group BG $_{i}$ represents all the bit nodes connected to

the edges of the i th row, this method verifies that for i ≠ j, BG_i and BG_j have no common nodes or are the same. It includes a step. In addition, if the check node group CG_i represents all check nodes connected to the edges of the i th row, the method verifies that for i ≠ j, CG_i and CG_j have no common nodes or are the same. It includes a step. The method also includes outputting and storing a matrix mapped to a memory accessible by a plurality of processors operating in parallel, and decoding the LDPC code according to the stored mapped matrix.

Other features and advantages of the invention will be apparent from the following detailed description with reference to numerous specific embodiments and implementations, including the best mode devised to practice the invention. In addition, the present invention may be possible in other embodiments, and various modifications may be made without departing from the spirit and scope of the present invention. Accordingly, the drawings and detailed description are by way of example and not limitation.

The invention is illustrated by way of example and not by way of limitation with reference to the accompanying drawings, in which like elements are designated by like reference numerals.

Systems, methods and software for efficiently decoding structured low density parity check codes are described. In the following description, for purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. However, one skilled in the art will recognize that the present invention may be practiced without these specific details or with an equivalent configuration. In many instances, well-known structures and devices are shown in block diagram form in order not to unnecessarily obscure the present invention.

1 illustrates a communication system configured to use a low density parity check code in accordance with an embodiment of the present invention. The digital communication system 100 includes a transmitter 101 for generating a signal waveform to be transmitted to the receiver 105 via a communication channel 103. In this separate communication system 100, the transmitter 101 has a message source that generates a separate set of possible messages, each of which has a corresponding signal waveform. These signal waveforms are attenuated or modified by the communication channel 103. LDPC codes are used to solve the noise channel 103.

The LDPC code generated by the transmitter 101 enables high speed implementation without incurring any performance loss. This LDPC code has a parallel decoding algorithm that advantageously includes simple operations such as addition, comparison and table search (unlike turbo codes). This parallel architecture is used to minimize the memory requirements and to increase flexibility in using a large number of parallel engines (ie processors). This will be described later with reference to FIGS. 7-11. Moreover, carefully designed LDPC codes do not exhibit any error floor indication in the range of interest.

According to one embodiment of the present invention, the transmitter 101 generates and communicates with the receiver 105 an LDPC code based on a parity check matrix that facilitates efficient memory access during decoding.

2 is a diagram of an exemplary transmitter in the system of FIG. The transmitter 200 is equipped with an LDPC encoder 203 which receives input from the information source 201 and outputs a coded stream with redundancy suitable for the error correction process at the receiver 105. Information source 201 generates k signals from discrete alphabet X. LDPC codes are specified using a parity check matrix.

The modulator 205 maps the encoded message from the encoder 203 to a signal waveform transmitted to the transmit antenna 207, which transmits this waveform through the communication channel 103. As such, the encoded message is modulated and distributed to the transmit antenna 207. As explained below, the transmission from the transmit antenna 207 propagates to the receiver.

3 is a diagram of an exemplary receiver in the system of FIG. On the receiving side, the receiver 300 includes a demodulator 301 that demodulates the signal received from the transmitter 200. This signal is received at the receive antenna 303 for demodulation. After demodulation, the received signal is sent to decoder 305, which attempts to reconstruct the original source message by generating message X 'in conjunction with bit metric generator 307. The bit metric generator 307 can exchange information (forwards and backwards) with the decoder during the decoding process. This decoding scheme was filed on July 3, 2003 and co-pended "Method and System for Routing in Low Density Parity Check Decoders" (Application No. 10 / 613,824, Agent Revision No. PD-203009, the entire contents of which

are incorporated herein by reference. In order to understand the advantages provided by the present invention, it is desirable to consider how LDPC codes are generated as described in FIG.

4 is a diagram of a sparse parity check matrix in accordance with an embodiment of the present invention. The LDPC code is a long linear block code with a sparse parity check matrix $H_{(nk)\,xn}$. Typically, block length n is in the range of thousands to tens of thousands of bits. For example, a parity check matrix for an LDPC code of length n = 8 and rate 1/2 is shown in FIG. The same code can be represented equivalently by the bipartite graph of FIG.

5 is a diagram of a bipartite graph of an LDPC code of the matrix of FIG. The parity check equation implies that for each check node the sum of all adjacent bit nodes (on the Galois Field) 2 is zero. As shown in the figure, the bit nodes occupy the left side of the graph and are associated with one or more check nodes according to a predetermined relationship. For example, corresponding to check node m_1 , there is a representation of $n_1 + n_4 + n_5 + n_8 = 0$ with respect to the bit node.

The LDPC decoder 305 considers a message passing decoder to find the value of the bit node. To accomplish this task, bit nodes and check nodes repeatedly communicate with each other. The nature of this communication is described below.

From the check node to the bit node, each check node provides one neighbor bit node with an estimate ("opinion") of the value of that bit node based on information from the other neighbor bit node. For example, in the above example, n_4, if n_5 and n_8 sum is "visible" to 0 to m_1 in, m_1 informs to n_1 that the estimated that the value of n_1 0 ($n_1 + n_4 + n_5 + n_8 = 0$). Otherwise, m_1 tells $n_{1\ that}$ the value of n 1 is assumed to be $_1$. In addition, for soft decision decoding, a reliability measure is added.

From the check node to the bit node, each bit node relays to one adjacent check node an estimate of the value of that check node based on feedback from the other adjacent check node. In this example, n_1 has only two adjacent check nodes m_1 and m_3 . If the feedback from m_3 to n_1 indicates that the value of n_1 is probably zero, n_1 will inform m_1 that the estimate of the value owned by n_1 is zero. If a bit node has more than two neighbor check nodes, the bit node performs a majority vote decision (soft decision)

on the feedback from the other neighbor check node before reporting the decision to the check node in communication with it. The process is repeated until all bit nodes are considered correct (i.e., until all parity checks are satisfied) or until a predetermined maximum number of iterations has been reached and a decoding failure is declared.

6 is a flowchart of the operation of an LDPC decoder according to an embodiment of the present invention. It is recognized that one of the most prominent advantages of LDPC is the possibility for parallel decoding. This is evident by examining the bipartite graph and decoding algorithm of the LDPC code. Obviously, when processing a bit node (or check node), since there is no intermediate data exchange during this process, all the bit nodes (or check nodes) can be processed at the same time. For each bit node i, there is a received value u_1 that indicates that the log likelihood ratio of a particular bit is logical 0 rather than logical $_1$, which is referred to as the "channel value". For example, a message parsing algorithm for decoding LDPC codes is described.

In step 601, initialization of the decoder 305 is performed, whereby a value is assigned to each edge of the bipartite graph (Figure 5). This value is equal to the channel value associated with the bit node to which the edge is connected. Next, the check node is updated through step 603. That is, at the end of the check node process, all edge values are updated. If e_1, e_2, ..., e_v represent all the values of all the edges connected to a particular check node, then the updated edge values are as follows.

$e_i = g(e_1, e_2, ..., e_{i-1}, e_i, ... e_v)$; i = 1, 2, ... v

From here,

$$g(a,b)=\ln\frac{1+e^{a+b}}{e^{a}+e^{b}}$$

to be. The function g (a, b, c) = g (g (a, b), c) can be calculated recursively.

After all check nodes have been processed, the bit nodes are processed with the updated edge values, as in step 605. The goal of the bit node process is to update the edge values further. Assuming that e_1, e_2, ..., e_v represent the edge values of all edges connected to this particular bit node, the updated edge values are as follows.

$e_i = e_1 + e_2 + \ldots + e_{i-1} + e_i + \ldots e_v$

The process outputs a posteriori probability information through step 607 and determines whether the parity check equation has been satisfied or the configurable maximum number of iterations has been reached (step 609). If these equations are not satisfied, steps 603-607 are executed repeatedly. At the final iteration, all edge values associated with one bit node are summed and further summed with the channel values through step 611 to form the final decision metric.

Since the LDPC decoder 305 can be implemented as a highly parallel system, there are two ways of realizing the interconnection between the check node and the bit node: (1) fully parallel and (2) partially parallel. do. In a fully parallel architecture, all of the nodes and their interconnections are physically implemented. The advantage of this architecture is speed. However, in a fully parallel architecture, realizing nodes and all of their interconnections introduces much greater complexity. Therefore, in a fully parallel architecture, smaller block sizes are required to reduce complexity.

The second way to implement an LDPC code is to physically realize only a subset of the total number of nodes, and to process all of the "functional" nodes of the code using this limited number of "physical" nodes. Although the operation of the LDPC decoder 305 can be very simple and can be performed in parallel, there is another design problem with how to set up communication between the "randomly" distributed bit nodes and the check nodes.

7 is a diagram of components of an LDPC decoder according to an embodiment of the present invention. Decoder 305 executes the decoding process of FIG. 7 using a parallel processor. As shown, the decoder 305 includes a bit node processor 701 and a check node processor 703, which perform bit node and check node operations, respectively. The intermediate edge value of the bit node or check node update is buffered in the edge memory 705. Also, the partial sum in forming a new edge value during one bit node process is typically buffered locally at the processor. Channel memory 707 maintains reliability information from the channel. As can be seen in the figure, channel reliability memory 707 is used only in a bit node process.

The design of the decoder 305 involves a number of engineering tradeoffs such as the number of parallel processors versus the speed of decoding, and the memory size. In one extreme example, one processor may be used to decode the entire code sequentially from one node to another. However, due to the number of nodes and the number of iterations for a typical good LDPC code, this approach is rarely useful. Another extreme example is to use as many processors as there are nodes. According to this, the highest decoding speed can be obtained, but unfortunately the cost is the highest. Practical decoders are designed between these two extremes.

Clearly, there is a tradeoff between decoding speed and the number of bit nodes and / or check node processors. Although the bit node processor 701 and the check node processor 703 are shown as separate components, the bit node processor 701 may be coupled to the check node processor 703 to allow them to share a portion of the data path. Will know. Alternatively, the bit node processor 701 may be identical to the check node processor 703 by performing the conversion of the edge value before the process. It should also be noted that the number of bit node processors 701 need not be the same as the number of check node processors 703.

Also, an efficient architecture should not arbitrarily handle the number of parallel process engines. It is desirable to have some freedom in the number of parallels to provide flexibility in implementation to accommodate advances in technical improvements. For example, if the code is designed to have 360 parallel engines, as integrated circuit technology advances, the circuit may operate at twice the speed. Thus, it is desirable to maintain the same decoding architecture while reconfiguring the same code to have 180 parallel engines.

In relation to the memory structure, the LDPC decoder 305 places the edge memory 705 in proximity to each of the processors 701 and 703. Registers or RAM may be used as the memory 705. For long codes, the use of registers is much more expensive than RAM. With regard to RAM, it is recognized that the cost per bit decreases as the size of RAM increases. Therefore, it is desirable to lump as many RAMs as possible so that all edge values are functionally lumped into one RAM.

It is recognized that arbitrary LDPC codes cannot efficiently support the decoder architecture of FIG. This is evident by looking at the standard RAM architecture. The following observations are noted. RAM is a memory having

a predefined width and depth and can be read and written in a row-wise manner. This means that in one access cycle, only values within the same row can be accessed. The term "random" merely means that rows can be accessed in any order in each access cycle. Therefore, in terms of efficiency, in any given access cycle, each bit node processor 701 (check node processor 703 when processing on the check node side) is supplied with the same amount of edge value input. Should be. Thus, at any given moment, an efficient decoder architecture must ensure that the supply to the process engine is not interrupted.

It is also desirable to dedicate one processor to a particular bit node (or check node) at any given time. That is, the processor must not switch to another bit node (or check node) before finishing updating the edge associated with one bit node.

As will be described below, the present invention introduces the concept of parallel decodable code in order to explain various characteristics of the architecture of FIG.

8 is a flowchart of a process for generating a mapped matrix for efficient parallel decoding according to various embodiments of the present invention. Parallel decodable code is defined to represent an efficient memory structure. If there is a one-to-one mapping that maps the nonzero item of the parity check matrix to a matrix of m x n (mapped matrices) that satisfies a number of defined attributes, then the LDPC code will have n parallel engines with package memory. Can be decoded. The process for generating this mapped matrix is described below.

In step 801, based on the parity check matrix, a mapped matrix is generated. This mapped matrix is then verified to ensure various conditions or attributes for efficient parallel decoding. Through step 803, the matrix is verified so that no two different items in the same row of the mapped matrix are connected to the same bit node or the same check node. If the bit node group i (i = 0, 1, ..., m-1) represented by BG_i is all bit nodes connected to the edge in the i th row, then for i ≠ j, the process is BG_i And ensure that BG_j has no common node or is the same (step 805). Similarly, if through step 807 the check node group i (i = 0, 1, ... m-1), denoted by CG_i, is all check nodes connected to the edge in the i th row, i ≠ For j, CG_i and CG_j have no common node or are verified to be the same.

The three conditions of steps 803-807 are necessary for an efficient parallel process. In step 809, the mapped matrix is output and stored in the edge memory 705 (step 811). In an exemplary embodiment, the mapped matrix represents a memory structure. These memories are physically different memory devices and are functionally equivalent to the mapped matrix. The attribute of step 803 ensures that in each access cycle, each processor (eg, processors 710 and 703) will receive only one value during each access cycle.

In n parallel process engines, every processor will be given an equal load. The second attribute (step 805) divides the bit nodes into groups of size n. Two bit nodes are in the same group as long as there is an edge associated with nodes mapped to the same row. In addition, there is a need to keep certain intermediate results active during the process for each bit node. Thus, a set of processes of bit nodes (check nodes) must be completed before processing a new bit node (or check node). And this is ensured by the second and third attributes of steps 805 and 807.

9 is a diagram of an LDPC decoder having a memory that stores a mapped matrix of edge values to support parallel decoding according to an embodiment of the present invention. A mapped matrix of m × n is stored in memory 901 and edge values are provided to n parallel processors 903. For illustration purposes, the use of a mapped matrix is described in connection with the parity check matrix below.

$$\begin{bmatrix} 1 & 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 1 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 1 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 1 & 0 & 0 & 1 \\ 0 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 0 & 0 \end{bmatrix}$$

During decoding, each of the nonzero items of the parity matrix is associated with an edge value. This edge value is labeled as follows.

$$\begin{bmatrix} e_1 & e_4 & 0 & 0 & 0 & 0 & 0 & e_{19} & 0 & 0 & e_{25} & 0 & 0 & 0 & 0 & 0 \\ 0 & e_5 & e_7 & 0 & 0 & 0 & e_{17} & 0 & 0 & 0 & 0 & e_{27} & 0 & 0 & 0 & 0 \\ 0 & 0 & e_8 & e_{10} & 0 & e_{15} & 0 & 0 & e_{21} & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ e_2 & 0 & 0 & e_{11} & e_{13} & 0 & 0 & 0 & 0 & e_{22} & 0 & 0 & 0 & 0 & 0 & 0 \\ e_3 & 0 & 0 & 0 & 0 & e_{16} & 0 & 0 & e_{22} & 0 & 0 & 0 & 0 & e_{31} & e_{33} & 0 \\ 0 & e_6 & 0 & 0 & 0 & 0 & e_{18} & 0 & 0 & e_{24} & 0 & 0 & 0 & 0 & e_{34} & e_{35} \\ 0 & 0 & e_9 & 0 & 0 & 0 & 0 & e_{20} & 0 & 0 & e_{26} & 0 & e_{29} & 0 & 0 & e_{36} \\ 0 & 0 & 0 & e_{12} & e_{14} & 0 & 0 & 0 & 0 & 0 & 0 & e_{28} & e_{30} & e_{32} & 0 & 0 \end{bmatrix}$$

Edge values are mapped one-to-one into the following 9 × 4 matrix.

$$\begin{bmatrix} e_1 & e_5 & e_8 & e_{11} \\ e_2 & e_4 & e_7 & e_{10} \\ e_3 & e_6 & e_9 & e_{12} \\ e_{13} & e_{15} & e_{17} & e_{19} \\ e_{14} & e_{16} & e_{18} & e_{20} \\ e_{21} & e_{23} & e_{25} & e_{27} \\ e_{22} & e_{24} & e_{26} & e_{28} \\ e_{29} & e_{32} & e_{33} & e_{35} \\ e_{30} & e_{31} & e_{34} & e_{36} \end{bmatrix}$$

The above matrix is a dense matrix that 100% uses, assuming that the matrix mirrors the memory. In other words, all the cells of the memory are fully used. It can easily be proved that this mapping satisfies all three characteristics for efficient parallel decoding, as described above.

In this example, the bit nodes are divided into four groups, each group containing four bit nodes, and the bit node i is a bit node associated with column i of the original parity check matrix. These bit node groups are {1,2,3,4}, {5,6,7,8}, {9,10,11,12}, and {13,14,15,16}.

Similarly, check node i is a check node associated with row i of the raw parity check matrix. Check nodes are divided into two groups: {1,2,3,4}, and {5,6,7,8}. In this example, 4 parallel processors are employed (n = 4). For bit node processing, the first row is read by 4 parallel processors 903, and the intermediate results are separated because the purpose is to update the edges associated with bit nodes {1,2,3,4}. It will be created and buffered in memory. This is accomplished after all original edge values of {1,2,3,4} have been processed.

Note that all edge values associated with the bit node {1,2,3,4} are located in the first three rows of the edge matrix. Therefore, in the next access cycle, the second row or the third row will continue to be accessed. If the process starts on any other row, additional intermediate results will need to be buffered. After all the rows associated with the bit node {1,2,3,4} have been processed and the edge values have been updated, other groups of bit nodes are processed.

Similarly, the above process is performed at check nodes. That is, check nodes are processed into groups, and after all edges associated with this group have been updated, the next group is processed. In each access cycle, exactly one value from the edge matrix is taken for each node currently being processed, so that none of the processors 903 are overloaded (or poorly used in that matter). .

It is recognized that the three conditions for efficient parallel processing (as described in steps 803-807 in Figure 8) may be somewhat limited, and that a random LDPC code will generally not meet these conditions. This restriction can be lifted by deriving another matrix from the mapped matrix, as described below.

10 is a flow diagram of a process for generating a derived matrix derived to accommodate parallel decoding in accordance with various embodiments of the present invention. To better understand the derived mapped matrix generation, certain techniques and theorems are described as follows. The n x n matrix is called a "dense balanced matrix" if each row and each column of the matrix is equal to the number of entries that are not exactly zero. One example of a density balanced matrix is an all-zero matrix; This matrix is widely used for LDPC code generation.

Another example of a density balance matrix is a matrix having the following characteristics. Assuming that $b_0 b_1 \ldots b_{n-1}$ is a binary vector and a is relatively prime to n, then the matrix whose entries in the i th row and j th column are b_{ai+j} is a density balanced matrix. In particular, when a = 1, it results in a circulant matrix.

It is noted that the permutation of any row and / or column of the density balance matrix results in a density balance matrix. However, because different component matrices can be exchanged differently, basically different LDPC codes result in using building blocks of different exchanges.

In the density balance matrix, if the number of nonzero entries in one row is k, then all nonzero entries can be selected (one-to-one) by taking one nonzero entry one row at a time and then mapping it to the k x n density matrix.

The following theorem is proposed (Theorem 1):

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1w} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2w} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{v1} & A_{v2} & \cdots & \cdots & A_{vw} \end{bmatrix}$$

All three conditions for n efficient parallel decoding are satisfied only when there is a parity check matrix of LDPC codes that are equivalent exchanges in the block matrix such that each of the component block matrices A_{ij} is an nxn density balanced matrix. Note that the matrix thus constructed does not need to generate generic LDPC code.

Given the limitations imposed by the theorem on the structure of the parity check matrix, it is obvious that the general parity check matrix will not satisfy these conditions. Therefore, there is a need to design code with this structural limitation in order to be able to efficiently decode and access memory.

It is recognized that code capable of efficient parallel decoding with n parallel engines may not be capable of efficient parallel decoding for other numbers (ie, with a number of engines other than n). As previously described, in order to account for future developments in processing, low parallelism is desirable for future audiences. However, all the conditions for efficient parallel processing need to be satisfied even in the case of low parallelism.

In order to maintain 3 conditions for efficient parallel processing for less than n processors, another theorem is proposed (Theorem 2). Assume that the following matrix is a parity check matrix, and each of the block matrices A_{ij} are nxn density balanced matrices.

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1w} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2w} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{v1} & A_{v2} & \cdots & \cdots & A_{vw} \end{bmatrix}$$

Where A_{11} is an m-by-m density balanced matrix, each of B_{ij}, and for all other $A_{ij} \neq A_{11}$, block matrices can be decomposed into block density balanced submatrices under the same row and column exchange as A_{11}. And if it can be decomposed into one matrix under heat exchange,

$$A_{11} = \begin{bmatrix} B_{11} & B_{12} & \cdots & \cdots & B_{1P} \\ B_{21} & B_{22} & \cdots & \cdots & B_{2P} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ B_{Q1} & B_{Q2} & \cdots & \cdots & B_{QP} \end{bmatrix}$$

LDPC codes are defined as follows:

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1w} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2w} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{v1} & A_{v2} & \cdots & \cdots & A_{vw} \end{bmatrix}$$

The above matrix satisfies the conditions for efficient parallel processing in n and m parallelism.

By way of example, the following is defined: A_{ij} is an nxn circulant matrices, and m is any integer fraction of n. The rows and columns of A_{ij} are leveled to 0,1, ..., n. The submatrix of A_{ij}, whose row and column indices consist of all elements such as the constant modulo m, is also a circulant. As a result, under the same row and column exchange, all A_{ij} can be decomposed back into circular submatrices. This implies that an LDPC code with A_{ij} as component code satisfies the conditions for efficient parallel processing with both n and m parallelism, where m is any integer fraction of n.

In any given LDPC code with parity check matrix H, other LDPC codes may be derived from H by converting some of the non-zero entries of H to zeros. Any code obtained in this manner is termed "derived code of the raw LDPC code." In step 1001, the mapped matrix for the n parallel decoded code is received. If the source code is n parallel decoding with full dense memory of n parallel engine check matrices, any derived code may also be n parallel decodable with a reduced density memory matrix.

In terms of a bipartite graph, the derivation code can be obtained by deleting some edges (step 1003). The mapped matrix of derived code is essentially the same as the mapped matrix of base code. However, entries

corresponding to deleted edges may be marked as "DON'T CARE" (ie, null values). Therefore, these memory cells are wasted, and access cycles are extended with fetching null values. However, as long as the total number of edge deletions is relatively small, it does not significantly affect the efficiency. Such derivation code may exhibit other desirable characteristics, such as easy encoding and improved distance characteristics. In step 1005, the derived mapped matrix is output.

FIG. 11 is a diagram illustrating an LDPC decoder having a memory storing a derived mapped matrix of edge values for supporting parallel decoding according to an embodiment of the present invention. This matrix is then stored in memory 1101 for access by m processors 1103. In this allocation, there is very flexibility in the number of parallel engines used, while maintaining an efficient memory structure.

Figure 12 illustrates a computer system in which one embodiment according to the present invention may be implemented. Computer system 1200 includes a bus 1201 or other communication mechanism for communicating information, and a processor 1203 coupled with bus 1201 for processing information. Computer system 1200 may access main memory 1205, such as random access memory (RAM) or other dynamic storage device, coupled to bus 1201 to store information and instructions executed by processor 1203. Also includes. Main memory 1205 may also be used to store temporary variables or other intermediate information during execution of instructions executed by processor 1203. Computer system 1200 further includes a read memory (ROM) 1207 or other static storage device coupled to bus 1201 to store static information and instructions for processor 1203. Storage device 1209, such as a magnetic or optical disk, is further coupled to bus 1201 to store information and instructions.

Computer system 1200 may display a display, such as a cathod ray tube (CRT), liquid crystal display (LCD), active matrix display, or plasma display, to display information to a computer user. 1211 may be connected via a bus 1201. Input device 1213, such as a keyboard that includes alphanumeric and other keys, is coupled to bus 1201 to communicate information and command selections to processor 1203. Another type of user input device is cursor control 1215, such as a mouse, trackball, or cursor direction keys, for communicating indication information and command selections to the processor 1203 and for controlling cursor operation on the display 1211.)to be.

In accordance with one embodiment of the present invention, parallel decoding of LDPC codes is provided by computer system 1200 in response to processor 1203 executing instructions contained in main memory 1205. Such instructions may be read into main memory 1205 from another computer readable medium, such as storage device 1209. Execution of instructions included in main memory 1205 causes processor 1230 to perform the process steps described herein. One or more processors in multiprocessing may also be employed to execute the instructions contained in main memory 1205. In still other embodiments, wired circuitry may be used in place of or in combination with software instructions to implement embodiments of the present invention. Therefore, embodiments of the present invention are not limited to any particular combination of hardware circuitry and software.

Computer system 1200 also includes a communication interface 1217 connected by bus 1201. The communication interface 1217 provides bidirectional data communication coupled to a network link 1219 coupled to a local area network 1221. For example, communication interface 1217 may be a digital subscriber line (DSL) card or modem, an integrated services digital network (ISDN) to provide a data communication connection to a corresponding type of telephone line. Card, cable modem, or telephone modem. In another example, communication interface 1217 may be a local area network (LAN) card (eg, Ethernet or Asynchronous Trasfer Model (ATM) network) that provides a data communication connection to a compatible LAN. Can be. Wireless links can also be implemented. In any such implementation, communication interface 1217 sends and receives electrical, electromagnetic, or optical signals that carry digital data streams representing various types of information. In addition, the communication interface 1217 may include peripheral interface devices, such as a universal serial bus (USB) interface, a Personal Computer Memory Card International Association (PCMCIA) interface, and the like.

Network link 1219 typically provides data communication to other data devices over one or more communication networks. For example, network link 1219 may be a network 1225 (e.g., a telecommunications network (WAN)) or a general purpose packet data network, now commonly referred to as the "Internet" or data device operated by a service provider. May provide a connection through local area network 1221 to a host computer 1223. Local area network 1221 and communication network 1225 both use electrical, electromagnetic, or optical signals to convey information and

commands. Signals over various networks and signals over communication link 1219 and communication interface 1217, which communicate digital data with computer system 1200, may be in the form of carrier waves containing information and instructions. Examples are.

Computer system 1200 may send and receive data and messages, including program code, via network (s), network link 1219, and communication interface 1217. In the Internet example, a server (not shown) may transmit the required code pertaining to an application for implementing one embodiment of the present invention via communication network 1225, local area network 1221, and communication interface 1217. . Processor 1203 may execute the transmitted code while receiving and / or storing code in storage 129 or non-volatile storage for later execution. In this manner, computer system 1200 may obtain application program code in the form of a carrier wave.

As used herein, the term “computer readable medium” refers to any medium that serves to provide instructions to the processor 1203 for execution. Such a medium may take many forms, including, but not limited to, nonvolatile media, volatile media, and transmission media. For example, nonvolatile media include optical or magnetic disks, such as storage device 1209. Volatile media include dynamic memory, such as main memory 1205. Transmission media include coaxial cables, lead wires, and fiber optics, including wires, including bus 1201. Transmission media may also take the form of acoustic, light, or electromagnetic waves, such as those generated during Radio Frequency (RF) and infrared (IR) data communications. Common forms of computer readable media include, for example, floppy disks, flexible disks, hard disks, magnetic tapes, any other magnetic media, CD-ROM, CD-RW, DVD, any Other optical media, punch cards, paper tape, optical mark sheets, holes or any other physical media having patterns of other optically recognizable marks, RAM, PROM, and EPROM, FLASH-EPROM, any other memory chip or cartridge, carrier wave, or any other medium readable by a computer.

Various forms of computer readable media may be involved in providing instructions to a processor for execution. For example, instructions to perform at least a portion of the invention may be first on a magnetic disk of a remote computer. In such a scenario, the remote computer loads the instructions into main memory and uses the modem to send the instructions over the telephone line. Modems in premises computer systems receive data

over telephone lines, use infrared transmitters to convert data into infrared signals, and transmit infrared signals to mobile computing devices, such as PDAs and laptops. The infrared detector of the mobile computing device receives the information and commands carried by the infrared signal and loads the data on the bus. The bus delivers data to main memory where the processor receives and executes the instructions. Instructions received by the main memory may optionally be stored on the storage device either before or after execution by the processor.

Accordingly, various embodiments of the present invention provide an approach for decoding structured Low Density Parity Check (LDPC) codes. Conditions of the parallel decodable codes are made to generate a mapped matrix according to the parity check matrix. These conditions result in LDPC codes that develop a parallel decoding structure of LDPC codes and provide an efficient memory structure. The approach also allows preservation of these parallel decodable conditions for efficient parallel processing for fewer than n processors (ie m processors where m is an integer fraction of n). Under the arrangement above, decoder resources, such as memory and processing engines, are effectively employed. In addition, implementation flexibility is improved and m parallel engines may be used.

While the invention has been described in connection with a number of embodiments, the invention includes, without limitation, various obvious modifications and equivalent arrangements, which are within the scope of the appended claims.

Claims (19)

Hide Dependent

1. A method of supporting decoding of low density parity check (LDPC) codes,

   Constructing a mapped matrix based on a parity check matrix associated with the LDPC code, the mapped matrix being a lumped memory structure that can be accessed by a plurality of processors 903 and 1103 operating in parallel; Satisfies a plurality of parallel decodable conditions allowing 901, 1101; And

Storing the mapped matrix in memory 901, 1101 for decoding the LDPC code by the processor 903, 1103.

Support method comprising a.

2. The method of claim 1, wherein the parallel decodable condition is:

   Determining that none of any two different entries in the same row of the mapped matrix are connected to the same bit node or the same check node;

   If the bit node group BG_i represents all the bit nodes connected to the edge of the i < th > row, for $i \neq j$, verifying that BG_i and BG_j have no or the same common node; And

   If the check node group CG_i represents all check nodes connected to the edge of the i th row, then for $i \neq j$, verifying that CG_i and CG_j have no or the same common node.

3. 2. The method of claim 1 wherein the parity check matrix is a permutation equivalent to a block matrix having entries of an n × n density balance matrix.
4. 4. The method of claim 3 wherein one of the entries of the n × n density balance matrix is an m × m density balance matrix and the other entries of the n × n density balance matrix can be decomposed into a block density balance submatrix.
5. The method of claim 4, wherein m is an integer fraction of n.
6. The LDPC code of claim 1, wherein the LDPC code is selected from 8-PSK (Phase Shift Keying), 16-QAM (Quadrature Amplitude Modulation), 16-APSK (Amplitude Phase Shift Keying), 32-APSK, and QPSK (Quadrature Phase Shift Keying). A support method represented by a signal modulated according to a signal constellation comprising one.
7. 7. The method of claim 6 wherein the signal is transmitted over a satellite link in accordance with a digital video broadcast.
8. A computer readable medium having instructions that support decoding of low density parity check (LDPC) codes, the instructions being computer readable that, when executed, cause one or more processors 903, 1103 to perform the support method described in claim 1. Media available.

9. A decoding device for decoding a low density parity check (LDPC) code,

Memory 901, 1101 configured to store a mapped matrix that satisfies a plurality of parallel decodable conditions allowing package memory structures 901, 1101; And

A plurality of processors 903 and 1103 operating in parallel, wherein the processors 903 and 1103 access memory 901 and 1101 to decode the LDPC code;

Decoding apparatus comprising a.

10. The method of claim 9, wherein the parallel decodable condition is:

Determining that none of any two different entries in the same row of the mapped matrix are connected to the same bit node or the same check node;

If the bit node group BG_i represents all the bit nodes connected to the edge of the i < th > row, for $i \neq j$, verifying that BG_i and BG_j have no or the same common node; And

If the check node group CG_i represents all check nodes connected to the edge of the i < th > row, then for $i \neq j$, verifying that CG_i and CG_j have no or the same common node.

11. 10. The derived mapped matrix of claim 9, wherein the number of processors (903, 1103) is n, and the memory (901, 1101) is constructed based on the mapped matrix by deleting edge values in the mapped matrix. And a new mapped matrix is used by the m parallel processors (903, 1103), where m is less than n.
12. The decoding apparatus of claim 11, wherein m is an integer fraction of n.
13. 12. The decoding apparatus of claim 11, wherein the parity check matrix is an exchange equivalent to a block matrix having entries of an $n \times n$ density balance matrix.

14. The decoding apparatus of claim 13, wherein one of the entries of the n × n density balance matrix is an m × m density balance matrix, and another entry of the n × n density balance matrix can be decomposed into a block density balance submatrix.
15. The LDPC code of claim 9, wherein the LDPC code is selected from 8-PSK (Phase Shift Keying), 16-QAM (Quadrature Amplitude Modulation), 16-APSK (Amplitude Phase Shift Keying), 32-APSK, and QPSK (Quadrature Phase Shift Keying). A decoding device represented by a signal modulated according to a signal constellation comprising one.
16. 16. The decoding apparatus of claim 15, wherein the signal is transmitted via a satellite link according to digital video broadcast.

17. A method of supporting decoding of low density parity check (LDPC) codes,

Storing a mapped matrix based on the parity check matrix corresponding to the LDPC code;

Determining that none of any two different entries in the same row of the mapped matrix are connected to the same bit node or the same check node;

If the bit node group BG_i represents all the bit nodes connected to the edge of the i < th > row, for $i \neq j$, verifying that BG_i and BG_j have no or the same common node;

If the check node group CG_i represents all check nodes connected to the edge of the i < th > row, for $i \neq j$, verifying that CG_i and CG_j have no or the same common node; And

Outputting the mapped matrix and storing it in memories 901 and 1101 accessible by a plurality of processors 903 and 1103 operating in parallel and decoding the LDPC code according to the stored mapped matrix.

Support method comprising a.

18. 18. The processor of claim 17, wherein the number of processors 903 and 1103 is n.

The support method,

Outputting the derived matrix by deleting edge values in the mapped matrix;

The derived mapped matrix is used by m parallel processors (903, 1103), where m is less than n.

19. A computer readable medium having instructions that support decoding of low density parity check (LDPC) codes, the instructions that, when executed, cause the one or more processors 903, 1103 to perform the support method described in claim 17. Readable Media.

Patent Citations (7)

Publication number	Priority date	Publication date	Assignee	Title
US20020042899A1	2000-06-16	2002-04-11	Tzannes Marcos C.	Systems and methods for LDPC coded modulation
JP3833457B2	2000-09-18	2006-10-11	シャープ株式会社	Satellite broadcast receiving system
US6633856B2	2001-06-15	2003-10-14	Flarion Technologies, Inc.	Methods and apparatus for decoding LDPC codes
CN1185796C *	2002-11-15	2005-01-19	清华大学	Improved correcting decoding method for non-regular low-density parity-check code
US6957375B2 *	2003-02-26	2005-10-18	Flarion Technologies, Inc.	Method and apparatus for performing low-density parity-check (LDPC) code operations using a multi-level permutation
JP4225163B2 *	2003-05-13	2009-02-18	ソニー株式会社	Decoding device, decoding method, and program
US7296208B2 *	2003-07-03	2007-11-13	The Directv Group, Inc.	Method and system for generating parallel decodable low density parity check (LDPC) codes

* Cited by examiner, † Cited by third party

Cited By (49)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
US7203887B2 *2002-07-032007-04-10The Directtv Group, Inc.Method and system for routing in low density parity check (LDPC) decoders
US7577207B22002-07-032009-08-18Dtvg Licensing, Inc.Bit labeling for amplitude phase shift constellation used with low density parity check (LDPC) codes
US7864869B2 *2002-07-262011-01-04Dtvg Licensing, Inc.Satellite communication system utilizing low density parity check codes
US7020829B22002-07-032006-03-28Hughes Electronics CorporationMethod and system for decoding low density parity check (LDPC) codes
US20040019845A1 *2002-07-262004-01-29Hughes ElectronicsMethod and system for generating low density parity check codes
US7296208B2 *2003-07-032007-11-13The Directv Group, Inc.Method and system for generating parallel decodable low density parity check (LDPC) codes
US8140931B22003-07-032012-03-20Dtvg Licensing, Inc.Method and system for generating parallel decodable low density parity check (LDPC) codes
US8374284B2 *2004-02-122013-02-12Apple, Inc.Universal decoder
US7760880B2 *2004-10-132010-07-20Viasat, Inc.Decoder architecture system and method
JP4519694B2 *2005-03-292010-08-04敏 後藤LDPC code detection apparatus and LDPC code detection method
JP4622654B22005-04-252011-02-02ソニー株式会社Decoding device and decoding method
EP1717959A1 *2005-04-292006-11-02STMicroelectronics N.V.Method and device for controlling the decoding of a LDPC encoded codeword, in particular for DVB-S2 LDPC encoded codewords
JP2008541630A *2005-05-132008-11-20フィッシャー−ローズマウント・システムズ・インコーポレーテッドＦｉｓｈｅｒ−Ｒｏｓｅｍｏｕｎｔ Ｓｙｓｔｅｍｓ， Ｉｎｃ. Fieldbus process communication using error correction
KR100791225B1 *2005-12-012008-01-03한국전자통신연구원Apparatus and Method for LDPC Decoding using type-classified index

US8122315B22005-12-012012-02-21Electronics And Telecommunications Research InstituteLDPC decoding apparatus and method using type-classified index

CN100425017C *2005-12-082008-10-08西安电子科技大学Encoder of parallel-convolution LDPC code based on precoding and its fast encoding method

KR100765522B1 *2006-02-062007-10-10엘지전자 주식회사Light Emitting Diode and Method for Manufacturing the same

JP2007215089A *2006-02-132007-08-23Fujitsu LtdDecoding apparatus and decoding method

US7644339B2 *2006-10-022010-01-05Broadcom CorporationOverlapping sub-matrix based LDPC (low density parity check) decoder

CN101192833B *2006-11-282011-12-07华为技术有限公司A device and method for low-density checksum LDPC parallel coding

US7992070B2 *2006-12-272011-08-02Nec Laboratories America, Inc.Bit-interleaved LDPC-coded modulation for high-speed optical transmission

US20100122143A12007-03-272010-05-13Hughes Network Systems, LlcMethod and system for providing low density parity check (ldpc) coding for scrambled coded multiple access (scma)

CN101364809B *2007-08-062011-09-07美国博通公司Decoder

JP4487213B22007-10-192010-06-23ソニー株式会社Decoding apparatus and method, and program

JP4487212B2 *2007-10-192010-06-23ソニー株式会社Decoding device and method, transmission / reception system, receiving device and method, and program

KR101502624B1 *2007-12-062015-03-17삼성전자주식회사Apparatus and method for channel encoding and decoding in communication system using low-density parity-check codes

JP5420566B22007-12-212014-02-19スリーエム イノベイティブ プロパティズ カンパニーMicrobial system and fluid sample analysis method

US8266493B1 *2008-01-092012-09-11L-3 Communications, Corp.Low-density parity check decoding using combined check node and variable node

WO2009143375A22008-05-212009-11-26The Regents Of The University Of CalforniaLower-complexity layered belief propagation deconding ldpc codes

US8335979B2 *2008-12-082012-12-18Samsung Electronics Co., Ltd.Contention-free parallel processing multimode LDPC decoder

US8296640B2 *2008-06-202012-10-23Samsung Electronics Co., Ltd.Method and apparatus for parallel processing multimode LDPC decoder
CN102460977A2009-05-272012-05-16诺沃尔赛特有限公司Iterative decoding of ldpc codes with iteration scheduling
CN101610229B *2009-07-162012-07-04清华大学Joint LDPC encoding recursive MSK modulation and demodulation system and method therefor
CN102098127B *2009-12-152014-11-05上海贝尔股份有限公司Multi-user domain parallel decoding method and device
US8566668B1 *2010-01-042013-10-22Viasat, Inc.Edge memory architecture for LDPC decoder
US8832534B1 *2010-01-042014-09-09Viasat, Inc.LDPC decoder architecture
US8402341B2 *2010-02-182013-03-19Mustafa ErozMethod and system for providing low density parity check (LDPC) encoding and decoding
US8782489B22010-02-182014-07-15Hughes Network Systems, LlcMethod and system for providing Low Density Parity Check (LDPC) encoding and decoding
WO2012097046A12011-01-142012-07-19Marvell World Trade Ltd.Ldpc multi-decoder architectures
US8972834B22012-08-282015-03-03Hughes Network Systems, LlcSystem and method for communicating with low density parity check codes
US9264182B22012-09-132016-02-16Novelsat Ltd.Iterative receiver loop
US8930789B1 *2013-01-232015-01-06Viasat, Inc.High-speed LDPC decoder
US9094132B12013-01-232015-07-28Viasat, Inc.High data rate optical transport network using 8-PSK
US9294131B22013-02-102016-03-22Hughes Network Systems, LlcApparatus and method for improved modulation and coding schemes for broadband satellite communications systems
US8887024B22013-02-102014-11-11Hughes Network Systems, LlcApparatus and method for improved modulation and coding schemes for broadband satellite communications systems
US9246634B22013-02-102016-01-26Hughes Network Systems, LlcApparatus and method for improved modulation and coding schemes for broadband satellite communications systems
US9203431B22013-03-152015-12-01Hughes Networks Systems, LlcLow density parity check (LDPC) encoding and decoding for small terminal applications

CN106330200B *2015-07-092019-09-10华邦电子股份有限公司It is implemented in the ldpc decoding method of flux graphics processor
DE102016222007B42016-11-092018-07-12Deutsches Zentrum für Luft- und Raumfahrt e.V.Method for transmitting data
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
KR20170103780A2017-09-13Data processing device and data processing method
US9148177B22015-09-29Method and system for error correction in transmitting data using low complexity systematic encoder
US10361724B22019-07-23Data processing device and data processing method for improving resistance to error of data
US20160204901A12016-07-14Systems and methods for ldpc coded modulation
JP5976960B22016-08-24Design for lifted LDPC codes with high parallelism, low error floor, and simple coding principles
EP2329602B12015-03-11Method and apparatus for adapting a bit interleaver to ldpc codes and modulations under awgn channel conditions using binary erasure surrogate channels
US8402337B22013-03-19Data processing apparatus and data processing method as well as encoding apparatus and encoding method
US7373581B22008-05-13Device, program, and method for decoding LDPC codes
JP5648852B22015-01-07Data processing apparatus and data processing method
CN100440736C2008-12-03Method and system for routing in low density parity check (ldpc) decoders
EP1528686B12006-11-22Method and apparatus for encoding long and short block length low density parity check (LDPC) codes
US9094043B22015-07-28Data processing device and data processing method
DE602005002815T22008-07-17Method and apparatus for generating a low-density parity-check (LDPC) code
CA2623465C2014-01-21A set of irregular ldpc codes with random structure and low encoding complexity

EP1513258B12008-02-20Method and system for encoding short block length low density parity check (LDPC) codes
US6950461B22005-09-27Modems utilizing low density parity check codes
JP4361924B22009-11-11Apparatus and method for data transmission / data reception in a communication system using a structured low density parity check code
US8341492B22012-12-25Quasi-cyclic LDPC (low density parity check) code construction
JP4555334B22010-09-29Apparatus and method for encoding / decoding block low density parity check code having variable block length
KR101492595B12015-02-11Memory-efficient ldpc decoding
US8392789B22013-03-05Method and system for decoding low density parity check codes
JP4163023B22008-10-08Parity check matrix generation method and parity check matrix generation apparatus
US9184768B22015-11-10Data-processing device and data-processing method
ES2529182T32015-02-17Structured LDPC design with grouping vector rows
JP4836962B22011-12-14Multilevel low density parity check coded modulation

Priority And Related Applications

Priority Applications (2)

ApplicationPriority dateFiling dateTitle
US48497403P2003-07-032003-07-03*US Provisional Application*
US60/484,9742003-07-03

Legal Events

DateCodeTitleDescription
2004-07-02A201Request for examination
2006-01-26E902Notification of reason for refusal
2006-09-26E701Decision to grant or registration of patent right
2006-10-09GRNTWritten decision to grant
2012-09-28FPAYAnnual fee payment
Payment date: 20121002

Year of fee payment: 7

2013-09-26FPAYAnnual fee payment
Payment date: 20130926

Year of fee payment: 8

2014-09-26FPAYAnnual fee payment
Payment date: 20140926

Year of fee payment: 9

2015-10-01FPAYAnnual fee payment
Payment date: 20151001

Year of fee payment: 10

2016-09-29FPAYAnnual fee payment
Payment date: 20160929

Year of fee payment: 11

2018-09-28FPAYAnnual fee payment
Payment date: 20180928

Year of fee payment: 13

2019-09-23FPAYAnnual fee payment
Payment date: 20190924

Year of fee payment: 14

(19)대한민국특허청(KR)
(12) 등록특허공보(B1)

(51) 。Int. Cl.
H03M 13/11 (2006.01)

(45) 공고일자 2006년10월16일
(11) 등록번호 10-0634250
(24) 등록일자 2006년10월09일

(21) 출원번호 10-2004-0051588
(22) 출원일자 2004년07월02일
(65) 공개번호 10-2005-0004118
(43) 공개일자 2005년01월12일

(30) 우선권주장 60/484,974 2003년07월03일 미국(US)

(73) 특허권자 더 디렉터브 그룹, 인크.
미국 90245 캘리포니아주 이엘 세군도 이. 임페리얼 하이웨이 2250

(72) 발명자 선,펭-웬
미국 20874 메릴랜드주 지밴타운 헤트릿지 드라이브 17904

에로즈,투스타파
미국 20874 메릴랜드주 지밴타운 인디안 그래스 드라이브 17007

리,린-난
미국 20854 메릴랜드주 포토맥 플라워 게이트 테라스 10004

(74) 대리인 주성민
백만기
이중희

심사관 : 송평중

(54) 병렬 디코딩이 가능한 저밀도 패리티 검사 코드를생성하기 위한 방법 및 시스템

요약

저밀도 패리티 검사(LDPC) 코드를 효율적으로 디코딩하는 방식이 제공된다. LDPC 디코더(305)는 꾸러미 메모리 구조(901, 1101)를 허용하는 특수의 병렬 디코딩 가능 조건을 만족하는 맵핑된 매트릭스를 저장하는 메모리(901, 1101)를 포함한다. 또한, 디코더(305)는 LDPC 코드를 디코딩하는 저장된 맵핑된 매트릭스로부터 에지 값을 액세스하는 병렬 프로세서(903, 1103)를 포함한다. 상기 방법은 위성 방송 시스템에 대한 특수 적용성을 갖는다.

대표도

도 8

색인어

저밀도 패리티 검사(LDPC) 코드, 맵핑된 매트릭스, 병렬 프로세서,

명세서

도면의 간단한 설명

도 1은 본 발명의 일 실시예에 따라 저밀도 패리티 검사(LDPC) 코드를 이용하도록 구성된 통신 시스템.

도 2는 도 1의 시스템의 예시적인 송신기.

도 3은 도 1의 시스템의 예시적인 수신기.

도 4는 본 발명의 일 실시예에 따른 저밀도 패리티 검사 매트릭스;

도 5는 도 4의 매트릭스의 LDPC 코드의 이분 그래프.

도 6은 본 발명의 일 실시예에 따른 도 3의 LDPC 디코더의 동작 순서도.

도 7은 본 발명의 일 실시예에 따른 LDPC 디코더의 컴포넌트들.

도 8은 본 발명의 다양한 실시예에 따른 효율적인 병렬 디코딩을 위한 압축된 매트릭스를 생성하는 프로세스의 순서도.

도 9는 본 발명의 일 실시예에 따른 병렬 디코딩을 지원하기 위한 여러 값들의 압축된 매트릭스를 저장하는 메모리를 가진 LDPC 디코더.

도 10은 본 발명의 다양한 실시예에 따른 병렬 디코딩을 수용하기 위해 도출된 압축된 매트릭스를 생성하는 프로세스의 순서도.

도 11은 본 발명의 일 실시예에 따른 병렬 디코딩을 지원하기 위한 여러 값들의 도출된 압축된 매트릭스를 저장하는 메모리를 가진 LDPC 디코더.

도 12는 본 발명의 실시예들에 따른 LDPC 코드의 인코딩 및 디코딩의 프로세스를 행할 수 있는 컴퓨터 시스템.

<도면의 주요 부분에 대한 부호의 설명>

101: 송신기

102: 채널

103: 수신기

200: 송신기

201: 소스

203: LDPC 인코더

205: 변조기

300: 수신기

301: 복조기

305: LDPC 디코더

307: 비트 메트릭 생성기

발명의 상세한 설명

발명의 목적

발명이 속하는 기술 및 그 분야의 종래기술

<관련 출원>

본 출원은 2003년 7월 3일자로 출원된 발명의 명칭이 "일반적인 병렬 디코딩이 가능한 LDPC 코드"인 미국 특허 가출원 번호 60/484,974와 관련된 것으로서 35 U.S.C. 119(e)조 하에서 그의 우선일을 주장하며, 그 전체 내용은 본 명세서에 한 영되어 있다.

본 발명은 통신 시스템에 관한 것으로서, 보다 상세하게는 코딩된 시스템에 관한 것이다.

통신 시스템은 잡음이 많은 통신 채널을 통한 신뢰성 있는 통신을 보장하기 위해 코딩을 이용한다. 통신 채널은 소정의 신호 대 잡음비(SNR)에서의 심볼당 비트로 표현될 수 있는 고정 용량을 갖는데, 이는 샤는 한계로 알려진 이론적인 상한을 정의한다. 결과적으로, 코딩 설계는 샤는 한계에 가까운 속도를 달성하는 것을 목표로 하여 왔다. 샤는 한계에 가까운 코드의 한 종류는 저밀도 패리티 검사(LDPC) 코드이다.

전통적으로 LDPC 코드는 많은 결점으로 인해 널리 보급되지 못했다. 결점 중 하나는 LDPC 인코딩 기술이 매우 복잡하다는 것이다. LDPC 코드를 그 생성자 매트릭스를 이용하여 인코딩하기 위해서는 매우 큰 넌-스파스(non-sparse) 매트릭스를 저장해야 한다. 또한, LDPC 코드는 큰 블록의 유효할 것을 요구하며, 결과적으로 LDPC 코드의 패리티 검사 매트릭스가 스파스(sparse)인 경우에도 매트릭스를 저장하는 데 문제가 발생한다.

구현에 있어서도 많은 도전에 직면하게 된다. 예컨대, 스토리지는 LDPC 코드가 실제로 널리 사용되지 못한 중요한 이유이다. LDPC 디코딩에 있어 주요한 이슈 중 하나는 메모리의 제공이다. 메모리 크기가 클수록 비트당 비용이 적다는 사실을 인식한다면, LDPC 디코더에서의 대량의 액기 값들에 대한 효율적인 구리의 메모리 구조(lumped memory structure)를 적용하는 코드 아키텍처에 대한 연구가 필요하다.

또한, LDPC 코드 구현에 있어서의 주요 과제는 디코더 내의 여러 처리 엔진들(노드들) 간의 접속 네트워크를 어떻게 이룰 것인가 하는 것이다. 또한, 디코딩 프로세스에서의 연산 부하, 특히 검사 노드 연산은 문제를 일으킨다.

일반적으로, 설정된 수의 병렬 엔진에 대응하는 코드는 효율적인 메모리 아키텍처를 유지하기 위하여 다른 수의 병렬 엔진을 수용하도록 재구성될 수 없는 것으로 인식되고 있다. 다른 애플리케이션은 다른 디코딩 속도를 요구하기 때문에 이러한 비유연성은 심각한 결점을 나타낸다. 또한, 이러한 제한은 반도체 성능에서의 향상을 수용하고 이용하는 능력을 방해한다. 예컨대, 처리 능력이 증가함에 따라 주어진 디코딩 속도에 필요한 병렬 엔진의 수가 감소해야 하지만, 고정된 수의 엔진의 설계는 사용되는 프로세서의 수에 있어서의 피접적인 감소를 허용하지 않는다.

따라서, 효율적인 디코딩 프로세스를 이용하는 LDPC 통신 시스템이 필요하다. 또한, LDPC 코딩을 구현하기 위한 스토리지 요건을 최소화할 필요가 있다. 미래의 기술 발달을 쉽게 수용할 수 있는 디코딩 스킴도 필요하다.

발명이 이루고자 하는 기술적 과제

본 발명은 이러한 과제를 해결하고자 하는 것으로서, 구조화된 저밀도 패리티 검사(LDPC) 코드를 디코딩하기 위한 해결책을 제공한다. 패리티 검사 매트릭스에 대응하는 컴팩된 매트릭스를 생성하기 위하여 병렬 디코딩이 가능한 코드에 대한 기준이 설정된다. 제1 기준은 컴팩된 매트릭스의 동일 행에 있는 임의의 다른 두 엔트리의 어느 것도 동일 비트 노드를 또는 동일 검사 노드들에 접속되지 않는다는 것이다. 동등로, $BG_i(i=0,1,\ldots,m-1)$로 표시되는 비트 노드 그룹이 i 번째 행의 액기들에 접속된 모든 비트 노드를 나타내는 경우, $i\neq j$에 대해, BG_i 및 BG_j은 어떠한 공통 노드도 갖지 않거나 동일하다는 것이 확실하다. 이러한 기준은 또한 검사 노드에도 적용되며, 이에 따라 $CG_i(i=0,1,\ldots,m-1)$로 표시되는 검사 노드 그룹은 i 번째

행의 에지들에 접속된 모든 검사 노드를 나타낸다. $i \neq j$에 대해, CG_i 및 CG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것이 검증된다. 이러한 기준은 LDPC 코드의 병렬 디코딩 구조를 이용하고 효율적인 메모리 아키텍처를 제공하는 LDPC 코드를 도출한다. n개의 프로세서보다 적은 프로세서(즉, m개의 프로세서, m은 n의 정수 프랙션(integer fraction)임)에 대한 효율적인 병렬 처리를 위한 3가지 조건을 유지하기 위하여 $n \times n$ 필드 균형 매트릭스인 블록 매트릭스 A_{ij}를 이용하여 패리티 검사 매트릭스가 구성된다. 블록 매트릭스는 행렬 교환에 의해 $m \times m$ 필드 균형 매트릭스인 B_{ij} 엔트리를 갖는 매트릭스로 분해된다. 이분 그래프에서, 도출된 코드는 랜덤된 매트릭스에서 소정의 에지들을 삭제함으로써 얻어질 수 있다. 이러한 구성에서 m개의 병렬 엔진이 사용될 수 있고, 따라서 구현의 유연성을 향상시킬 수 있게 된다.

본 발명의 일 실시예의 한 특징에 따르면, 저밀도 패리티 검사(LDPC) 코드의 디코딩을 지원하는 방법이 개시된다. 이 방법은 LDPC 코드와 관련된 패리티 검사 매트릭스에 기초하여 랜덤된 매트릭스를 구성하는 단계를 포함한다. 랜덤된 매트릭스는 병렬로 동작하는 복수의 프로세서에 의해 액세스 가능한 구조의 메모리 구조를 허용하기 위한 다수의 병렬 디코딩 가능 조건을 만족한다. 또한, 이 방법은 프로세서들에 의해 LDPC 코드를 디코딩하기 위해 메모리에 랜덤된 매트릭스를 저장하는 단계를 포함한다.

본 발명의 일 실시예의 다른 특징에 따르면, 저밀도 패리티 검사 코드의 디코딩을 위한 디코딩 장치가 개시된다. 이 장치는 구조의 메모리 구조를 허용하기 위한 다수의 병렬 디코딩 가능 조건을 만족하는 랜덤된 매트릭스를 저장하도록 구성된 메모리를 포함한다. 또한, 이 장치는 병렬로 동작하는 복수의 프로세스를 포함하는데, 이 프로세서들은 메모리에 액세스하여 LDPC 코드를 디코딩한다.

본 발명의 일 실시예의 또 다른 특징에 따르면, 저밀도 패리티 검사 코드의 디코딩을 지원하기 위한 방법이 개시된다. 이 방법은 LDPC 코드에 대응하는 패리티 검사 매트릭스에 기초하여 랜덤된 매트릭스를 초기화하는 단계를 포함한다. 이 방법은 또한 랜덤된 매트릭스의 동일 행에 있는 임의의 다른 두 엔트리의 어느 것도 동일 비트 노드들 또는 동일 검사 노드들에 접속되지 않는 것으로 결정하는 단계를 포함한다. 비트 노드 그룹(BG_i)이 i 번째 행의 에지들에 접속된 모든 비트 노드들을 나타내는 경우, 이 방법은, $i \neq j$에 대해, BG_i 및 BG_j가 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계를 포함한다. 또한, 검사 노드 그룹 CG_i가 i 번째 행의 에지들에 접속된 모든 검사 노드를 나타내는 경우, 이 방법은, $i \neq j$에 대해, CG_i 및 CG_j가 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계를 포함한다. 또한, 이 방법은 병렬로 동작하는 복수의 프로세서에 의해 액세스 가능한 메모리로 랜덤된 매트릭스를 출력하여 저장하여, 저장된 랜덤된 매트릭스에 따라 LDPC 코드를 디코딩하는 단계를 포함한다.

본 발명의 다른 목적 및 이점은 본 발명을 실시하기 위해 고안된 최상 모드를 포함한 다수의 특정 실시예 및 구현예를 참조하여 아래의 상세한 설명으로부터 명확히 알 수 있을 것이다. 또한, 본 발명은 다른 실시예도 가능할 것이며, 본 발명의 사상 및 범위를 벗어나지 않고 다양한 수정이 가능할 것이다. 따라서, 도면 및 상세한 설명은 예시적인 것일 뿐 제한적인 것이 아니다.

발명의 구성 및 작용

본 발명은 동일 요소에 동일 부호가 부여된 첨부된 도면들을 참조하여 제한적이 아니라 예시적으로 설명된다.

구조화된 저밀도 패리티 검사 코드를 효율적으로 디코딩하기 위한 시스템, 방법 및 소프트웨어가 설명된다. 아래의 설명에서는 설명의 목적으로 본 발명의 철저한 이해를 제공하기 위하여 많은 특정 사항이 설명된다. 그러나, 당해 분야의 기술자는 본 발명이 이러한 특정 사항이 아니고도 또는 동가적인 구성으로 실시될 수 있다는 것을 알 것이다. 여러 예에서 본 발명을 불필요하게 모호하게 하지 않기 위해 공지된 구조 및 장치는 블록도 형태로 도시된다.

도 1은 본 발명의 일 실시예에 따른 저밀도 패리티 검사 코드를 이용하도록 구성된 통신 시스템을 나타낸다. 디지털 통신 시스템(100)은 통신 채널(103)을 통해 수신기(105)로 전송될 신호 파형을 생성하는 송신기(101)를 포함한다. 이러한 개별 통신 시스템(100)에 있어서, 송신기(101)는 가능한 메시지들의 개별 세트를 생성하는 메시지 소스를 갖추고 있는데, 이들 가능한 메시지의 각각은 대응하는 신호 파형을 갖는다. 이들 신호 파형은 통신 채널(103)에 의해 감쇄되거나 변형된다. 잡음 채널(103)을 해결하기 위하여 LDPC 코드가 사용된다.

송신기(101)에 의해 생성되는 LDPC 코드는 여러한 성능 손실을 유발하지 않고 고속 구현을 가능하게 한다. 이러한 LDPC 코드는 (터보 코드와 달리) 가산, 비교 및 룩업 테이블 등과 같은 간단한 연산을 유익하게 포함하는 병렬 디코딩 알고리즘을 갖

는다. 이러한 병렬 구조는 메모리 요건을 최소화하여 다양한 수의 병렬 엔진(즉 프로세서)을 이용하는 데 있어서 유연성을 향상시키기 위해 이용된다. 이에 대해서는 도 7-11을 참조하여 후술된다. 더욱이, 주의 깊게 설계된 LDPC 코드는 관심 범위에서 어떠한 에러 플로어 표시를 나타내지 않는다.

본 발명의 일 실시예에 따르면, 송신기(101)는 디코딩 동안에 효율적인 메모리 액세스를 용이하게 하는 패리티 검사 매트릭스에 기초하여 LDPC 코드를 생성하여 수신기(105)와 통신한다.

도 2는 도 1의 시스템 내의 예시적인 송신기의 도면이다. 송신기(200)에는, 정보 소스(201)로부터의 입력을 받아, 수신기(105)에서의 에러 정정 프로세스에 적합한 리던던시를 갖는 코딩된 스트림을 출력하는 LDPC 인코더(203)가 장착된다. 정보 소스(201)는 이산 알파벳 X로부터 k개의 신호를 생성한다. LDPC 코드는 패리티 검사 매트릭스를 이용하여 지정된다.

변조기(205)는 인코더(203)로부터의 인코딩된 메시지를 송신 안테나(207)에 송신되는 신호 파형으로 맵핑하며, 송신 안테나(207)는 이러한 파형을 통신 채널(103)을 통해 방출한다. 이에 따라, 인코딩된 메시지가 변조되어 송신 안테나(207)에 분산된다. 이하에 설명되는 바와 같이, 송신 안테나(207)로부터의 송신은 수신기로 전파된다.

도 3은 도 1의 시스템 내의 예시적인 수신기의 도면이다. 수신측에서, 수신기(300)는 송신기(200)로부터 수신된 신호의 복조를 수행하는 복조기(301)를 포함한다. 이러한 신호는 복조를 위해 수신 안테나(303)에서 수신된다. 복조 후, 수신된 신호는 디코더(305)로 전송되며, 디코더(305)는 비트 매트릭 생성기(307)와 관련하여 메시지 X'를 생성함으로써 원시 소스 메시지의 재구성을 시도한다. 비트 매트릭 생성기(307)는 디코딩 프로세스 동안 디코더와 순방향 및 역방향으로(반복적으로) 정보를 교환할 수 있다. 이러한 디코딩 방식은 2003년 7월 3일자로 출원되어 공동 계류 중인 "저밀도 패리티 검사(LDPC) 디코더 내의 라우팅을 위한 방법 및 시스템(Method and System for Routing in Low Density Parity Check Decoders)"(출원 번호 10/613,824; 대리인 정리번호 PD-203009)에 보다 더 상세하게 설명되어 있으며, 그 전체 내용이 본 명세서에 참조로서 포함된다. 본 발명에 의해 제공되는 이점을 이해하기 위하여, 도 4에 설명되는 바와 같이 LDPC 코드가 어떻게 생성되는지를 고찰하는 것이 바람직하다.

도 4는 본 발명의 실시예에 따른 스파스 패리티 검사 매트릭스의 도면이다. LDPC 코드는 스파스 패리티 검사 매트릭스 $H_{(n-k)xn}$을 갖는 긴 선형 블록 코드이다. 전형적으로, 블록 길이 n은 수천 내지 수만 비트의 범위 내에 있다. 예를 들어, 길이 n=8이고 레이트 1/2인 LDPC 코드를 위한 패리티 검사 매트릭스가 도 4에 도시되어 있다. 동일한 코드가 도 5의 이분 그래프에 의해 등가적으로 표현될 수 있다.

도 5는 도 4의 매트릭스의 LDPC 코드의 이분 그래프의 도면이다. 패리티 검사식은, 각각의 검사 노드에 대하여 모든 인접 비트 노드의 합계{GF(Galois Field)(2) 상에서의}가 0임을 암시한다. 도면에 나타난 바와 같이, 비트 노드는 그래프의 좌측을 점유하며, 미리 결정된 관계에 따라 하나 이상의 검사 노드와 연관된다. 예를 들어, 검사 노드 m_1에 대응하여, 비트 노드와 관련하여 $n_1 + n_4 + n_5 + n_8 = 0$의 표현이 존재한다.

LDPC 디코더(305)는 메시지 패싱 디코더(message passing decoder)를 고려하며, 그에 의해 비트 노드의 값을 찾게 된다. 이러한 패스코를 달성하기 위하여, 비트 노드 및 검사 노드는 반복적으로 서로 통신한다. 이러한 통신의 본질에 대해서는 이하에 설명된다.

검사 노드로부터 비트 노드로, 각각의 검사 노드는 하나의 인접 비트 노드에, 다른 인접 비트 노드로부터의 정보에 기초하는 그 비트 노드의 값에 관한 추정치("평가"(opinion))를 제공한다. 예를 들어, 상기의 예에서, n_4, n_5 및 n_8의 합계가 m_1에게 0으로 "보이는" 경우, m_1은 n_1의 값이 0인 것으로 추정된다는 것을 n_1에게 알려준다 ($n_1 + n_4 + n_5 + n_8 = 0$ 이므로). 그렇지 않으면, m_1은 n_1의 값이 1로 추정된다는 것을 n_1에게 알려준다. 또한, 소프트 결정 디코딩에 대하여, 신뢰도 측정이 추가된다.

검사 노드로부터 비트 노드로, 각각의 비트 노드는 하나의 인접 검사 노드에, 다른 인접 검사 노드로부터의 피드백에 기초하는 그 검사 노드의 값에 관한 추정치를 공제한다. 상기 예에서, n_1은 단 2개의 인접 검사 노드 m_1 및 m_2만을 갖는다. m_2로부터 n_1으로의 피드백이 n_1의 값이 아마 0일 것임을 나타내는 경우, n_1은 m_1에게 n_1이 소유하는 값의 추정치가 0임을 알릴 것이다. 비트 노드가 2보다 많은 인접 검사 노드를 갖는 경우, 비트 노드는 자신과 통신하는 검사 노드에 결정을 보고하

기 전에, 다른 인접 검사 노드로부터의 피드백에 따라 결정(소프트 결정)을 수행한다. 상기 프로세스는, 모든 비트 노드가 올바른 것으로 간주될 때까지(즉, 모든 패리티 검사식이 만족될 때까지), 또는 미리 결정된 최대 반복 횟수에 도달되어 디코딩 실패가 선언될 때까지 반복된다.

도 6은 본 발명의 실시예에 따른 LDPC 디코더의 동작의 순서도이다. LDPC의 가장 두드러진 이점 중 하나는 병렬 디코딩에 대한 가능성인 것으로 인식된다. 이것은 LDPC 코드의 이분 그래프 및 디코딩 알고리즘을 조사함으로써 분명해진다. 명확하게, 비트 노드(또는 검사 노드)를 프로세스할 때, 이러한 프로세스 동안에는 중간 메시지 교환이 없기 때문에, 모든 비트 노드(또는 검사 노드)가 동시에 프로세스될 수 있다. 각각의 비트 노드 j에 대하여, 특정 비트의 로그 가능성 비율(log likelihood ratio)이 논리 1이 아닌 논리 0임을 표현하는 수신 값 u_j이 존재하며, 이것은 "채널 값"으로 참조된다. 계속 돌아, LDPC 코드를 디코딩하기 위한 메시지 파싱 알고리즘(message parsing algorithm)이 설명된다.

단계(601)에서, 디코더(305)의 초기화가 수행되고, 이에 의하여 이분 그래프(도 5)의 각각의 에지에 값이 할당된다. 이러한 값은 해당 에지가 접속되는 비트 노드에 연관된 채널 값과 동일하다. 다음으로, 단계(603)를 통하여 검사 노드가 갱신된다. 즉, 검사 노드 프로세스의 라지막에, 모든 에지 값이 갱신된다. $e_1, e_2, ..., e_v$가 특정 검사 노드에 접속된 에지 전부의 값을 모두 나타낸다고 하면, 갱신된 에지 값은 다음과 같다.

$$e_i = g(e_1, e_2, ..., e_{i-1}, e_{i+1}, ...e_v) : i = 1, 2, ...v$$

여기에서, $g(a,b) = \ln\frac{1+e^{a+b}}{e^a+e^b}$ 이다. 함수 g(a, b, c) = g(g(a, b), c)는 회귀적으로 계산될 수 있다.

모든 검사 노드가 처리되고 난 후, 단계(605)에서와 같이, 갱신된 에지 값으로 비트 노드가 프로세스된다. 비트 노드 프로세스의 목표는 에지 값을 더 갱신하는 것이다. $e_1, e_2, ..., e_v$가 이러한 특정 비트 노드에 접속된 모든 에지의 에지 값을 나타낸다고 가정하면, 갱신된 에지 값은 다음과 같다.

$$e_i = e_1 + e_2 + ... + e_{i-1} + e_{i+1} + ...e_v$$

프로세스는 단계(607)를 통하여 사후 확률 정보(a posteriori probability information)를 출력하고, 패리티 검사식이 만족되거나 또는 구성가능한 최대 반복수에 도달되었는지를 결정한다(단계(609)). 이러한 식들이 만족되지 않는 경우, 단계(603-607)가 반복적으로 실행된다. 최종 반복시에, 하나의 비트 노드에 연관된 모든 에지 값이 합산되고, 단계(611)를 통하여 채널 값과 더 합산되어 최종 결정 메트릭을 생성한다.

LDPC 디코더(305)는 고도의 병렬 시스템으로서 구현될 수 있으므로, 검사 노드와 비트 노드 간의 상호 접속을 실현하기 위한 방식으로서, (1) 완전 병렬 방식 및 (2) 부분 병렬 방식의 2가지 방식이 존재한다. 완전 병렬 아키텍처에서, 노드 및 그들의 상호접속 전부가 물리적으로 구현된다. 이러한 아키텍처의 이점은 속도이다. 그러나, 완전 병렬 아키텍처에서는, 노드 및 그들의 상호접속 전부를 실현하는 것으로 인해 훨씬 큰 복잡성을 야기한다. 그러므로, 완전 병렬 아키텍처에서는, 복잡성을 감소시키기 위하여 블록 크기가 더 작아질 것이 요구된다.

LDPC 코드를 구현하기 위한 2번째 방식은, 노드의 총 개수의 부분집합만을 물리적으로 실현하고, 이러한 제한된 개수의 "물리적" 노드를 사용하여 코드의 "기능적" 노드 전부를 프로세스하는 것이다. LDPC 디코더(305) 동작이 매우 간단해질 수 있고 병렬로 수행될 수 있긴 하지만, "무작위로" 분산된 비트 노드와 검사 노드 간의 통신을 어떻게 설정할 것인가와 관련한 또 다른 설계상의 문제점이 있다.

도 7은 본 발명의 실시예에 따른 LDPC 디코더의 컴포넌트의 도면이다. 디코더(305)는 병렬 프로세서를 사용하여 도 7의 디코딩 프로세스를 실행한다. 도시된 바와 같이, 디코더(305)는 비트 노드 프로세서(701) 및 검사 노드 프로세서(703)를 포함하며, 이들은 각각 비트 노드 및 검사 노드 동작을 수행한다. 비트 노드 또는 검사 노드 갱신의 중간 에지 값은 에지 메모리(705)에 버퍼링된다. 또한, 하나의 비트 노드 프로세스 동안 새로운 에지 값을 생성하는 데에 있어서의 부분 합계는, 진행적으로 프로세서에 국부적으로 버퍼링된다. 채널 메모리(707)는 채널로부터의 신뢰도 정보를 유지한다. 도면에서 볼 수 있는 바와 같이, 채널 신뢰도 메모리(707)는 비트 노드 프로세스에서만 사용된다.

디코더(305)의 설계는 병렬 프로세서의 수 대 디코딩의 속도, 및 메모리 사이즈와 같은 다수의 엔지니어링 상충관계를 수반한다. 하나의 극단적인 예에서, 하나의 프로세서를 사용하여 하나의 노드로부터 다른 노드로 순차적으로 전체 코드를 디

코딩할 수 있다. 그러나, 전형적인 양호한 LDPC 코드에 대한 다수의 노드 및 다수의 반복으로 인해, 이러한 방식은 거의 유용하지 않다. 다른 극단적인 예는 노드의 개수만큼의 프로세서를 사용하는 것이다. 이에 따르면, 최고의 디코딩 속도를 얻을 수 있지만, 불행하게도 비용도 최고가 된다. 실용적인 디코더는 이러한 양 극단의 방식 사이에서 설계된다.

명백하게, 디코딩 속도와 비트 노드 및/또는 검사 노드 프로세서의 개수 간에 상충 관계가 존재한다. 비트 노드 프로세서(701) 및 검사 노드 프로세서(703)가 별개의 블록들로서 도시되어 있지만, 비트 노드 프로세서(701)가 검사 노드 프로세서(703)에 결합되어, 이들이 데이터 경로의 일부를 공유할 수 있음을 알 것이다. 다르게는, 프로세스 전에 에지 값의 변환을 수행함으로써, 비트 노드 프로세서(701)가 검사 노드 프로세서(703)와 동일해질 수 있다. 또한, 비트 노드 프로세서(701)의 개수가 검사 노드 프로세서(703)의 개수와 동일할 필요는 없음에 유의해야 한다.

또한, 효율적인 아키텍처는 병렬 프로세스 엔진의 수를 임의로 제공해서는 안 된다. 기술적 개선의 전보를 수용할 구현의 유연성을 제공하기 위하여, 병렬의 수에 어느 정도 자유를 갖는 것이 바람직하다. 예를 들어, 코드가 300개의 병렬 엔진을 갖도록 설계되는 경우, 집적 회로 기술이 발전함에 따라, 회로는 2배의 속도로 동작할 수 있다. 따라서, 동일한 코드가 180개의 병렬 엔진을 갖도록 재구성하면서, 동일한 디코딩 아키텍처를 유지하는 것이 바람직하다.

메모리 구조와 관련하여, LDPC 디코더(305)는 에지 메모리(705)를 프로세서(701, 703) 각각에 근접하여 배치한다. 레지스터 또는 RAM이 메모리(705)로 이용될 수 있다. 긴 코드에 대하여, 레지스터의 사용은 RAM보다 훨씬 더 고가이다. RAM과 관련하여, RAM의 크기가 증가할수록 비트당 비용이 감소하는 것으로 인식된다. 그러므로, 되도록 많은 RAM을 일괄하여, 모든 에지 값이 하나의 RAM에 기능적으로 입출되도록 하는 것이 바람직하다.

임의적인 LDPC 코드는 도 7의 디코더 아키텍처를 효율적으로 지원할 수 없는 것으로 인식된다. 이것은 표준 RAM 아키텍처를 실패품으로써 판명해준다. 이하의 관찰이 주목된다. RAM은 미리 정의된 폭 및 깊이를 갖는 메모리이며, 행 방식(row-wise)으로 판독 및 기입될 수 있다. 이것은, 하나의 액세스 사이클에서, 동일 행 내의 값만이 액세스될 수 있음을 의미한다. "랜덤"이라는 용어는 단지 각각의 액세스 사이클마다 임의적인 순서로 행들이 액세스될 수 있음을 의미하는 것이다. 그러므로, 효율성의 관점에서 볼 때, 임의의 주어진 액세스 사이클에서, 각각의 비트 노드 프로세서(701)(검사 노드측에서 프로세스하는 경우에는 검사 노드 프로세서(703))에는, 동일한 양의 에지 값 입력이 공급되어야만 한다. 따라서, 임의의 주어진 순간에, 효율적인 디코더 아키텍처는 프로세스 엔진으로의 공급이 중단되지 않게 해야 한다.

또한, 임의의 주어진 시간에, 특정 비트 노드(또는 검사 노드)에 하나의 프로세서를 전용으로 하는 것이 바람직하다. 즉, 프로세서는 하나의 비트 노드에 연관된 에지의 갱신을 끝내기 전에 다른 비트 노드(또는 검사 노드)로 절환해서는 안 된다.

이하에 설명되는 바와 같이, 본 발명은 도 7의 아키텍처의 다양한 특성을 설명하기 위하여, 병렬 디코딩 가능한 코드의 개념을 도입한다.

도 8은 본 발명의 다양한 실시예에 따른 효율적인 병렬 디코딩을 위한 맵핑된 매트릭스를 생성하는 프로세스의 순서도이다. 병렬 디코딩 가능한 코드는 효율적인 메모리 구조를 나타내도록 정의된다. 패리티 검사 매트릭스의 0이 아닌 항목을 다수의 정의된 속성을 만족시키는 m×n의 매트릭스(맵핑된 매트릭스)로 맵핑시키는 일대일 맵핑이 존재하는 경우, LDPC 코드는 우리의 메모리를 갖는 n개의 병렬 엔진으로 디코딩될 수 있다. 이러한 맵핑된 매트릭스를 생성하기 위한 프로세스가 이하에 설명된다.

단계(801)에서, 패리티 검사 매트릭스에 기초하여, 맵핑된 매트릭스가 생성된다. 그리고나서, 효율적인 병렬 디코딩을 위한 다양한 조건 또는 속성을 보증하기 위하여, 이러한 맵핑된 매트릭스가 검증된다. 단계(803)를 통하여, 맵핑된 매트릭스의 동일 행 내의 어떠한 2개의 상이한 항목도 동일한 비트 노드 또는 동일한 검사 노드에 접속되지 않도록, 매트릭스가 검증된다. BG_i로 표시되는 비트 노드 그룹 $i(i = 0, 1, ...m-1)$가 i번째 행에 있는 에지에 접속되는 모든 비트 노드라고 하면, $i \neq j$에 대해, 프로세스는 BG_i 및 BG_j가 어떠한 공통 노드도 갖지 않거나, 또는 동일한 것을 보증한다(단계(805)). 마찬가지로, 단계(807)를 통하여, CG_i로 표시되는 검사 노드 그룹 $i(i = 0, 1, ...m-1)$가 i번째 행에 있는 에지에 접속되는 모든 검사 노드라고 하면, $i \neq j$에 대하여, CG_i 및 CG_j가 어떠한 공통 노드도 갖지 않거나 또는 동일한 것으로 검증된다.

단계(803-807)의 3개의 조건은 효율적인 병렬 프로세스를 위하여 필요한 것이다. 단계(809)에서, 맵핑된 매트릭스가 출력되어 에지 메모리(705)에 저장된다(단계(811)). 예시적인 실시예에서, 맵핑된 매트릭스는 메모리 구조를 나타낸다. 이

하나 메모리는 물리적으로는 상이한 메모리 장치이고, 기능적으로는 맵핑된 매트릭스과 동가이다. 단계(803)의 속성은, 각각의 액세스 사이클에서, 각각의 프로세서(예를 들어, 프로세서(710, 703))가 각각의 액세스 사이클 동안 단 하나의 값만을 수신할 것임을 보증한다.

n개의 병렬 프로세스 엔진에서, 모든 프로세서에는 균등한 부하가 주어질 것이다. 제2 속성(단계(805))은 비트 노드들을 크기 n의 그룹들로 분할한다. 2개의 비트 노드는, 동일 체크로 맵핑된 노드들에 연관된 에지가 존재하는 한, 동일 그룹 내에 있다. 또한, 각각의 비트 노드에 대한 프로세스 동안 소정의 중간 결과가 능동적으로 프로세스되도록 유지할 필요가 있다. 따라서, 비트 노드(검사 노드)의 한 집합의 프로세스는 새로운 비트 노드(또는 검사 노드)를 프로세스하기 전에 완료되어야만 한다. 그리고, 이것은 단계(805 및 807)의 제2 및 제3 속성에 의해 보증된다.

도 9는 본 발명의 실시예에 따른 병렬 디코딩을 지원하기 위하여 에지 값의 맵핑된 매트릭스를 저장하는 메모리를 갖는 LDPC 디코더의 도면이다. m ×n의 맵핑된 매트릭스가 메모리(901) 내에 저장되며, 에지 값은 n 개의 병렬 프로세서(903)에 제공된다. 설명을 위하여, 맵핑된 매트릭스의 사용은 이하의 패리티 검사 매트릭스과 관련하여 설명된다.

$$\begin{bmatrix} 1 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 1 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 1 \\ 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 0 & 0 \end{bmatrix}$$

디코딩 동안, 패리티 매트릭스의 0이 아닌 항목들 각각은 에지 값에 연관된다. 이러한 에지 값은 다음과 같이 라벨링된다.

$$\begin{bmatrix} e_1 & e_2 & 0 & 0 & 0 & 0 & 0 & e_9 & 0 & 0 & e_{13} & 0 & 0 & 0 & 0 & 0 \\ 0 & e_3 & e_4 & 0 & 0 & 0 & e_{10} & 0 & 0 & 0 & 0 & e_{17} & 0 & 0 & 0 & 0 \\ 0 & 0 & e_5 & e_{11} & 0 & e_{14} & 0 & 0 & e_{18} & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ e_{21} & 0 & 0 & e_{12} & e_{22} & 0 & 0 & 0 & 0 & e_{23} & 0 & 0 & 0 & 0 & 0 & 0 \\ e_{24} & 0 & 0 & 0 & 0 & e_{25} & 0 & 0 & e_{26} & 0 & 0 & 0 & 0 & e_{27} & e_{28} & 0 \\ 0 & e_{29} & 0 & 0 & 0 & 0 & e_{30} & 0 & 0 & e_{31} & 0 & 0 & 0 & 0 & e_{32} & e_{33} \\ 0 & 0 & e_{34} & 0 & 0 & 0 & 0 & e_{35} & 0 & 0 & e_{36} & 0 & e_{37} & 0 & 0 & e_{38} \\ 0 & 0 & 0 & e_{39} & e_{40} & 0 & 0 & 0 & 0 & 0 & 0 & e_{41} & e_{42} & e_{43} & 0 & 0 \end{bmatrix}$$

에지 값은 이하의 9 ×4 매트릭스로 일대일 맵핑된다.

$$\begin{bmatrix} e_1 & e_2 & e_9 & e_{13} \\ e_3 & e_4 & e_{10} & e_{17} \\ e_5 & e_{11} & e_{14} & e_{18} \\ e_{21} & e_{12} & e_{22} & e_{23} \\ e_{24} & e_{25} & e_{26} & e_{27} \\ e_{28} & e_{29} & e_{30} & e_{31} \\ e_{32} & e_{33} & e_{34} & e_{35} \\ e_{36} & e_{37} & e_{38} & e_{39} \\ e_{40} & e_{41} & e_{42} & e_{43} \end{bmatrix}$$

위의 매트릭스는, 매트릭스가 메모리를 미러화(mirror)한다고 가정해서, 100% 사용하는 조밀 매트릭스 (dense matrix)이다. 다시 말하면, 메모리의 모든 셀들이 완전히 사용된다. 이 맵핑이, 위에 기술된 바대로, 효율적 병렬 디코딩을 위한 모든 세가지 특성들을 만족시킨다는 것이 쉽게 증명될 수 있다.

이 예에서, 비트 노드들은 각 그룹이 4 비트 노드들을 포함하는 4 그룹들로 분리되고, 비트 노드 i는 원서 패리티 검사 매트릭스 (original parity check matrix)의 열 i와 연계된 비트 노드이다. 이러한 비트 노드 그룹들은 {1,2,3,4},{5,6,7,8}, {9,10,11,12}, 및 {13,14,15,16}이다.

유사하게, 검사 노드 i는 원시 패리티 검사 매트릭스의 행 i와 연계되는 검사 노드이다. 검사 노드들은 2 그룹들로 분리된다: {1,2,3,4}, 및 {5,6,7,8}. 이 예에서, 4 병렬 프로세서들이 채용된다(n=4). 비트 노드 프로세싱에 대해, 제1 행은 4 병렬 프로세서들(903)로 입력되고, 중간 결과들은, 특정의 비트 노드 {1,2,3,4}와 연계된 메시지들을 갱신하기 위한 것이므로, 분리된 메모리에 생성되고 버퍼링될 것이다. 이것은 {1,2,3,4}의 모든 연계 메시지 값들이 처리된 후에 성취된다.

비트 노드 {1,2,3,4}와 연계되는 모든 에지 값들은 에지 매트릭스의 적은 3행들에 위치함이 주목된다. 그러므로, 다음 액세스 사이클에서, 제2 행 또는 제3 행은 액세스가 계속될 것이다. 프로세스가 임의의 다른 행에서 시작하면, 추가 중간 결과들은 버퍼링될 필요가 있을 것이다. 비트 노드 {1,2,3,4}와 연계되는 모든 행들이 처리되고 에지 값들이 갱신된 후에, 비트 노드들의 다른 그룹들이 처리된다.

유사하게, 위의 과정은 검사 노드들에서 수행된다. 즉, 검사 노드들은 그룹들로 처리되고, 이 그룹과 연계되는 모든 에지들이 갱신된 후에, 다음 그룹이 처리된다. 각 액세스 사이클에서, 에지 매트릭스로부터의 정확히 한 개 값이 현재 처리되고 있는 각 노드에 대해 갖고오게 되고, 그래서 프로세서들(903) 중의 어떤 것도 파두하(또는, 그 문제에 있어서, 적조한 사용)되지 않도록 한다.

효율적 병렬 처리를 (도8의 스텝들(803~807)에서 기재된 바와 같이) 위한 3 가지 조건들은 어느 정도 제한적일 수 있고, 무작위 LDPC 코드는 일반적으로 이러한 조건들을 만족하지 않을 것임이 인식된다. 이 제한은, 아래 설명된 대로, 랜덤된 매트릭스로부터 다른 매트릭스를 도출해서 해제될 수 있다.

도 10은, 본 발명의 다양한 실시예에 따른 병렬 디코딩을 수용하기 위해 도출된 랜덤된 매트릭스를 생성하는 프로세스의 흐름도이다. 도출된 랜덤된 매트릭스 생성을 더 잘 이해하기 위해, 특정 기술들 및 정리들이 아래와 같이 기술된다. n x n 매트릭스는, 매트릭스의 각 행 및 각 열이 정확히 0 이 아닌 엔트리들의 수와 동일하면, "밀도 균형 매트릭스 (dense balanced matrix)"이라고 이름지어 진다. 밀도 균형 매트릭스의 한 예는 올제로 (all-zero) 매트릭스이다; 이 매트릭스는 LDPC 코드 생성을 위해 널리 사용된다.

밀도 균형 매트릭스의 또 다른 예는 다음의 특성들을 갖는 매트릭스이다. $b_0b_1...b_{n-1}$이 바이너리 벡터이고, n는 n에 비교적 프라임 (prime)이라고 가정하면, i번째 행 및 j번째 열의 엔트리가 b_{ai+j}인 매트릭스는 밀도 균형 매트릭스이다. 특히, a=1일 때, 써큘런트 매트릭스(circulant matrix)로 귀결된다.

밀도 균형 매트릭스의 임의의 행 및/또는 열의 순열(permutation)은 밀도 균형 매트릭스로 귀결함이 주목된다. 그러나, 상이한 컴포넌트 매트릭스들은 다르게 교환될 수 있기 때문에, 기본적으로 다른 LDPC 코드들은 다른 교환들의 빌딩 블록(building block)들을 사용해서 귀결된다. 밀도 균형 매트릭스에서, 한 개의 행에서 0 이 아닌 엔트리들의 수가 k이면, 모든 0 이 아닌 엔트리들은 한 번에 한 행씩 한 개의 0 이 아닌 엔트리를 택해서 k x n 밀도 매트릭스로 랜덤될 (일대일) 수 있다.

다음 정리가 제안된다 (정리 1): LDPC 코드는,

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1n} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2n} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{k1} & A_{k2} & \cdots & \cdots & A_{kn} \end{bmatrix}$$

컴포넌트 블록 매트릭스 A_{ij}의 각각이 n x n 밀도 균형 매트릭스가 되도록, 블록 매트릭스에 등가인 교환인 LDPC 코드의 패리티 검사 매트릭스가 있을 때만 n 효율적 병렬 디코딩을 위한 모든 3 조건들을 만족한다. 그렇게 구성된 매트릭스는 일반 LDPC 코드를 생성할 필요가 없음을 주목해야 한다.

패리티 검사 매트릭스의 구조에 정리에 의해 부가된 주어진 제한에서, 일반 패리티 검사 매트릭스는 이러한 조건들을 만족하지 않을 것임은 가정하다. 그러므로, 효율적 메모리 액세스와 디코딩할 수 있기 위해 이 구조적 제한을 갖는 코드를 디자인할 필요가 있다.

n 병렬 엔진들과 효율적 병렬 디코딩할 수 있는 코드는 기타 수들에 대해 (즉, n 이외의 수의 엔진들과 함께) 효율적 병렬 디코딩할 수 없을 수도 있음이 인식된다. 이것에 기술된 대로, 프로세싱에서 미래의 전개들을 설명하기 위해서, 저병렬관계(less parallelism)가 미래 수용자들을 위해 바람직하다. 그러나, 효율적 병렬 프로세싱을 위한 모든 조건들은 저 병렬관계의 경우에서도 만족될 필요가 있다.

n 보다 적은 수의 프로세서들을 위해 효율적 병렬 프로세싱을 위한 3 조건들을 유지하기 위해, 또 다른 정리가 제안된다 (정리 2). 다음의 매트릭스가 패리티 검사 매트릭스라고 하고, 블록 매트릭스 A_{ij}의 각각은 n x n 밀도 균형 매트릭스들이라고 하자.

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1c} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2c} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{r1} & A_{r2} & \cdots & \cdots & A_{rc} \end{bmatrix}$$

A_{11}이, B_{ij}의 각각이 m x m 밀도 균형 매트릭스이고, 모든 다른 A_{ij}에 대해서, 블록 매트릭스들은 A_{11}과 동일한 행 및 열 교환 하에 블록 밀도 균형 서브매트릭스들로 분해될 수 있도록, 행 및 열 교환 하에서 한 개의 매트릭스로 분해될 수 있다면,

$$A_{11} = \begin{bmatrix} B_{11} & B_{12} & \cdots & \cdots & B_{1p} \\ B_{21} & B_{22} & \cdots & \cdots & B_{2p} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ B_{q1} & B_{q2} & \cdots & \cdots & B_{qp} \end{bmatrix}$$

LDPC 코드는 다음과 같이 정의된다:

$$\begin{bmatrix} A_{11} & A_{12} & \cdots & \cdots & A_{1c} \\ A_{21} & A_{22} & \cdots & \cdots & A_{2c} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ A_{r1} & A_{r2} & \cdots & \cdots & A_{rc} \end{bmatrix}$$

위의 매트릭스는 n 및 m 병렬관계로 효율적 병렬 프로세싱을 위한 조건들을 만족한다.

예로써, 다음이 정의된다: A_{ij}는 n x n 서큘런트 매트릭스들이고, m은 n의 임의의 정수 프랙션(fraction)이다. A_{ij}의 행 및 열들은 0,1,...,n으로 레벨화된다. 행 및 열 인덱스들이 상수 모듈로 m과 같은 모든 소자들로 구성되는 A_{ij}의 서브매트릭스는 또한 서큘런트이다. 결과적으로, 동일 행 및 열 교환 하에 모든 A_{ij}는 다시 서큘런트 서브매트릭스들로 분해 가능하다. 이것은, 컴포넌트 코드로서 A_{ij}들을 갖는 LDPC 코드는 n 및 m 병렬관계 모두와 함께 (여기서, m은 n의 임의의 정수 프랙션임) 효율적 병렬 프로세싱을 위한 조건들을 만족함을 함축한다.

패리티 검사 매트릭스 H를 갖는 주어진 임의의 LDPC 코드에서, 기타 LDPC 코드들은 H의 0 이 아닌 엔트리들의 일부를 0 들로 변환해서 H로부터 도출될 수 있다. 이 방식으로 얻어진 임의의 코드는 "원시 LDPC 코드의 도출 코드"라고 용어화된다. 스텝(1001)에서, n 병렬 디코딩되는 코드를 위한 맵핑되는 매트릭스는 수취된다. 원시 코드가 n 병렬 엔진 검사 매트릭스의 완전 조밀 메모리를 갖는 n 병렬 디코딩이라면, 임의의 도출 코드는 또한 감소된 밀도 메모리 매트릭스를 갖고 n 병렬 디코딩 가능할 수 있다.

이분 그래프(bipartite graph)의 관점에서, 도출 코드는 일부 에지들을 삭제해서 얻어질 수 있다 (스텝(1003)). 도출된 코드의 맵핑되는 매트릭스는 근본적으로 기본 코드의 맵핑된 매트릭스와 동일하다. 그러나, 삭제된 에지들에 대응하는 엔트리들은 "돈케어 (DON'T CARE)"로서 (즉, 널 (null) 값들) 표시될 수 있다. 그러므로, 이러한 메모리 셀들은 낭비되고, 액세스 사이클들은 널 값들을 가져오면서 확장된다. 그러나, 에지 삭제들의 숫자가 비교적 작은 한, 그것은 효율성에 크게 영향을 주지 않는다. 그런 도출 코드는, 쉬운 인코딩 및 향상된 거리 (distance) 특성들과 같은, 기타 바람직한 특성들을 나타낼 수 있다. 스텝(1005)에서, 도출된 맵핑된 매트릭스가 출력된다.

도11은, 본 발명의 실시예에 따른 병렬 디코딩을 지원하기 위한 에지 값들의 도움된 맵핑된 매트릭스를 저장하는 메모리를 갖는 LDPC 디코더를 나타내는 도면이다. 그 다음, 이 매트릭스는 m 프로세서들(1103)에 의해 액세스되기 위해 메모리(1101)에 저장된다. 이 배열에서, 효율적 메모리 구조를 유지하면서, 사용되는 병렬 연산들의 수에 있어서 매우 유연성이 있다.

도12는 본 발명에 따른 한 실시예가 구현될 수 있는 컴퓨터 시스템을 도시한다. 컴퓨터 시스템(1200)은 정보를 통신하기 위해 버스(1201) 또는 기타 통신 메카니즘, 및 정보를 처리하기 위해 버스(1201)와 결합된 프로세서(1203)를 포함한다. 컴퓨터 시스템(1200)은, 프로세서(1203)에 의해 실행되는 정보 및 명령들을 저장하기 위해 버스(1201)로 결합된, 랜덤 액세스 메모리 (RAM) 또는 기타 동적 저장 장치와 같은, 메인 메모리(1205)를 또한 포함한다. 메인 메모리(1205)는 프로세서(1203)에 의해 실행되는 명령들의 실행 동안 임시 변수들 또는 기타 중간 정보를 저장하기 위해 또한 사용될 수 있다. 컴퓨터 시스템(1200)은 프로세서(1203)을 위한 정적 정보 및 명령들을 저장하기 위해 버스(1201)로 연결된 읽기용 메모리 (ROM)(1207) 또는 기타 정적 저장 장치를 더 포함한다. 자기 디스크 또는 광 디스크와 같은, 저장 장치(1209)는 정보 및 명령들을 저장하기 위해 버스(1201)로 추가 연결된다.

컴퓨터 시스템(1200)은 컴퓨터 사용자에게 정보를 디스플레이하기 위해서, CRT (cathod ray tube), LCD (liquid crystal display), 액티브 매트릭스 디스플레이 (active matrix display), 또는 플라즈마 디스플레이 (plasma display)와 같은, 디스플레이(1211)로 버스(1201)를 통해 연결될 수 있다. 문자숫자 (alphanumeric) 및 기타 키들을 포함하는 키보드와 같은, 입력 디바이스(1213)는 프로세서(1203)로 정보 및 명령 선택들을 통신하기 위해 버스(1201)로 연결된다. 또 다른 타입의 사용자 입력 장치는, 프로세서(1203)로 지시 정보 및 명령 선택들을 통신하기 위해서 및 디스플레이(1211)에 커서 동작을 제어하기 위해서 마우스, 트랙볼, 또는 커서 방향 키들과 같은, 커서 제어(1215)이다.

본 발명의 한 실시예에 따라서, LDPC 코드들의 병렬 디코딩은 메인 메모리(1205)에 포함되는 명령들을 실행하는 프로세서(1203)에 응답해서 컴퓨터 시스템(1200)에 의해 제공된다. 그러한 명령들은, 저장 장치(1209)와 같은, 또 다른 컴퓨터 판독 가능 매체로부터 메인 메모리(1205)로 판독될 수 있다. 메인 메모리(1205)에 포함되는 명령들의 실행은 프로세서(1230)가 본 명세서에 기재된 프로세스 스텝들을 수행하도록 한다. 멀티 프로세싱에서 한 개 이상의 프로세서들은 메인 메모리(1205)에 포함된 명령들을 실행하기 위해 또한 채용될 수 있다. 또 다른 실시예들에서, 유선 회로부는 본 발명의 실시예를 구현하기 위해 소프트웨어 명령들을 대체하거나 조합해서 사용될 수 있다. 그러므로, 본 발명의 실시예들은 하드웨어 회로부 및 소프트웨어의 어떤 특정 조합에 제한되지 않는다.

컴퓨터 시스템(1200)은 버스(1201)로 연결된 통신 인터페이스(1217)를 또한 포함한다. 통신 인터페이스(1217)는 근거리 통신망(1221)에 연결된 통신망 링크(1219)로 연결된 양방향 데이타 통신을 제공한다. 예를 들어, 통신 인터페이스(1217)는, 대응하는 타입의 전화선으로 데이타 통신 연결을 제공하기 위한 디지털 가입자선(digital subscriber line, DSL) 카드 또는 모뎀, 통합 서비스 디지털 통신망(Intergrated services digital network, ISDN) 카드, 케이블 모뎀, 또는 전화 모뎀일 수 있다. 또 다른 예에서, 통신 인터페이스(1217)는 호환성 LAN으로 데이타 통신 연결을 제공하는 근거리 통신망(LAN) 카드 (예를 들어, 이써넷 (Ethernet) 또는 비동기 전송 모델 (Asynchronous Trasfer Model,ATM) 통신망)일 수 있다. 무선 링크들은 또한 구현될 수 있다. 임의의 그런 구현에서, 통신 인터페이스(1217)는, 다양한 타입들의 정보를 표절하는 디지털 데이타 스트림들을 운반하는 전기적, 전자기적, 또는 광 신호들을 주고 받는다. 게다가, 통신 인터페이스(1217)는, 범용 직렬 버스 (USB) 인터페이스, PCMCIA (Personal Computer Memory Card International Association) 인터페이스, 기타와 같은, 주변 인터페이스 장치들을 포함할 수 있다.

네트워크 링크(1219)는 한 개 이상의 통신망들을 통해 다른 데이타 장치들로 데이타 통신을 전형적으로 제공한다. 예를 들어, 네트워크 링크(1219)는, 네트워크(1225) (예를 들어, 원거리 통신망 (WAN)) 또는 "인터넷"으로서 현재 일반적으로 알려져 있는 범용 패킷 데이타 통신망) 또는 서비스 제공자에 의해 운영되는 데이타 장치로 연결되는, 호스트 컴퓨터(1223)로 근거리 통신망(1221)을 통해 연결을 제공할 수 있다. 근거리 통신망(1221) 및 통신망(1225) 모두는 정보 및 명령들을 전달하기 위해 전기적, 전자기적, 또는 광 신호들을 사용한다. 컴퓨터 시스템(1200)과 디지털 데이타를 통신하는, 다양한 통신망들을 통한 신호들 및 통신 링크(1219) 및 통신 인터페이스(1217)를 통한 신호들은, 정보 및 명령들을 포함하는 캐리어 웨이브 (carrier wave)들의 형태들의 예들이다.

컴퓨터 시스템(1200)은 통신망(들), 통신망 링크(1219), 및 통신 인터페이스(1217)를 통해, 프로그램 코드를 포함한 데이타와 메세지들을 주고 받을 수 있다. 인터넷 예에서, 서버 (도시 안됨)는 통신망(1225), 근거리 통신망(1221), 및 통신 인터페이스(1217)를 통해 본 발명의 한 실시예를 구현하기 위한 응용 프로그램에 속한 요구된 코드를 전송할 수 있다. 프로

세서(1203)는 수후 실행을 위해 저장 장치(129) 또는 비휘발성 저장 장치에 프로그램 코드를 수신하고 그리고/또는 저장하는 것을 예, 전송된 코드를 실행할 수 있다. 이 방식으로, 컴퓨터 시스템(1200)은 캐리어 웨이브 형태로 응용 프로그램 코드를 얻을 수 있다.

본 명세서에 사용된 것처럼 "컴퓨터 판독 가능 매체"라는 용어는 실행을 위해 프로세서(1203)에 명령들을 제공하는 역할을 하는 임의의 매체를 일컫는다. 그런 매체는, 비휘발성 매체들, 휘발성 매체들, 및 전송 매체들을 포함하지만 거기에 국한되지 않는 많은 형태들을 택할 수 있다. 예를 들어, 비휘발성 매체들은 저장 장치(1209)와 같은 광 또는 자기 디스크들을 포함한다. 휘발성 매체들은 메인 메모리(1205)와 같은 동적 메모리를 포함한다. 전송 매체들은, 버스(1201)를 포함하는 유선들을 포함하는 동축 케이블들, 납선 및 광섬유를 포함한다. 전송 매체들은, RF (Radio Frequency) 및 적외선 (Infrared,IR) 데이터 통신 동안 생성되는 것들과 같은, 음향, 광, 또는 전자기파들의 형태를 또한 가질 수 있다. 컴퓨터 판독 가능 매체들의 일반 형태들은, 예를 들어, 플로피 디스크, 플렉시블 디스크 (flexible disk), 하드 디스크 (hard disk), 자기 테잎, 임의의 기타 자기 매체, CD-ROM, CD-RW, DVD, 임의의 기타 광 매체, 펀치 카드들, 종이 테잎, 홀 마크 시트들 (optic mark sheets), 구멍들 또는 기타 광학적으로 인식 가능 표시들의 패턴들을 갖는 임의의 기타 물리 매체, RAM, PROM, 및 EPROM, FLASH-EPROM, 임의의 기타 메모리 칩 또는 카트리지, 캐리어 웨이브, 또는 컴퓨터가 판독할 수 있는 임의의 기타 매체를 포함한다.

다양한 형태들의 컴퓨터 판독 가능 매체들은 실행을 위해 프로세서로 명령들을 제공하는 것과 관련될 수 있다. 예를 들어, 본 발명의 적어도 일부분을 수행하는 명령들은 원격 컴퓨터의 자기 디스크에 저장되어 있을 수 있다. 그런 시나리오에서, 원격 컴퓨터는 명령들을 메인 메모리에 로드하고, 모뎀을 사용해서 명령들을 전화선을 통해 전송한다. 구내 컴퓨터 시스템의 모뎀은 전화선으로 데이터를 수신하고, 데이터를 적외선 신호로 변환하기 위해 적외선 전송기를 사용하고, PDA 및 랩탑과 같은, 모바일 컴퓨팅 디바이스로 적외선 신호를 전송한다. 모바일 컴퓨팅 디바이스의 적외선 감지기는 적외선 신호에 의해 전달되는 정보 및 명령들을 수신하고, 버스에 데이터를 싣는다. 버스는, 프로세서가 그 명령들을 수신하고 실행하는, 메인 메모리로 데이터를 전달한다. 메인 메모리에 의해 수신되는 명령들은 프로세서에 의한 실행 전 또는 후에 저장 장치에 선택적으로 저장될 수 있다.

발명의 효과

따라서, 본 발명의 다양한 실시예들은 구조화된 저밀도 패리티 검사 (Low Density Parity Check,LDPC) 코드들을 디코딩하는 새로운 방법을 제공한다. 병렬 디코딩 가능한 코드들의 조건들은 패리티 검사 매트릭스에 따른 맵핑된 매트릭스를 생성하기 위해 만들어진다. 이러한 조건들은, LDPC 코드들의 병렬 디코딩 구조를 개발하고 효율적 메모리 구조를 제공하는 LDPC 코드들로 귀결한다. 그 접근법은 n 보다 적은 수의 프로세서들을 위한 (즉, m이 n의 정수 프랙션인 m 프로세서들) 효율적 병렬 프로세싱을 위한 이러한 병렬 디코딩 가능 조건들의 보존을 또한 허용한다. 위의 배열 하에, 메모리 및 프로세싱 엔진들과 같은, 디코더 리소스들은 효과적으로 재용된다. 또한, 구현 유연성은 향상되고, m 병렬 엔진들이 사용될 수 있다.

본 발명이 다수의 실시예들과 연관해서 기술되는 가운데, 본 발명은 거기에 제한되지 않고, 첨부된 청구항들의 범위 내에 있는, 다양한 명백한 수정본들 및 동격 배치들을 포함한다.

(57) 청구의 범위

청구항 1.

저밀도 패리티 검사(LDPC) 코드의 디코딩을 지원하는 방법으로서,

상기 LDPC 코드와 관련된 패리티 검사 매트릭스에 기초하여 맵핑된 매트릭스를 구성하는 단계 - 상기 맵핑된 매트릭스는 병렬로 동작하는 복수의 프로세서(903, 1103)에 의해 액세스될 수 있는 루럴리 메모리 구조(lumped memory structure: 901, 1101)를 허용하는 복수의 병렬 디코딩 가능 조건을 만족함 -; 및

상기 프로세서(903, 1103)에 의해 상기 LDPC 코드를 디코딩하기 위하여 상기 맵핑된 매트릭스를 메모리(901, 1101)에 저장하는 단계

를 포함하는 지원 방법.

청구항 2.

제1항에 있어서, 상기 병렬 디코딩 가능 조건은,

상기 패리티 체크 매트릭스의 동일 행 내의 임의의 2개의 상이한 엔트리들 중 어느 것도 동일 비트 노드 또는 동일 검사 노드에 접속되지 않는 것으로 결정하는 단계;

비트 노드 그룹 BG_i가 i번째 행의 에지에 접속된 모든 비트 노드를 나타내면, i≠j에 대하여, BG_i 및 BG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 결정하는 단계; 및

검사 노드 그룹 CG_i가 i번째 행의 에지에 접속된 모든 검사 노드를 나타내면, i≠j에 대하여, CG_i 및 CG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 결정하는 단계를 포함하는 지원 방법.

청구항 3.

제1항에 있어서, 상기 패리티 검사 매트릭스는 n×n 밀도 균형 매트릭스의 엔트리를 갖는 블록 매트릭스에 동가인 순열인 지원 방법.

청구항 4.

제3항에 있어서, 상기 n×n 밀도 균형 매트릭스의 엔트리 중 하나는 m×m 밀도 균형 매트릭스이고, 상기 n×n 밀도 균형 매트릭스의 다른 엔트리는 블록 밀도 균형 서브매트릭스로 분해될 수 있는 지원 방법.

청구항 5.

제4항에 있어서, m은 n의 정수 프랙션(integer fraction)인 지원 방법.

청구항 6.

제1항에 있어서, 상기 LDPC 코드는 8-PSK(Phase Shift Keying), 16-QAM(Quadrature Amplitude Modulation), 16-APSK(Amplitude Phase Shift Keying), 32-APSK 및 QPSK(Quadrature Phase Shift Keying) 중의 하나를 포함하는 신호 컨스텔레이션(constellation)에 따라 변조되는 신호에 의해 표현되는 지원 방법.

청구항 7.

제6항에 있어서, 상기 신호는 디지털 비디오 방송에 따른 위성 링크를 통해 송신되는 지원 방법.

청구항 8.

저밀도 패리티 검사(LDPC) 코드의 디코딩을 지원하는 명령을 갖는 컴퓨터 판독가능 매체로서, 상기 명령은, 실행시, 하나 이상의 프로세서(903, 1103)가 제1항에 기재된 지원 방법을 수행하도록 하는 컴퓨터 판독가능 매체.

청구항 9.

저밀도 패리티 검사(LDPC) 코드를 디코딩하는 디코딩 장치로서,

루터의 메모리 구조(901, 1101)를 허용하는 복수의 병렬 디코딩 가능 조건을 만족하는 맵핑된 매트릭스를 저장하도록 구성된 메모리(901, 1101); 및

병렬로 동작하는 복수의 프로세서(903, 1103) - 상기 프로세서(903, 1103)는 메모리(901, 1101)를 액세스하여 상기 LDPC 코드를 디코딩함 -

를 포함하는 디코딩 장치.

청구항 10.

제9항에 있어서, 상기 병렬 디코딩 가능 조건은,

상기 맵핑된 매트릭스의 동일 행 내의 임의의 2개의 상이한 엔트리들 중 어느 것도 동일 비트 노드 또는 동일 검사 노드에 접속되지 않는 것으로 결정하는 단계;

비트 노드 그룹 BG_i가 i번째 행의 에지에 접속된 모든 비트 노드를 나타내면, i≠j에 대하여, BG_i 및 BG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계; 및

검사 노드 그룹 CG_i가 i번째 행의 에지에 접속된 모든 검사 노드를 나타내면, i≠j에 대하여, CG_i 및 CG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계를 포함하는 디코딩 장치.

청구항 11.

제9항에 있어서, 상기 프로세서(903, 1103)의 수는 n이고, 상기 메모리(901, 1101)는 상기 맵핑된 매트릭스 내의 에지 값을 삭제함으로써 상기 맵핑된 매트릭스에 기초하여 구성된 도출된 맵핑된 매트릭스를 저장하고, 새로운 맵핑된 매트릭스가 m개의 병렬 프로세서(903, 1103)에 의해 사용되고, m은 n보다 작은 디코딩 장치.

청구항 12.

제11항에 있어서, m은 n의 정수 프랙션인 디코딩 장치.

청구항 13.

제11항에 있어서, 상기 패리티 검사 매트릭스는 n×n 밀도 균형 매트릭스의 엔트리를 갖는 블록 매트릭스에 동가인 표현인 디코딩 장치.

청구항 14.

제13항에 있어서, 상기 n×n 밀도 균형 매트릭스의 엔트리 중 하나는 m×m 밀도 균형 매트릭스이고, 상기 n×n 밀도 균형 매트릭스의 다른 엔트리는 블록 밀도 균형 서브매트릭스로 분해될 수 있는 디코딩 장치.

청구항 15.

제9항에 있어서, 상기 LDPC 코드는 8-PSK(Phase Shift Keying), 16-QAM(Quadrature Amplitude Modulation), 16-APSK(Amplitude Phase Shift Keying), 32-APSK 및 QPSK(Quadrature Phase Shift Keying) 중의 하나를 포함하는 신호 컨스텔레이션(constellation)에 따라 변조되는 신호에 의해 표현되는 디코딩 장치.

청구항 16.

제15항에 있어서, 상기 신호는 디지털 비디오 방송에 따른 위성 링크를 통해 송신되는 디코딩 장치.

청구항 17.

저밀도 패리티 검사(LDPC) 코드의 디코딩을 지원하는 방법으로서,

상기 LDPC 코드에 대응하는 패리티 검사 매트릭스에 기초하여 맵핑된 매트릭스를 저장하는 단계;

상기 맵핑된 매트릭스의 동일 행 내의 임의의 2개의 상이한 엔트리 중 어느 것도 동일 비트 노드 또는 동일 검사 노드에 집속되지 않는 것으로 결정하는 단계;

비트 노드 그룹 BG_i가 i번째 행의 에지에 접속된 모든 비트 노드를 나타내면, i≠j에 대하여, BG_i 및 BG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계;

검사 노드 그룹 CG_i가 i번째 행의 에지에 접속된 모든 검사 노드를 나타내면, i≠j에 대하여, CG_i 및 CG_j는 어떠한 공통 노드도 갖지 않거나 동일하다는 것을 검증하는 단계; 및

상기 맵핑된 매트릭스를 출력하여 병렬로 동작하는 복수의 프로세서(903, 1103)에 의해 액세스가능한 메모리(901, 1101)에 저장하고 상기 저장된 맵핑된 매트릭스에 따라 상기 LDPC 코드를 디코딩하는 단계

를 포함하는 지원 방법.

청구항 18.

제17항에 있어서, 상기 프로세서(903, 1103)의 수는 n이고,

상기 지원 방법은,

상기 맵핑된 매트릭스 내의 에지 값을 삭제하여 도출된 맵핑된 매트릭스를 출력하는 단계를 더 포함하고,

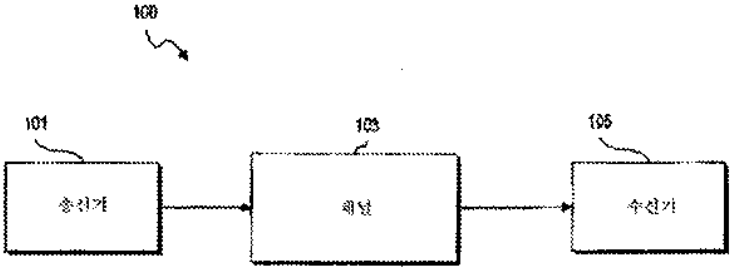
상기 도출된 맵핑된 매트릭스는 m개의 병렬 프로세서(903, 1103)에 의해 사용되고, m은 n보다 작은 지원 방법.

청구항 19.

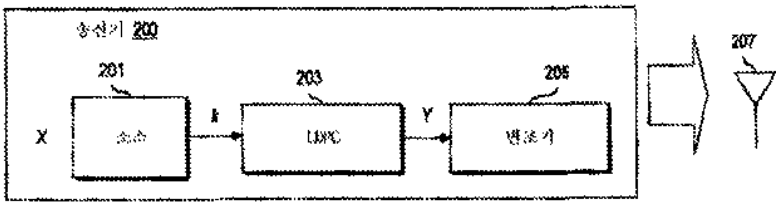
저밀도 패리티 검사(LDPC) 코드의 디코딩을 지원하는 명령을 갖는 컴퓨터 판독가능 매체에 있어서, 상기 명령은, 실행시, 하나 이상의 프로세서(903, 1103)가 제17항에 기재된 지원 방법을 수행하도록 하는 컴퓨터 판독가능 매체.

도면

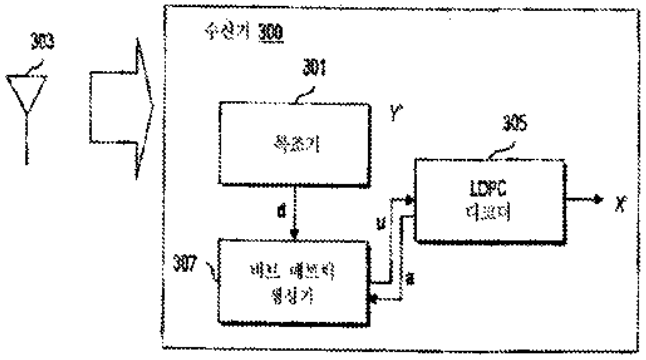
도면1



도면2



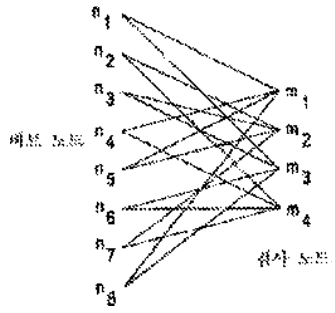
도면3



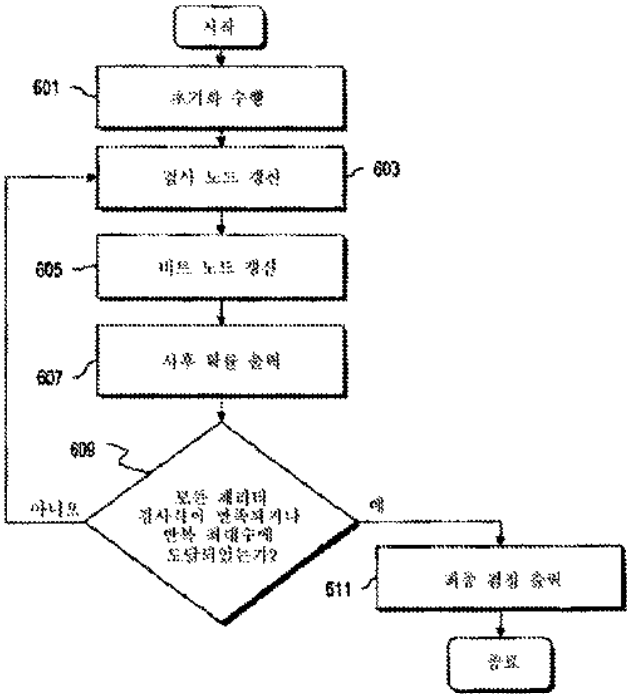
도면4

$$H = \begin{array}{c} \begin{matrix} n_1 & n_2 & n_3 & n_4 & n_5 & n_6 & n_7 & n_8 \end{matrix} \\ \begin{bmatrix} 1 & 0 & 0 & 1 & 1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 & 1 & 0 & 1 & 0 \\ 1 & 0 & 1 & 0 & 0 & 1 & 0 & 1 \\ 0 & 1 & 0 & 1 & 0 & 1 & 1 & 0 \end{bmatrix} \begin{matrix} m_1 \\ m_2 \\ m_3 \\ m_4 \end{matrix} \end{array}$$

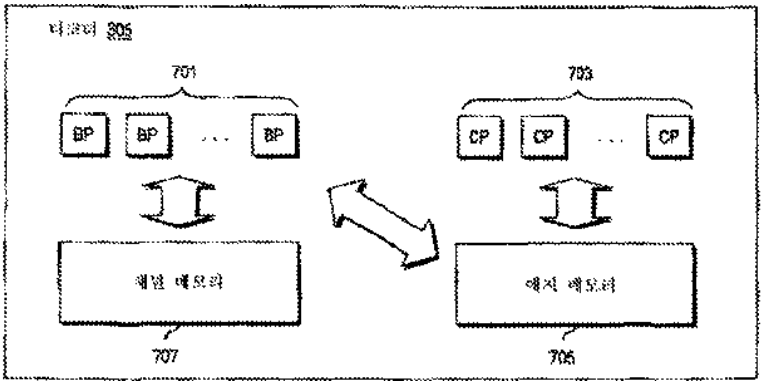
도면5


n_1
n_2
n_3
비트 노드 n_4
n_5
n_6
n_7
n_8
m_1
m_2
m_3
m_4
검사 노드

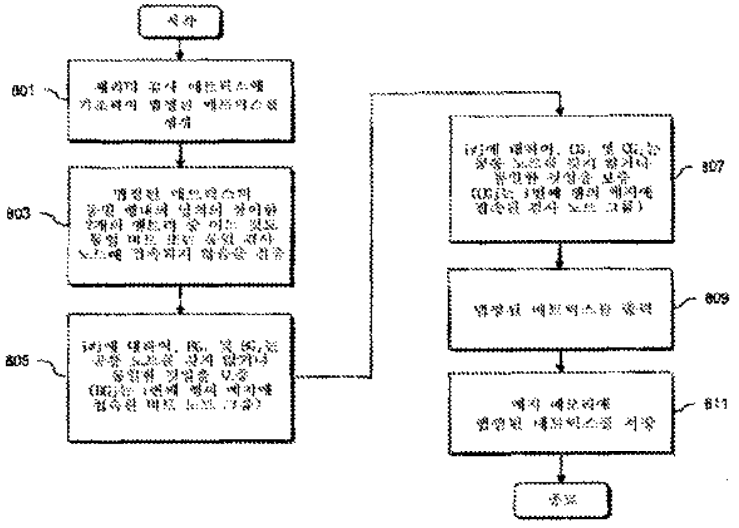
도면6


시작
601
초기화 수행
검사 노드 갱신
603
비트 노드 갱신
605
사후 확률 출력
607
609
모든 패리티
검사식이 만족되거나
반복 최대수에
도달되었는가?
아니오
예
611
최종 결정 출력
종료

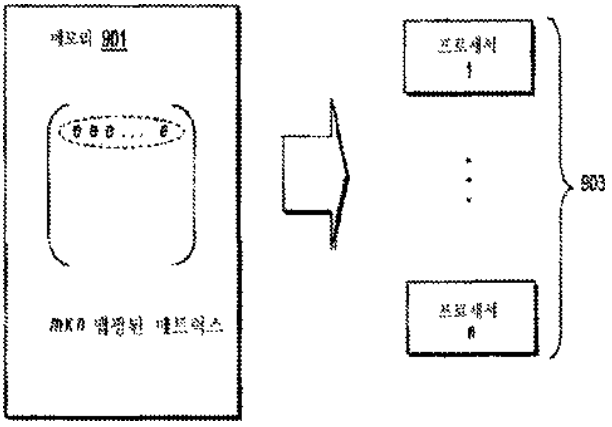
도면7



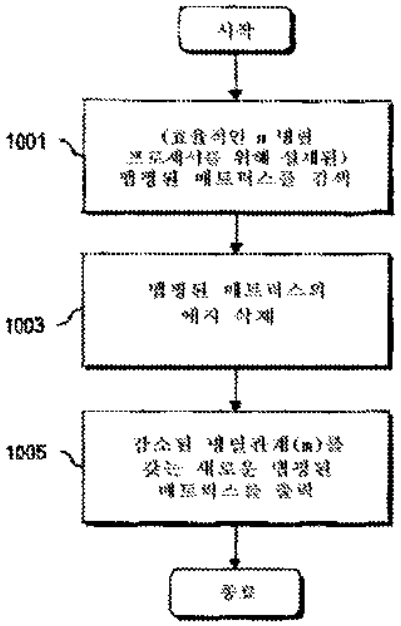
도면8



도면9


메모리 901
m×n 맵핑된 매트릭스
프로세서 1
프로세서 n
903

도면10


시작
1001
(효율적인 n 병렬 프로세서를 위해 설계된) 맵핑된 매트릭스를 검색
1003
맵핑된 매트릭스의 에지 삭제
1005
감소된 병렬관계(m)를 갖는 새로운 맵핑된 매트릭스를 출력
종료

1225
1221
LAN
1223
1219
1201
1209
1217
1230
1207
ROM
1203
1205
1211
1213
1215

도면12

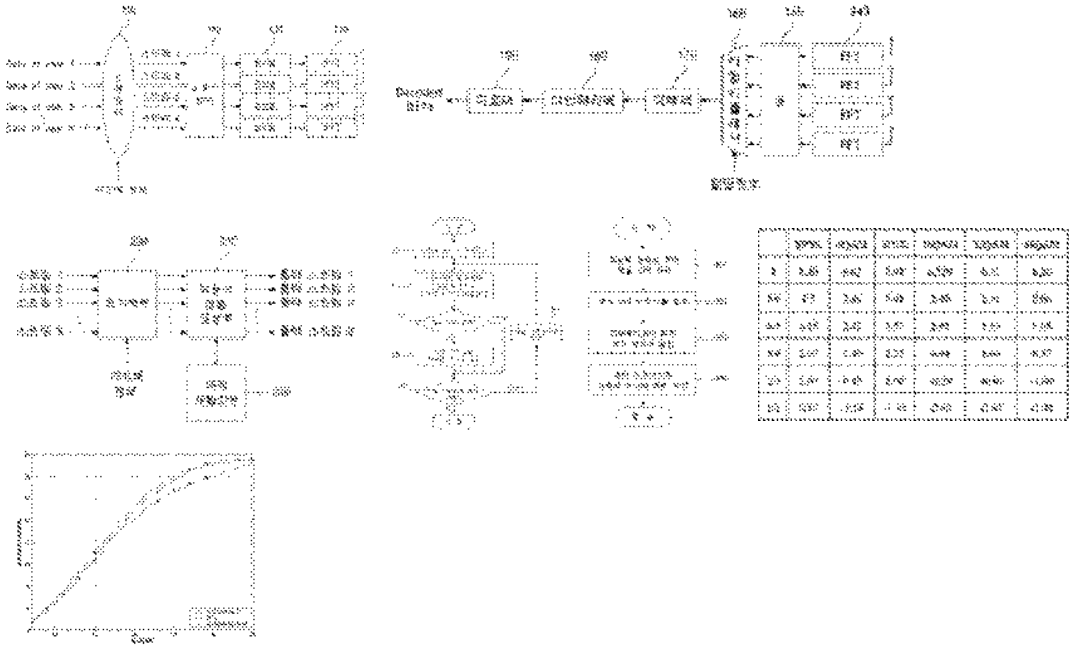
도면13

KR100723018B1

Method for adaptive modulation using residual power and apparatus thereof

Abstract

An adaptive modulation method and apparatus thereof using residual power are disclosed.
According to an embodiment of the present invention, a plurality of streams are arranged in order of eigenvalues, and a predetermined power is assigned to each of the streams. Determining an energy level of a symbol to be transmitted at, if there is residual power in the priority stream, further assigning the residual power to a next stream having an eigenvalue less than the priority stream, and the code rate of the next stream and the Determining an energy level of a symbol to be transmitted in the next stream using feedback information.
According to the present invention, the energy level of a symbol is determined in consideration of different code rates for each stream, and the remaining power is additionally allocated to a stream having a low eigenvalue, thereby preventing waste of residual power and having a constraint of a target bit error rate. Even if the maximum size of the constellation is limited, the performance of the system can be improved, and no additional computation is required for the implementation.



Classifications

H04B7/0434 Power distribution using multiple eigenmodes
View 3 more classifications
South Korea
Download PDF Find Prior Art Similar
Other languages
Korean
Inventor
성창경
이인규
조정호

Worldwide applications
2006 KR

Application KR1020060027259A events

2005-12-29

Priority to KR20050134343

2005-12-29

Priority to KR1020050134343

2006-03-27

Application filed by 고려대학교 산학협력단

2007-05-30

Application granted

2007-05-30

Publication of KR100723018B1

Info

Patent citations (4)

Cited by (1)

Legal events

Similar documents

Priority and Related Applications

External links

Espacenet
Global Dossier
Discuss

Description

{Method for adaptive modulation using residual power and Apparatus}

1A and 1B illustrate a MIMO BIC-OFDM system according to an embodiment of the present invention.

2 is a block diagram of the present invention.

3 is a flowchart of the present invention.

4 is a detailed flowchart of an energy level determination process of FIG. 3.

FIG. 5 illustrates an example of a coding offset table used in the energy level determination process of FIG. 3.

6 is a graph showing a conventional modulation method and the performance according to the present invention.

The present invention relates to orthogonal frequency multiple division modulation, and more particularly, to an adaptive modulation method and apparatus using residual power.

Recently, in the field of wireless communication, a system that supports high-speed packet data transmission is pursued. As a technique for supporting a high data rate, a multiple input multiple output (MIMO) system is attracting attention, and the application of an orthogonal frequency multiple division (OFDM) modulation scheme can effectively overcome the frequency selective nature of the channel. Here

The combination of Bit Interleaved Coded Modulation (BICM) to achieve good diversity gain in the frequency domain has been widely applied to wireless standard technology in recent years. Examples are technologies such as IEEE 802.11a wireless LAN (WLAN) and IEEE 802.16e wireless metropolitan area network (WMAN).

When channel state information (CSI) is known to the transmitter, the transmission rate may be maximized through an adaptive modulation and coding (AMC) scheme. Recently, a bit allocation scheme using the Levin-Campello (LC) algorithm has been proposed in the SISO BIC-OFDM system, which uses an approximate coding gain (asymptotic coding gain) to apply the LC algorithm. The constraints of the bit error rate (BER) are relaxed.

Also, if the transmitter knows the channel state information, the multiple input multiple output (MIMO) channel corresponding to each subcarrier frequency

can be divided into independent parallel channels by applying a single value decomposition (SVD) technique. Can be. System performance can be maximized by using an optimized power allocation scheme through water-filling (WF) in the spatial domain and a bit allocation scheme in each subchannel. However, according to the conventional modulation method as described above, in the high SNR range, there is residual power even after allocating the maximum possible number of bits.

Therefore, in the conventional modulation method, when the maximum size of the constellation is limited, there is a problem of limiting performance improvement and wasting residual power due to the inherent property of the single-value decomposition technique.

Accordingly, the first technical problem to be achieved by the present invention is to provide an adaptive modulation method using residual power that can prevent the waste of residual power and improve performance even when the maximum size of the constellation is limited.

A second technical problem to be achieved by the present invention is to provide an adaptive modulation device using the residual power to which the adaptive modulation method using the residual power is applied.

In order to solve the first technical problem described above, the present invention comprises the steps of arranging a plurality of streams in order of eigenvalues, allocating a predetermined power for each of the streams, Determining an energy level of a symbol to be transmitted in the priority stream by using feedback information of a receiver; if residual power is present in the priority stream, additionally assigning the residual power to a next stream having a lower eigenvalue than the priority stream And determining an energy level of a symbol to be transmitted in the next stream using the code rate of the next stream and the feedback information.

In order to solve the above second technical problem, the present invention provides an initialization unit for arranging a plurality of streams in order of eigenvalues and allocating a predetermined power for each of the streams. Adaptive symbol generation unit for determining the energy level of a symbol to be transmitted in the streams using the code rate of the streams and the feedback information of the receiving end, after determining the energy level of the symbol in any of the streams to the arbitrary stream If there is residual power, there is provided an adaptive modulation apparatus that uses the

residual power including a power reassignment for further allocating the residual power to the next stream having an eigenvalue less than the arbitrary stream.

The present invention is characterized by an adaptive modulation scheme that provides improved performance in a MIMO BIC-OFDM system. In particular, unlike the conventional scheme using a coding gain, the bit allocation scheme is considered in consideration of the code rate used for each stream. It is further optimized and increases the power efficiency by delivering the remaining power allocated to each stream to the lower stream.

Hereinafter, with reference to the drawings will be described a preferred embodiment of the present invention.

1A and 1B illustrate a MIMO BIC-OFDM system according to an embodiment of the present invention. 1A and 1B assume an adaptive modulated MIMO BIC-OFDM system with the number of transmit antennas N_t = 4 and the number of receive antennas N_r = 4.

1A is a block diagram illustrating a transmitting end of an adaptive modulated MIMO BIC-OFDM system.

Referring to FIG. 1A, the user information (Data of user 1, 2, 3, ..., K) transmitted from the base station forms a total of N_t pieces, that is, four streams. Here, the i-th stream means user information transmitted through the i-th antenna.

The scheduler 100 outputs a plurality of user information in a limited stream (1, 2, 3, 4).

Then, the feedback information is used to determine the current channel status and adjust the data rate and power level based on the current channel status. Once the rate and power are determined, the data modulated according to the BICM scheme for each stream is transmitted through an orthogonal frequency multiplexing (OFDM) modulation technique having N_c subcarriers.

Feedback channels that provide channel information usually use good performance channel coding and low modulation levels to ensure reliable transmission. Therefore, the present invention assumes an ideal feedback channel without error.

The AMC block 110 allocates and outputs bits to be transmitted to each stream.

The plurality of BICMs 120 expands the size of data according to the code rate of each stream.

The plurality of IFFTs 130 performs an inverse fast Fourier transform on each stream.

Here, by using the BICM technique, it is possible to form a diversity gain in the frequency domain through channel coding in a frequency selective channel. OFDM is implemented by combining a binary convolutional coder and a symbol mapper through a bitwise interleaver. BIC-OFDM is implemented through the inverse fast Fourier transform (IFFT) and fast Fourier transform (FFT) modulation schemes in the BICM technique.

1B is a block diagram illustrating a receiving end of an adaptive modulated MIMO BIC-OFDM system.

The plurality of FFTs 140 performs fast Fourier transform on the signal received by each receive antenna. The weight applying unit 150 applies a weight to each stream, the demultiplexer 160 combines each stream according to the allocation information, and the demapper 170, the deinterleaver 180, and the decoder 190 are original bits. Restore it.

2 is a block diagram of the present invention when the number of streams is N. FIG.

The initialization unit 200 arranges the plurality of streams in order of eigenvalues, and allocates a predetermined power to each stream. In addition, the initialization unit 200 may calculate the eigenvalue for each stream from the feedback information of the receiver. In addition, the initialization unit 200 may detect a state of a channel through which the symbol generated by the adaptive symbol generator 210 is transmitted from feedback information of the receiver, and determine a predetermined power according to the state of the detected channel.

The adaptive symbol generator 210 determines the energy level of a symbol to be transmitted in the streams using the code rate of the streams and the feedback information of the receiver in the order of high eigenvalues for each

stream. In this case, the symbol is information of a bit unit consisting of a plurality of bits included in the user information.

In addition, the adaptive symbol generator 210 may compensate an energy level of a symbol to be transmitted in the streams by using a coding offset according to the code rate of the streams. Preferably, the coding offset may be determined with reference to a coding offset table which previously stored an offset corresponding to the code rate of the streams based on a predetermined bit error rate. At this time, the output streams 1 to N are streams which have completed the energy level determination process and the additional allocation process of residual power.

The power reallocator 220 further allocates the remaining power to the next stream having a lower eigenvalue than the above stream if there is residual power in the above stream after determining the energy level of the symbol in any of the streams.

Preferably, the adaptive modulation device of FIG. 2 may be implemented within the AMC block 110 of FIG. 1A.

3 is a flowchart of the present invention.

First, a plurality of streams are arranged in order of eigenvalues, and a predetermined power is allocated to each of the streams (step 300). This process 300 may include detecting a state of a channel through which a symbol is transmitted from feedback information of a receiver and determining a predetermined power according to the state of the channel. In addition, this process (process 300) may include a process of calculating the eigenvalue for each stream from the feedback information of the receiving end.

Next, the streams are selected in order of increasing eigenvalues among the streams, and the energy level of the symbol to be transmitted in the selected stream is determined using the code rate of the selected stream and the feedback information of the receiver (step 310). This process 310 may include generating a symbol to be transmitted from the user information. At this time, the step 310 of determining the energy level of the symbol consumes at least a portion of power initially allocated to each stream.

When the process of determining the energy level of the symbol in the selected stream is completed, it is determined whether residual power exists in the selected stream (step 320).

If there is residual power in the selected stream, the remaining power is additionally allocated to the next stream having a smaller eigenvalue than the selected stream (step 330). On the other hand, if there is no residual power in the selected stream, the process proceeds to the next step (340).

Next, it is determined whether the above processes (steps 310-330) are completed for all streams (step 340). At this time, if the above process (steps 310-330) is not completed for all streams, the process proceeds to the next stream (350) and repeats the above steps (310-330) (step 340).

On the other hand, if the above steps (steps 310-330) have been completed for all streams, all steps are terminated (step 340).

Here, a stream having the largest eigenvalue may be defined as a priority stream, a stream having a unique value smaller than the priority stream, a next stream, and a stream having a unique value less than the next stream as a subordinate stream.

4 is a detailed flowchart of the energy level determination process 310 of FIG. 3.

First, the state of the channel through which the symbol for each stream is transmitted is determined from the feedback information (step 411).

Next, a specific position of the coding offset table is referred to according to the code rate and the channel state for each stream (step 412).

Next, the coding offset according to the code rate for each stream is determined (step 413).

Finally, the energy level of the symbol transmitted in the stream is compensated with the determined coding offset (step 414).

FIG. 5 illustrates an example of a coding offset table used in the energy level determination process 310 of FIG. 3.

FIG. 5 is a table illustrating coding offsets corresponding to various channel codes and modulation schemes used in a MIMO BIC-OFDM system according to an embodiment of the present invention. Here, the leftmost column represents the code rate of the stream, and the topmost column represents various modulation schemes. The coding offset described here is an experimental value obtained based on the bit error rate (BER) 10 ^ 4 and is expressed in decibels (dB).

The present invention is described as follows.

The received signal in the nth subcarrier is represented by Equation 1 below.

$$\mathbf{r}_n = \begin{bmatrix} r_1^n \\ \vdots \\ r_{N_r}^n \end{bmatrix} = \mathbf{H}_n \mathbf{x}_n + \mathbf{z}_n$$

$$= \begin{bmatrix} H_{11}^n & \cdots & H_{N_t 1}^n \\ \vdots & \ddots & \vdots \\ H_{1N_r}^n & \cdots & H_{N_t N_r}^n \end{bmatrix} \begin{bmatrix} x_1^n \\ \vdots \\ x_{N_t}^n \end{bmatrix} + \begin{bmatrix} z_1^n \\ \vdots \\ z_{N_r}^n \end{bmatrix}$$

In Equation 1, H_ (ij) ^ n indicates a channel frequency response between the i th transmit antenna and the j th receive antenna in the n th subcarrier. z_j ^ n is a complex Gaussian noise of independent coincidence distribution following the variance of σ_z ^ 2, and x_i ^ n denotes a symbol transmitted to the i th transmit antenna having a variance of σ_x ^ 2. It is assumed that the total power used in the transmission stage is P = N_t * σ_x ^ 2.

The available channel codes and the magnitude of the maximum constellation are assumed to be V and 2 ^ (m_max), respectively. In each transmission, the i-th stream undergoes encoding by selecting one of V different channel codes, and the n-th subcarrier of the i-th stream is b_n ^ i ∈ {m_1, m_2,... , m_max} (n = 1, 2, ..., N_c) bits are transmitted. [b_1 ^ i, b_2 ^ i,... , B_ (N_c) ^ i)] If the bit vector **b** _i la of the i-th stream, i spectral efficiency (spectral efficiency of the second stream) is expressed as the following equation (2).

$$R_T(\mathbf{b}; C_{v^*}) = R_s(C_{v^*}) \frac{1}{N_c} \sum_{n=1}^{N_c} b_n$$

In Equation 2, R_c (C_v *) denotes a coding rate selected by the channel encoder C_v *. The channel encoder C_v * and the number of bits b_n ^ i allocated to the nth subcarrier are determined by the AMC scheme.

In more detail, the n-th subchannel H_n can be expressed by Equation 3 through the SVD method.

$$\mathbf{H}_n = \mathbf{V}_n \mathbf{\Sigma}_n \mathbf{U}_n^*$$

In Equation 3, V_n and U_n are unitary matrices, and Σ_n is a diagonal matrix having eigenvalues of H_n as components. Knowing the SVD information, the MIMO channel H_n can be diagonalized through filtering using U_n and V_n ^ *. The equivalent model of the signal received at the receiver can be expressed as Equation 4 below.

$$\begin{aligned} \mathbf{y}_n &= \mathbf{V}_n^*(\mathbf{H}_n \mathbf{U}_n \mathbf{x}_n + \mathbf{z}_n) \\ &= \mathbf{\Sigma}_n \mathbf{x}_n + \mathbf{V}_n^* \mathbf{z}_n \end{aligned}$$

As in Equation 4 above, **x** _ n indicates a symbol transmitted in the nth subcarrier. **The** most common method of controlling **x** _ n is a power allocation scheme through eigenmodes. The optimized power allocation technique is obtained through the well-known water-filling technique in the spatial domain. However, this requires a lot of complexity.

On the other hand, the bit allocation scheme is a method of allocating more bits to a mode having a larger eigenvalue. The bit rate is maximized in the MIMO BIC-OFDM system according to another embodiment of the present invention. This is optimized in terms of channel capacity through spatio-temporal vector coding (STVC). However, this method is very complicated because SVD must be performed in N_rN_c × N_tN_c dimension. In addition, it is known that the algorithm performing N_c times the SVD of N_r × N_t dimension almost follows the optimized performance as N_c increases. Thus, the present invention takes the latter method to reduce

complexity. This type of rate maximization method can be formulated as in Equation 5 below.

$$\text{maximize} : \quad \sum_{i=1}^{N_t} R_i(C_{ij}) \sum_{n=1}^{N_c} b_{n,i}(\lambda_{n,i})$$

$$\text{subject to} : \quad \sum_{i=1}^{N_t} \sum_{n=1}^{N_c} \mathcal{E}_{n,i} = N_t N_c \sigma_x$$

$$P_b \le P_t$$

As can be seen from Equation 5, the conventional Levin-Campello (LC) algorithm is optimized bit allocation vector **b** _i through equal power distribution of each subcarrier when the maximum constellation size is not limited. And the code index v_i ^ *. However, in practical situations where the maximum supported feature size is limited, the above-mentioned residual power problem arises, resulting in poor performance. SVDs performed on each subcarrier are sorted in descending order of eigenvalues. At high SNR, this concentrates on streams of indices where the number of bits of m_max is low. In this case, the energy required to transmit the m_max bits is less than E_x,

$\sum_{n=1}^{N_c} \mathcal{E}_{n,i}$

Also

$N_c \mathcal{E}_x$

Has a smaller value. therefore

$N_c \mathcal{E}_x - \sum_{n=1}^{N_c} \mathcal{E}_{n,i}$

As much residual power is generated.

Conversely, high index streams with small eigenvalues have room to carry more information when allocating additional power. Therefore, delivering extra power from a low index stream to another stream can increase the overall throughput of the system. This may be arranged as an equation as shown in Equation 6 below.

$$\hat{\mathcal{E}}_{i+1} = \left(\hat{\mathcal{E}}_i - \frac{1}{N_c} \sum_{n=1}^{N_c} \mathcal{E}_{n,i} \right)_+ + \mathcal{E}_x$$

$$(x)_+ = \begin{cases} x & \text{if } x \geq 0 \\ 0 & \text{if } x < 0 \end{cases}$$

As shown in Equation 6 above, when applied to the LC Efficientizing (EF) algorithm and the E-Tightening (ET) algorithm, a bit rate is allocated to a stream having a small eigenvalue, thereby increasing the transmission rate.

Meanwhile, the discrete bit allocation algorithm (MIMO) (BLCM / RPD (E_x)) will be described below.

1) Obtain $\boldsymbol{\Sigma}_n = \mathbf{V}_n^* \mathbf{H}_n \mathbf{U}_n$ for $n = 1, \cdots, N_c$

2) $\mathcal{E}_1 \leftarrow \mathcal{E}_x$

3) FOR $1 \leq i \leq N_t$

a) $\hat{\mathbf{H}} \leftarrow [\lambda_{1,i}, \cdots, \lambda_{N_c,i}]$

b) obtain $\mathbf{b}_i$ by performing the DBLC/CS($\hat{\mathbf{H}}, \sigma_x$)) with the constraint of $\hat{\mathcal{E}}_i$

c) $\hat{\mathcal{E}}_{i+1} = \left(\hat{\mathcal{E}}_i - \frac{1}{N_c}\sum_{n=1}^{N_c} \mathcal{E}_{n,i}\right)_+ + \mathcal{E}_x$

In Equation 7, the value of c) can be directly obtained through the ET algorithm without any additional calculation amount.

In addition, the discrete bit allocation algorithm (DBLC / CS (H, σ_x)) in the encoding system will be described with Equation (8).

1) FOR $1 \leq v \leq V$

a) Obtain $\mathbf{b}_v$ with GLCA($\hat{\mathbf{H}}, v, \sigma_x$)

b) Compute $R_T(\mathbf{b}_v; \mathcal{E}_v)$

2) $v^* = \arg\max_{1 \leq v \leq V} R_T(\mathbf{b}_v; \mathcal{E}_v)$

3) $\mathbf{b} = \mathbf{b}_{v^*}$.

As shown in Equation 8, the generalized LC algorithm (GLC) of the present invention will be described. The conventional LC algorithm is one of the best known optimized bit allocation algorithms, and performs the algorithm using an approximate coding gain through a fixed channel gap (Γ). The energy function obtained through gap approximation in M-QAM is expressed by Equation 9 below.

$$\mathcal{E}_n(b_n) = (2^{b_n} - 1)\frac{\sigma_n^2}{|H_n|^2} \cdot \Gamma$$

As shown in Equation 9, in the uncoded system, the channel gap is assumed to be constant, but in the encoding system, a value corresponding to the gap is changed according to the characteristics of the applied channel code. The gap of the system to which the code rate R_c is applied is expressed by Equation 10 below.

$$\Gamma_{coded} = 10\log_{10}(2^{mR_c} - 1) - \Psi(m, R_c)\ (\mathrm{dB})$$

(M, R_c) is an SNR value required to satisfy the bit error rate (BER) 10^{-4} when the code rate and the constellation size are applied as R_c and 2^m, respectively. When channel coding is applied to each stream, the number of bits carried in each subcarrier no longer means the number of information bits. In this case, the channel gap should be replaced with a coding offset defined as in Equation 11 below.

$$\gamma_v(m) = \Psi(m, R_c(C_v)) - 10\log_{10}(2^m - 1) \quad (dB)$$

The energy function considering this coding offset as a new parameter is given by Equation 12 below.

$$\mathcal{E}_n(b_n; v) = 2^{b_n - 1}\frac{\sigma_n^2}{|H_n|^2} \cdot \gamma_v(b_n)$$

6 is a graph showing a conventional modulation method and the performance according to the present invention.

In this case, we consider an OFDM system consisting of N_c = 64 samples and CP = 16 samples, and used a channel with exponentially decaying power consisting of five taps. A 64-state RCPC code was used and the target frame error rate (FER) was set to 1%. The corresponding bit error rate is 10^{-4}.

6 is a graph showing the performance when a single user of N_t = N_r = 4 is used. The conventional SVD, distinguished by the rightmost asterisk, represents the throughput of a conventional water-filling bit allocation system. In other words, the system of the conventional SVD uses a fixed channel gap in the conventional LC algorithm and does not consider the remaining power distribution.

Performance Curve of System for Reallocating Power According to the Invention (RPD with GLCRPD) According to the present invention, the performance curve (RPD) of the system which compensates for the energy of the symbol in consideration of the code rate is higher than the conventional SVD of the conventional system. It can be seen that the throughput is high, that is, the performance is improved by about 4 dB based on the spectral efficiency of 18 bps / Hz.

Since most practical systems all have a limit on the maximum feature size that can be supported, the problem of residual power in the high SNR range is very common. The present invention can obtain a large performance gain especially in the high SNR range by proposing a bit allocation scheme that compensates for this.

Preferably, a program for executing the adaptive modulation method using the residual power of the present invention on a computer can be recorded on a computer-readable recording medium.

The invention can be implemented via software. When implemented in software, the constituent means of the present invention are code segments that perform the necessary work. The program or code segments may be stored on a processor readable medium or transmitted by a computer data signal coupled with a carrier on a transmission medium or network.

Although the present invention has been described with reference to one embodiment shown in the drawings, this is merely exemplary and will be understood by those of ordinary skill in the art that various modifications and variations can be made therefrom. However, such modifications should be considered to be within the technical protection scope of the present invention. Therefore, the true technical protection scope of the present invention will be defined by the technical spirit of the appended claims.

As described above, according to the present invention, the energy level of the symbol is determined in consideration of different code rates for each stream, and the remaining power is additionally allocated to the stream having a low eigenvalue, thereby preventing waste of residual power and reducing the target bit error rate. Even when there are constraints or when the maximum size of constellations is limited, the performance of the system can be improved and the implementation does not require additional computation.

Claims (15)

Hide Dependent

1. Arranging a plurality of streams in order of eigenvalues, and allocating a predetermined power for each of the streams;

   Determining an energy level of a symbol to be transmitted in the priority stream by using a code rate of a priority stream having a high eigen value among the streams and feedback information of a receiver;

   If there is residual power in the seniority stream, further assigning the residual power to the next stream having an eigenvalue less than the seniority stream; And

   Determining an energy level of a symbol to transmit in the next stream using the code rate of the next stream and the feedback information.

2. The method of claim 1,

   Allocating a predetermined power for each stream

Detecting a state of a channel through which the symbol is transmitted from feedback information of the receiving end; And

And determining the predetermined power in accordance with the state of the channel.

3. The method of claim 1,

Allocating a predetermined power for each stream

And calculating a eigenvalue for each stream from the feedback information of the receiving end.

4. The method of claim 1,

Determining an energy level of a symbol to be transmitted in the priority stream

Compensating for an energy level of a symbol to be transmitted in the priority stream using a coding offset according to the code rate of the priority stream.

5. The method of claim 4, wherein

The coding offset is

And a coding offset table that stores an offset corresponding to a code rate of the priority stream based on a predetermined bit error rate.

6. The method of claim 1,

Determining the energy level of the symbol to be transmitted in the next stream

Compensating for an energy level of a symbol to be transmitted in the next stream using a coding offset according to the code rate of the next stream.

7. The method of claim 6,

The coding offset is

Adaptive modulation method using the residual power, characterized in that determined using a coding offset table that previously stored an offset corresponding to the code rate of the next stream on the basis of a predetermined bit error rate.

8. The method of claim 1,

If there is residual power in the next stream, further assigning the residual power to a subordinate stream having an eigenvalue less than the next stream; And

Determining an energy level of a symbol to transmit in the subordinate stream using the code rate of the subordinate stream and the feedback information;

And further allocating the residual power and determining the energy level of a symbol to be transmitted in the subordinate stream for all streams.

9. The method of claim 8,

Determining the energy level of the symbol to be transmitted in the subordinate stream

Compensating for an energy level of a symbol to be transmitted in the subordinate stream using a coding offset according to the code rate of the subordinate stream;

And the coding offset is determined by using a coding offset table which previously stores an offset corresponding to a code rate of the subordinate stream based on a predetermined bit error rate.

10. A computer-readable recording medium having recorded thereon a program for executing the method of claim 1 on a computer.

11. An initialization unit for arranging a plurality of streams in order of eigenvalues and allocating a predetermined power for each of the streams;

An adaptive symbol generator configured to determine an energy level of a symbol to be transmitted in the streams by using the code rate of the streams and feedback information of a receiver in the order of increasing eigenvalues for each of the streams;

If there is residual power in the arbitrary stream after determining the energy level of the symbol in any of the streams, a power reallocation section for further allocating the residual power to the next stream having an eigenvalue less than the arbitrary stream Adaptive modulation device using the residual power comprising.

12. The method of claim 11, wherein

The initialization unit

And detecting the state of a channel through which the symbol is transmitted from the feedback information of the receiving end, and determining the predetermined power according to the state of the channel.

13. The method of claim 11, wherein

The initialization unit

Adaptive modulation apparatus using the residual power, characterized in that for calculating the eigen value for each stream from the feedback information of the receiving end.

14. The method of claim 11, wherein

The adaptive symbol generation unit

And an energy level of a symbol to be transmitted in the streams by using a coding offset according to the code rate of the streams.

15. The method of claim 14,

The coding offset is

And an offset corresponding to a code rate of the streams based on a predetermined bit error rate.

Patent Citations (4)

Publication numberPriority datePublication dateAssigneeTitle
EP1037426A11998-10-272000-09-20Texas Instruments IncorporatedData allocation in multicarrier commumication systems
KR20020050299A *1999-11-222002-06-26추후제출Variable rate coding for forward link
KR20050027186A *2003-09-132005-03-18리전츠 어브 더 유니버시티 오브 미네소타Adaptive modulation for multi-antenna transmission with partial channel knowledge
KR20050053312A *2003-12-022005-06-08한국전자통신연구원A method for allocating and accessing radio resource in ofdma tele-communication system
Family To Family Citations
* Cited by examiner, † Cited by third party

Cited By (1)

Publication numberPriority datePublication dateAssigneeTitle
KR101015657B12007-12-032011-02-22아주대학교산학협력단Method for assigning power to multiuser for multiplexing using superposition coding in communication system
Family To Family Citations
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
US9900066B22018-02-20Quality control scheme for multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems
JP2016195464A2016-11-17Method and apparatus for implementing space frequency block coding in orthogonal frequency division multiplexing wireless communication system

JP5497218B22014-05-21User grouping for MIMO transmission in wireless communication systems
US20150318906A12015-11-05Method and apparatus for transmitting/receiving feedback information in a wireless packet data communication system
US9577730B22017-02-21Channel quality index feedback reduction for broadband systems
JP5376539B22013-12-25Method and apparatus for performing transmission in a multi-carrier communication system
JP5579647B22014-08-27Method and system for processing data for transmission in a multi-channel communication system using selective channel transmission
US20150030097A12015-01-29Method and apparatus for subcarrier and antenna selection in mimo-ofdm system
JP5850509B22016-02-03Method and apparatus for modulation of signal field in WLAN frame header
JP4927976B22012-05-09Beam-steering and beam-forming for wideband MIMO / MISO systems
JP5671078B22015-02-18MIMO transmitter and receiver for supporting single channel codeword downlink communication
Miyashita et al.2002High data-rate transmission with eigenbeam-space division multiplexing (E-SDM) in a MIMO channel
US8503555B22013-08-06Feedback of differentially encoded channel state information for multiple-input multiple-output (MIMO) and subband scheduling in a wireless communication system
Xia et al.2004Adaptive MIMO-OFDM based on partial channel state information
JP5265590B22013-08-14Multiple input, multiple output (MIMO) system with multiple transmission modes
US10264558B22019-04-16System and method for resource block-specific control signaling
KR101293373B12013-08-05Method for transmitting data in multiple antenna system
US8271043B22012-09-18Approach to a unified SU-MIMO/MU-MIMO operation
DK2267926T32017-04-10MIMO system with multiple spatial multiplexing modes
JP4359495B22009-11-04Adaptive modulation / demodulation apparatus and method applied to a multiple input / output system having a hierarchical space-time detector

TWI326165B2010-06-11Systems and methods for beamforming feedback in multi antenna communication systems
CA2600783C2012-07-17Systems and methods for beamforming and rate control in a multi-input multi-output communication systems
EP1897258B12018-05-23Robust channel rank prediction in MIMO communications systems
KR100983231B12010-09-20Coded mimo systems with selective channel inversion applied per eigenmode
US7095790B22006-08-22Transmission schemes for multi-antenna communication systems utilizing multi-carrier modulation

Priority And Related Applications

Priority Applications (2)

ApplicationPriority dateFiling dateTitle
KR200501343432005-12-29
KR10200501343432005-12-29

Legal Events

DateCodeTitleDescription
2006-03-27A201Request for examination
2007-05-17E701Decision to grant or registration of patent right
2007-05-22GRNTWritten decision to grant
2013-04-05FPAYAnnual fee payment
Payment date: 20130405

Year of fee payment: 7

2014-03-04FPAYAnnual fee payment
Payment date: 20140304

Year of fee payment: 8

2015-05-23LAPSLapse due to unpaid annual fee

(19)대한민국특허청(KR)
(12) 등록특허공보(B1)

(51) 。Int. Cl.
H04L 27/26 (2006.01)
H04B 7/04 (2006.01)
H04J 11/00 (2006.01)
H04L 12/28 (2006.01)

(45) 공고일자 2007년05월30일
(11) 등록번호 10-0723018
(24) 등록일자 2007년05월22일

(21) 출원번호 10-2006-0027259
(22) 출원일자 2006년03월27일
심사청구일자 2006년03월27일

(65) 공개번호
(43) 공개일자

(30) 우선권주장 1020050134343 2005년12월29일 대한민국(KR)

(73) 특허권자 고려대학교 산학협력단
서울 성북구 안암동5가1 고려대학교 내

(72) 발명자 이인규
서울 서초구 서초동 삼풍아파트 21동 1203호

성장경
서울 강남구 논현2동 동현아파트 5동 309호

조경호
경기 수원시 영통구 매탄2동 동남빌라 10동 101호

(74) 대리인 현종철

(56) 선행기술조사문헌
EP1037426 A1
KR1020050027186 A
KR1020020050209 A
KR1020050063312 A

심사관 : 박상범

전체 청구항 수 : 총 15 항

(54) 잔여 파워를 이용하는 적응 변조 방법 및 그 장치

(57) 요약

잔여 파워를 이용하는 적응 변조 방법 및 그 장치가 개시된다.

본 발명은 복수의 스트림을 고유치가 큰 순서대로 정렬하고, 상기 스트림마다 소정의 파워를 할당하는 단계, 상기 스트림을 중 고유치가 큰 선순위 스트림의 코드 레이트 및 수신단의 피드백 정보를 이용하여 상기 선순위 스트림에서 전송할

심볼의 에너지 레벨을 결정하는 단계, 상기 선순위 스트림에 잔여 파워가 존재하면, 상기 선순위 스트림보다 작은 고유치를 갖는 다음 스트림에 상기 잔여 파워를 추가로 할당하는 단계 및 상기 다음 스트림의 코드 레이트 및 상기 피드백 정보를 이용하여 상기 다음 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계를 포함한다.

본 발명에 의하면, 스트림마다 다른 코드 레이트를 고려하여 심볼의 에너지 레벨을 결정하고, 잔여 파워를 고유치가 낮은 스트림에 추가로 할당함으로써, 잔여 파워의 낭비를 막고, 목표 비트 에러율의 제약조건이 있는 경우나 성상의 최대크기가 제한된 경우에도 시스템의 성능을 향상시킬 수 있으며, 구현 과정에 추가적인 계산량을 요구하지 않는다.

대표도

도 2

특허청구의 범위

청구항 1.

복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 상기 스트림들마다 소정의 파워를 할당하는 단계;

상기 스트림들 중 고유치가 큰 선순위 스트림의 코드 레이트 및 수신단의 피드백 정보를 이용하여 상기 선순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계;

상기 선순위 스트림에 잔여 파워가 존재하면, 상기 선순위 스트림보다 작은 고유치를 갖는 다음 스트림에 상기 잔여 파워를 추가로 할당하는 단계; 및

상기 다음 스트림의 코드 레이트 및 상기 피드백 정보를 이용하여 상기 다음 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계를 포함하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 2.

제 1 항에 있어서,

상기 스트림마다 소정의 파워를 할당하는 단계는

상기 수신단의 피드백 정보로부터 상기 심볼이 전송되는 채널의 상태를 검출하는 단계; 및

상기 채널의 상태에 따라 상기 소정의 파워를 결정하는 단계를 포함하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 3.

제 1 항에 있어서,

상기 스트림마다 소정의 파워를 할당하는 단계는

상기 수신단의 피드백 정보로부터 스트림별 고유치를 연산하는 단계를 포함하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 4.

제 1 항에 있어서,

상기 선순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계는

상기 선순위 스트림의 코드 레이트에 따른 코딩 오프셋을 이용하여 상기 선순위 스트림에서 전송할 심볼의 에너지 레벨을 보상하는 단계를 포함하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 5.

제 4 항에 있어서,

상기 코딩 오프셋은

소정의 비트 에러율을 기준으로 상기 선순위 스트림의 코드 레이트에 대응하는 오프셋을 미리 저장한 코딩 오프셋 데이블을 이용하여 결정되는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 6.

제 1 항에 있어서,

상기 다음 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계는

상기 다음 스트림의 코드 레이트에 따른 코딩 오프셋을 이용하여 상기 다음 스트림에서 전송할 심볼의 에너지 레벨을 보상하는 단계를 포함하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 7.

제 6 항에 있어서,

상기 코딩 오프셋은

소정의 비트 에러율을 기준으로 상기 다음 스트림의 코드 레이트에 대응하는 오프셋을 미리 저장한 코딩 오프셋 데이블을 이용하여 결정되는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 8.

제 1 항에 있어서,

상기 다음 스트림에 잔여 파워가 존재하면, 상기 다음 스트림보다 작은 고유치를 갖는 후순위 스트림에 상기 잔여 파워를 추가로 할당하는 단계; 및

상기 후순위 스트림의 코드 레이트 및 상기 피드백 정보를 이용하여 상기 후순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계를 더 포함하고,

상기 잔여 파워를 추가로 할당하는 단계 및 상기 후순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계를 모든 스트림에 대해 반복하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 9.

제 8 항에 있어서,

상기 후순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계는

상기 후순위 스트림의 코드 레이트에 따른 코딩 오프셋을 이용하여 상기 후순위 스트림에서 전송할 심볼의 에너지 레벨을 보상하는 단계를 포함하고,

상기 코딩 오프셋은 소정의 비트 에러율을 기준으로 상기 후순위 스트림의 코드 레이트에 대응하는 오프셋을 미리 저장한 코딩 오프셋 테이블을 이용하여 결정되는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 방법.

청구항 10.

제 1 항 내지 제 9 항 중 어느 한 항의 방법을 컴퓨터에서 실행시키기 위한 프로그램을 기록한 컴퓨터로 읽을 수 있는 기록 매체.

청구항 11.

복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 상기 스트림들마다 소정의 파워를 할당하는 초기화부;

상기 스트림들마다 상기 고유치가 큰 순서대로 상기 스트림들의 코드 레이트 및 수신단의 피드백 정보를 이용하여 상기 스트림들에서 전송할 심볼의 에너지 레벨을 결정하는 적응적 심볼 생성부;

상기 스트림들 중 임의의 스트림에서 심볼의 에너지 레벨을 결정한 후에 상기 임의의 스트림에 잔여 파워가 존재하면, 상기 임의의 스트림보다 작은 고유치를 갖는 다음 스트림에 상기 잔여 파워를 추가로 할당하는 파워 재할당부를 포함하는 잔여 파워를 이용하는 적응 변조 장치.

청구항 12.

제 11 항에 있어서,

상기 초기화부는

상기 수신단의 피드백 정보로부터 상기 심볼이 전송되는 채널의 상태를 검출하고, 상기 채널의 상태에 따라 상기 소정의 파워를 결정하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 장치.

청구항 13.

제 11 항에 있어서,

상기 초기화부는

상기 수신단의 피드백 정보로부터 스트림별 고유치를 연산하는 것을 특징으로 하는 잔여 파워를 이용하는 적응 변조 장치.

청구항 14.

제 11 항에 있어서,

상기 적응적 심볼 생성부는

상기 스트림들의 코드 레이트에 따른 코딩 오프셋을 이용하여 상기 스트림들에서 전송할 심볼의 에너지 레벨을 보상하는 것을 특징으로 하는 관여 파워를 이용하는 적응 변조 장치.

청구항 15.

제 14 항에 있어서,

상기 코딩 오프셋은

소정의 비트 에러율을 기준으로 상기 스트림들의 코드 레이트에 대응하는 오프셋을 미리 저장한 코딩 오프셋 테이블을 이용하여 결정되는 것을 특징으로 하는 관여 파워를 이용하는 적응 변조 장치.

명세서

발명의 상세한 설명

발명의 목적

발명이 속하는 기술 및 그 분야의 종래기술

본 발명은 직교 주파수 다중 분할 변조에 관한 것으로 특히, 관여 파워를 이용하는 적응 변조 방법 및 그 장치에 관한 것이다.

근래의 무선통신 분야에서는 고속의 패킷 데이터 전송을 지원하는 시스템을 추구한다. 높은 데이터 전송률을 지원하기 위한 기법으로 다중입력 다중출력(MIMO) 시스템이 주목받고 있으며, 직교 주파수 다중 분할(OFDM) 변조 방식의 적용을 통해 채널의 주파수 선택적인 성질을 효과적으로 극복할 수 있다. 여기에

비트 인터리브드 코디드 모듈레이션(Bit Interleaved Coded Modulation: BICM)을 결합하여 주파수 도메인에서 좋은 다이버시티 이득을 얻는 방식이, 최근 무선 표준 기술에 광범위하게 적용되고 있다. IEEE 802.11a 무선 랜 (WLAN)과 IEEE 802.16e 무선 도시권 통신망 (WMAN)과 같은 기술이 그 예이다.

채널 상태 정보(CSI: channel state information)를 송신단에서 알고 있는 경우 적응 변조 (AMC: adaptive modulation and coding) 방식을 통해 전송률을 극대화시킬 수 있다. 근래에, SISO BIC-OFDM 시스템에서 레빈 캠펠로(Levin-Campello : LC) 알고리즘을 이용한 비트 할당 기법이 제안되었는데, 이 기법은 LC 알고리즘을 적용하기 위해 근사적인 코딩 이득(asymptotic coding gain)을 이용하므로, 비트 에러율 (BER: bit error rate)의 제약조건이 느슨해진다.

또한 송신단에서 채널 상태 정보를 완전히 알고 있는 경우, 각 부반송파 주파수에 해당되는 다중입력 다중출력(MIMO) 채널은 단일값 분해(singular value decomposition: SVD) 기법을 적용함으로써 독립적인 평행한 채널들로 분할될 수 있다. 공간 도메인에서의 워터 필링(water-filling: WF)을 통한 최적화된 파워 할당 기법과 각 부채널에서의 비트 할당 기법을 사용하면 시스템 성능이 극대화될 수 있다. 그러나, 위와 같은 종래의 변조 방법에 따르면 높은 SNR 범위에서, 최대로 가능한 숫자만큼의 비트를 할당하고도 관여 파워가 존재하게 된다.

따라서, 종래의 변조 방법은 성상(constellation)의 최대 크기가 제한된 경우 단일값 분해 기법의 고유한 성질로 인하여 성능 향상을 제한하고, 관여 파워를 낭비하는 문제점이 있다.

발명이 이루고자 하는 기술적 과제

따라서, 본 발명이 이루고자 하는 첫번째 기술적 과제는 관여 파워의 낭비를 막고, 성상의 최대크기가 제한된 경우에도 성능을 향상시킬 수 있는 관여 파워를 이용하는 적응 변조 방법을 제공하는데 있다.

본 발명이 이루고자 하는 두번째 기술적 과제는 상기의 잔여 파워를 이용하는 적응 변조 방법이 적용된 잔여 파워를 이용하는 적응 변조 장치를 제공하는데 있다.

발명의 구성

상기의 첫번째 기술적 과제를 해결하기 위하여, 본 발명은 복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 상기 스트림들마다 소정의 파워를 할당하는 단계, 상기 스트림들 중 고유치가 큰 선순위 스트림의 코드 레이트 및 수신단의 피드백 정보를 이용하여 상기 선순위 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계, 상기 선순위 스트림에 잔여 파워가 존재하면, 상기 선순위 스트림보다 작은 고유치를 갖는 다음 스트림에 상기 잔여 파워를 추가로 할당하는 단계 및 상기 다음 스트림의 코드 레이트 및 상기 피드백 정보를 이용하여 상기 다음 스트림에서 전송할 심볼의 에너지 레벨을 결정하는 단계를 포함하는 잔여 파워를 이용하는 적응 변조 방법을 제공한다.

상기의 두번째 기술적 과제를 해결하기 위하여, 본 발명은 복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 상기 스트림들마다 소정의 파워를 할당하는 초기화부, 상기 스트림들마다 상기 고유치가 큰 순서대로 상기 스트림들의 코드 레이트 및 수신단의 피드백 정보를 이용하여 상기 스트림들에서 전송할 심볼의 에너지 레벨을 결정하는 적응적 심볼 생성부, 상기 스트림들 중 임의의 스트림에서 심볼의 에너지 레벨을 결정한 후에 상기 임의의 스트림에 잔여 파워가 존재하면, 상기 임의의 스트림보다 작은 고유치를 갖는 다음 스트림에 상기 잔여 파워를 추가로 할당하는 파워 재할당부를 포함하는 잔여 파워를 이용하는 적응 변조 장치를 제공한다.

본 발명은 MIMO BIC-OFDM 시스템에서 더욱 향상된 성능을 제공하는 적응 변조 방식에 특징이 있는데, 특히 종래에 큰 사적인 코딩 이득을 이용하는 방식과 달리, 각 스트림에 사용되는 코드 레이트를 고려하여 비트 할당 기법을 더욱 최적화하고 있으며, 각 스트림에서 할당되고 남은 잔여 파워를 하위 스트림에 전달시켜줌으로써 파워 효율을 증가시키는 특징이 있다.

이하에서는 도면을 참조하여 본 발명의 바람직한 실시예를 설명하기로 한다.

도 1a 및 도 1b는 본 발명의 일 실시예에 따른 MIMO BIC-OFDM 시스템을 도시한 것이다. 도 1a 및 도 1b는 송신 안테나 개수 $N_t=4$, 수신 안테나 개수 $N_r=4$인 적응 변조 MIMO BIC-OFDM 시스템을 가정한 것이다.

도 1a는 적응 변조 MIMO BIC-OFDM 시스템의 전송단을 도시한 블록도이다.

도 1a를 참조하면, 기지국에서 전송된 사용자 정보(Data of user 1, 2, 3, ... , K)는 총 N_t개 즉, 4개의 스트림을 형성한다. 여기서 i번째 스트림은 i번째 안테나를 통해 송신되는 사용자 정보를 의미한다.

스케쥴러(100)는 복수의 사용자 정보를 한정된 스트림(1, 2, 3, 4)으로 출력한다.

다음, 피드백 정보를 통해 현재의 채널 상태를 파악하고 이를 바탕으로 하여 데이터 전송률과 파워 레벨을 조절한다. 전송률과 파워가 결정되면, 각 스트림 별로 BICM 방식을 따라 변조된 데이터는 N_c개의 부반송파를 갖는 직교 주파수 다중 분할 (OFDM) 변조 기법을 통해 전송된다.

채널 정보를 제공하는 피드백 채널은 보통 좋은 성능의 채널 코딩과 낮은 변조 레벨을 사용함으로써 신뢰성이 높은 전송을 보장한다. 따라서 본 발명에서는 에러가 없는 이상적인 피드백 채널을 가정한다.

AMC 블록(110)에서는 각 스트림에 전송할 비트를 할당하여 출력한다.

복수의 BICM들(120)은 각 스트림의 코드 레이트에 따라 데이터의 크기를 확장시킨다.

복수의 IFFT들(130)은 각 스트림에 대해 역 고속 푸리에 변환을 수행한다.

여기서 BICM 기법을 사용함으로써 주파수 선택적 채널에서, 채널 코딩을 통하여 주파수 도메인에서의 다이버시티 이득을 형성할 수 있다. OFDM은 2진 길쌈 부호기와 심볼 매퍼를 비트 단위 인터리버를 통해 결합함으로써 구현된다. BIC-OFDM은 BICM 기법에 역 고속 푸리에 변환 (IFFT)와 고속 푸리에 변환 (FFT)의 변조 방식을 통해 구현된다.

도 1b는 적응 변조 MIMO BIC-OFDM 시스템의 수신단을 도시한 블록도이다.

복수의 FFT들(140)은 각 수신 안테나로 수신된 신호에 대해 고속 푸리에 변환을 수행한다. 가중치 적용부(150)는 각 스트림에 가중치를 적용하고, 디맵퍼/역섞기(160)은 할당 정보에 따라 각 스트림을 결합하고, 디매퍼(170), 디인터리버(180) 및 디코더(190)는 원래의 비트를 복원한다.

도 2는 스트림의 갯수가 N인 경우에 본 발명의 블록도이다.

초기화부(200)는 복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 스트림들마다 소정의 파워를 할당한다. 또한, 초기화부(200)는 수신단의 피드백 정보로부터 스트림별 고유치를 연산할 수 있다. 또한, 초기화부(200)는 수신단의 피드백 정보로부터 적응적 심볼을 생성부(210)에서 생성된 심볼이 전송되는 채널의 상태를 검출하고, 검출된 채널의 상태에 따라 소정의 파워를 결정할 수 있다.

적응적 심볼 생성부(210)는 스트림들마다 고유치가 큰 순서대로 스트림들의 코드 레이트 및 수신단의 피드백 정보를 이용하여 스트림들에서 전송할 심볼의 에너지 레벨을 결정한다. 이때, 심볼은 사용자 정보에 포함된 복수의 비트들로 구성되는 비트 단위의 정보이다.

또한, 적응적 심볼 생성부(210)는 스트림들의 코드 레이트에 따른 코딩 오프셋을 이용하여 스트림들에서 전송할 심볼의 에너지 레벨을 보상할 수 있다. 바람직하게는, 코딩 오프셋은 소정의 비트 에러율을 기준으로 스트림들의 코드 레이트에 대응하는 오프셋을 미리 저장한 코딩 오프셋 테이블을 참조하여 결정될 수 있다. 이때, 출력 스트림 1 내지 N은 심볼의 에너지 레벨 결정 과정과 잔여 파워의 추가 할당 과정을 완료한 스트림들이다.

파워 재할당부(220)는 스트림들 중 임의의 스트림에서 심볼의 에너지 레벨을 결정한 후에 위 스트림에 잔여 파워가 존재하면, 위 스트림보다 작은 고유치를 갖는 다음 스트림에 잔여 파워를 추가로 할당한다.

바람직하게는, 도 2의 적응적 변조 장치는 도 1a의 AMC 블럭(110) 내에서 구현할 수 있다.

도 3은 본 발명의 흐름도이다.

먼저, 복수의 스트림들을 고유치가 큰 순서대로 정렬하고, 상기 스트림들마다 소정의 파워를 할당한다(300 과정). 이 과정(300 과정)은 수신단의 피드백 정보로부터 심볼이 전송되는 채널의 상태를 결정하고, 채널의 상태에 따라 소정의 파워를 결정하는 과정을 포함할 수 있다. 또한, 이 과정(300 과정)은 수신단의 피드백 정보로부터 스트림별 고유치를 연산하는 과정을 포함할 수 있다.

다음, 스트림들 중 고유치가 큰 순서대로 스트림을 선택하고, 선택된 스트림의 코드 레이트 및 수신단의 피드백 정보를 이용하여 선택된 스트림에서 전송할 심볼의 에너지 레벨을 결정한다(310 과정). 이 과정(310 과정)은 사용자 정보로부터 전송할 심볼을 생성하는 과정을 포함할 수 있다. 이때, 심볼의 에너지 레벨을 결정하는 과정(310 과정)은 최초에 스트림마다 할당된 파워를 적어도 일정부분 소모하게 된다.

선택된 스트림에서 심볼의 에너지 레벨을 결정하는 과정이 완료되면, 선택된 스트림에 잔여 파워가 존재하는지 판단한다(320 과정).

선택된 스트림에 잔여 파워가 존재하면, 선택된 스트림보다 작은 고유치를 갖는 다음 스트림에 잔여 파워를 추가로 할당한다(330 과정). 반면에, 선택된 스트림에 잔여 파워가 존재하지 않으면, 다음 과정(340 과정)으로 진행한다.

다음, 모든 스트림에 대해 위 과정(310~330 과정)이 완료되었는지 판단한다(340 과정). 이때, 모든 스트림에 대해 위 과정(310~330 과정)이 완료되지 않았다면, 다음 스트림으로 넘어가서(350 과정) 위 과정(310~330 과정)을 반복한다(340 과정).

반면에, 모든 스트림에 대해 위 과정(310~330 과정)이 완료되었으면, 모든 절차를 종료한다(340 과정).

여기서, 고유치가 가장 큰 스트림을 선순위 스트림, 선순위 스트림보다 고유치가 작은 스트림을 다음 스트림, 다음 스트림보다 고유치가 작은 스트림을 후순위 스트림으로 정의할 수 있다.

도 4는 도 3의 에너지 레벨 결정 과정(310 과정)의 상세 흐름도이다.

먼저, 피드백 정보로부터 스트림별 심볼이 전송되는 채널의 상태를 파악한다(411 과정).

다음, 스트림별 코드 레이트 및 채널 상태에 따라 코딩 오프셋 테이블의 특정 위치를 참조한다(412 과정).

다음, 스트림별 코드 레이트에 따른 코딩 오프셋을 결정한다(413 과정).

마지막으로, 결정된 코딩 오프셋으로 해당 스트림에서 전송되는 심볼의 에너지 레벨을 보상한다(414 과정).

도 5는 도 3의 에너지 레벨 결정 과정(310 과정)에서 사용되는 코딩 오프셋 테이블의 일 예를 도시한 것이다.

도 5는 본 발명의 일 실시예에 따른 MIMO BIC-OFDM 시스템에서 사용되는 여러 채널 코드와 변조방식에 해당하는 코딩 오프셋을 계산한 표이다. 여기에서, 가장 좌측열은 스트림의 코드 레이트를 나타내고, 가장 상단의 행은 여러가지 변조방식을 나타낸다. 여기에 기재된 코딩 오프셋은 비트 에러율(BER) 10^{-4}을 기준으로 하여 실험적으로 구해진 값으로 단위는 데시벨(dB)이다.

본 발명을 수학식으로 설명하면 다음과 같다.

n번째 부반송파에서의 수신 신호는 다음 수학식 1과 같다.

수학식 1

$$\mathbf{r}_n = \begin{bmatrix} r_1^n \\ \vdots \\ r_{N_r}^n \end{bmatrix} = \mathbf{H}_n \mathbf{x}_n + \mathbf{z}_n$$

$$= \begin{bmatrix} H_{11}^n & \cdots & H_{1N_t}^n \\ \vdots & \ddots & \vdots \\ H_{N_r1}^n & \cdots & H_{N_rN_t}^n \end{bmatrix} \begin{bmatrix} x_1^n \\ \vdots \\ x_{N_t}^n \end{bmatrix} + \begin{bmatrix} z_1^n \\ \vdots \\ z_{N_r}^n \end{bmatrix}$$

위 수학식 1에서, H_{ij}^n은 n번째 부반송파에서, i번째 송신 안테나와 j번째 수신 안테나 사이의 채널 푸리에 응답을 가리킨다. z_j^n은 σ_x^2의 분산을 따르는 독립 일치 분포의 복소 가우시안 잡음이며, x_j^n은 σ_x^2의 분산을 갖는 i번째 송신 안테나로 전송되는 심볼을 의미한다. 전송단에서 사용되는 총 파워는 $P=N_t\sigma_x^2$이라고 가정한다.

사용 가능한 채널 코드와 최대 성상의 크기는 각각 V와 $2^{m_{max}}$로 가정한다. 각 전송에서 i번째 스트림은 V개의 서로 다른 채널 코드 중의 하나를 택해 부호화 과정을 거치고, i번째 스트림의 n번째 부반송파는 $b_n^i \in \{m_1, m_2, \cdots, m_{max}\}$ $(n=1,2,\cdots,N_c)$ 개의 비트를 전송한다. $\{b_1^i, b_2^i, \cdots, b_{N_c}^i\}$를 i번째 스트림의 비트 벡터 b_i라 하면, i번째 스트림에서의 스펙트럼 효율 (spectral efficiency)은 다음 수학식 2와 같이 표현된다.

수학식 2

$$R_i(\mathbf{b}_i, C_{v_i}) = R_c(C_{v_i}) \frac{1}{N_c} \sum_{n=1}^{N_c} b_n^i$$

위 수학식 2에서, $R_c(C_{v_i})$는 채널 부호기 C_{v_i}에 의해 선택된 코딩 레이트를 의미한다. 채널 부호기 C_{v_i}와 n번째 부반송파에 할당되는 비트 수 b_n^i는 AMC 기법에 의해 결정된다.

좀 더 자세히 설명하면, SVD 기법을 통해 n번째 부채널 H_n을 다음 수학식 3과 같이 표현할 수 있다.

수학식 3

$$\mathbf{H}_n = \mathbf{V}_n \mathbf{\Sigma}_n \mathbf{U}_n^*$$

위 수학식 3에서, V_n과 U_n은 유니터리 (unitary) 행렬이고, Σ_n은 H_n의 고유치들을 성분으로 갖는 대각 행렬이다. SVD 정보를 알면, MIMO 채널 H_n은 U_n과 V_n^*을 이용한 필터링을 통해 대각화될 수 있다. 수신단에서 수신된 신호의 등가 모델은 다음 수학식 4와 같이 나타낼 수 있다.

수학식 4

$$\begin{aligned} \mathbf{y}_n &= \mathbf{V}_n^*(\mathbf{H}_n \mathbf{U}_n \mathbf{x}_n + \mathbf{z}_n) \\ &= \mathbf{\Sigma}_n \mathbf{x}_n + \mathbf{V}_n^* \mathbf{z}_n \end{aligned}$$

위 수학식 4에서와 같이, x_n은 n번째 부반송파에서 전송되는 심벌을 가리킨다. x_n을 제어하는 방법 중에 가장 일반적인 것은 고유모드 (eigenmode)를 통한 파워 할당 기법이다. 최적화된 파워 할당 기법은 잘 알려진 공간 도메인에서의 water-filling 기법을 통해 얻어진다. 하지만 이는 많은 복잡도를 필요로 한다.

한편, 비트 할당 기법은 더 큰 고유치를 갖는 모드에 더 많은 비트를 할당하는 방법이다. 본 발명의 다른 실시예에 따르는 MIMO BIC-OFDM 시스템에서의 비트 할당 방법을 이용하여 전송률을 극대화시킨다. 이는 시-공간 벡터 코딩 (STVC: spatio-temporal vector coding) 기법을 통하여 채널 용량 (capacity) 관점에서 최적화하는 것이다. 그러나 이 방법은 N_rN_c×N_tN_c 차원에서 SVD를 수행해야 하므로 복잡도가 매우 크다. 또한 N_r×N_t 차원의 SVD를 N_c번 수행하는 알고리즘은 N_c가 커짐에 따라 최적화된 성능을 거의 따라간다는 사실이 알려져 있다. 따라서, 본 발명은 복잡도를 줄이기 위해 후자의 방법을 택한다. 이러한 형태의 레이트 극대화 방법은 다음 수학식 5와 같이 수식화할 수 있다.

수학식 5

$$\begin{aligned} \text{maximize} \quad & \sum_{n=1}^{N_c} R_n(C_n) \sum_{m=1}^{N_t} b_{n,m}(\lambda_{n,m}) \\ \text{subject to} \quad & \sum_{n=1}^{N_c} \sum_{m=1}^{N_t} \varepsilon_{n,m} = N_c N_t \varepsilon_x \\ & P_t \le P_x \end{aligned}$$

위 수학식 5에서 볼 수 있는 바와 같이, 종래의 레빈 캠펠로(Levin-Campello ; LC) 알고리즘은 최대 성상 크기가 제한되지 않은 경우에는 각 부반송파의 균등한 파워 분배를 통해 최적화된 비트 할당 벡터 b_j와 코드 색인 c_j^*를 얻는다. 그러나, 지원 가능한 최대 성상 크기가 제한되는 실용적인 상황에서는 위에서 언급한 균여 파워 문제가 발생하여 성능이 떨어진다. 각 부반송파별로 수행되는 SVD는 고유치를 내림차순으로 정렬하는데, 높은 SNR에서는 이로 인하여 m_max의 비트 수가 할당될 확률이 낮은 색인의 스트림에 집중된다. 이 경우 m_max 비트를 전송하는 데 필요한 에너지가 E_x보다 작아지므로, $\sum_{m=1}^{N_t} \varepsilon_{n,m}$ 역시 $N_t\varepsilon_x$ 보다 작은 값을 갖는다. 따라서 $N_t\varepsilon_x - \sum_{m=1}^{N_t} \varepsilon_{n,m}$ 만큼의 잔여 파워가 생겨난다.

반대로 작은 고유치를 갖는 높은 색인의 스트림은 추가적인 파워를 할당하게 되면 더 많은 정보를 실을 수 있는 여지가 존재한다. 따라서 낮은 색인의 스트림에서 생겨나는 여분의 파워를 다른 스트림으로 전달시켜주면 시스템의 전체 전송률을 증가시킬 수 있다. 이를 다음 수학식 6과 같이 수식으로 정리할 수 있다.

수학식 6

$$\varepsilon_{n,i} = \left(\varepsilon_x - \frac{1}{N_t} \sum_{m=1}^{N_t} \varepsilon_{n,m} \right)_+ + \varepsilon_x$$

$$(x)_+ = \begin{cases} x & \text{if } x \geq 0 \\ 0 & \text{if } x < 0 \end{cases}$$

위 수학식 6에서와 같이, LC Efficientizing (EF) 알고리즘과 E-Tightening (ET) 알고리즘에 적용하면, 작은 고유치를 갖는 스트림에 기존보다 더 많은 비트를 할당하므로 전송률을 증가시킬 수 있다.

한편, 이산 비트 할당 알고리즘 (MIMO(BLCM/RPD($E_{-}s$))에 대하여 살펴보면, 다음 수학식 7과 같다.

수학식 7

1) Obtain $\Sigma_n = \mathbf{V}_n^{\dagger}\mathbf{H}_n\mathbf{U}_n$ for $n = 1, \cdots, N_t$

2) $\mathcal{E}_1 = \mathcal{E}_s$

3) FOR $1 \leq k \leq N_t$

a) $\tilde{\mathbf{H}} = [\mathbf{x}_{1,k}, \cdots, \mathbf{x}_{N_t,k}]$

b) obtain $\mathbf{b}_k$ by performing the DBLC/CS($\tilde{\mathbf{H}}, \sigma_n$) with the constraint of $\hat{\mathcal{E}}_k$

c) $\hat{\mathcal{E}}_{k+1} = \left(\hat{\mathcal{E}}_k - \frac{1}{N_t}\sum_{n=1}^{N_t}\mathcal{E}_{n,k}\right)_+ + \mathcal{E}_s$

위 수학식 7에서는 c)의 값을 별도의 추가적인 계산량 없이 ET 알고리즘을 통해 바로 얻을 수 있다.

또한, 부호화 시스템에서의 이산 비트 할당 알고리즘 (DBLC/CS(H,$\sigma_{-}n$))에 대하여 살펴보면, 다음 수학식 8과 같다.

수학식 8

1) FOR $1 \leq n \leq V$

a) Obtain $\mathbf{b}_n$ with GLCA($\mathbf{H}_n, \sigma_n$)

b) Compute $R_n(\mathbf{b}_n, \mathcal{C}_n)$

2) $n^* = \arg\max_{1 \leq n \leq V} R_n(\mathbf{b}_n, \mathcal{C}_n)$

3) $\mathbf{b} = \mathbf{b}_{n^*}$,

위 수학식 8에서 와 같이 본 발명의 generalized LC 알고리즘 (GLC)을 살펴본다. 종래의 LC 알고리즘은 가정 잘 알려진 최적화된 비트 할당 알고리즘 중의 하나로, 고정된 채널 갭 (Γ: gap)을 통한 근사적인 크링 이득을 이용하여 알고리즘을 수행한다. M-QAM에서 갭 근사화를 통하여 구한 레이지 함수는 다음 수학식 9와 같다.

수학식 9

$$\mathcal{E}_n(b_n) = (2^{b_n} - 1)\frac{\sigma_n^2}{|H_n|^2} \cdot \Gamma$$

위 수학식 9에서 허림 비부호화 시스템에서는 채널 갭이 일정하다는 가정을 따르지만, 부호화 시스템에서는 적용된 채널 코드의 특성에 따라 갭에 해당하는 값이 변하게 된다. 코드 레이트 R_c를 적용한 시스템의 갭은 다음 수학식 10과 같이 표현된다.

수학식 10

$$\Gamma_{coded} = 10\log_{10}(2^{mR_c} - 1) - \Psi(m, R_c)\ (\mathrm{dB})$$

여기서 Ψ(m,R_c)는 코드 레이트와 성상의 크기를 각각 R_c와 2^m으로 적용하였을 때, 비트 에러율 (BER: bit error rate) 10^{-5}을 만족시키기 위해 필요한 SNR 값이다. 각 스트림에 채널 코딩이 적용되면 각 부반송파에 실리는 비트 수가 더 이상 정보 비트 수를 의미하지 않게 되므로, 이 경우 채널 갭은 다음 수학식 11과 같이 정의되는 코딩 오프셋으로 대체되어야 한다.

수학식 11

$$\gamma_c(m) = \Psi(m, R_c(C_c)) - 10\log_{10}(2^m - 1) \quad (dB)$$

이 코딩 오프셋을 새로운 파라미터로 고려한 에너지 함수는 다음 수학식 12와 같다.

수학식 12

$$\mathcal{E}_n(b_n; c) = 2^{b_n - 1}\frac{\sigma_n^2}{|H_n|^2} \cdot \gamma_c(b_n)$$

도 6은 종래의 변조 방법 및 본 발명에 따른 성능을 도시한 그래프이다.

여기서는 N_c=64개, CP=16샘플로 구성되는 OFDM 시스템을 고려하였고, 채널은 5개의 탭으로 구성된, 파워가 지수적으로 감소하는(exponentially decaying) 채널을 사용하였다. 64-상태 RCPC 코드를 사용하였고, 목표 프레임 에러율 (FER: frame error rate)을 1%로 설정하였다. 이에 해당하는 비트 에러율은 10^{-5}이다.

도 6은 N_t=N_r=4의 단일 사용자가 사용하는 전우의 성능을 나타낸 그래프이다. 가장 오른쪽의 *(asterisk)로 구별된 성능 곡선(conventional SVD)은 종래의 워터 필링(water-filling) 비트 할당 시스템의 스루풋(throughput)을 나타낸다. 즉, 성능 곡선(conventional SVD)의 시스템은 종래의 LC 알고리즘에 고정된 채널 갭(gap)을 사용하고, 전역 파워 분배를 고려하지 않은 시스템이다.

본 발명에 따라 파워를 재할당하는 시스템의 성능 곡선(RPD with GLCRPD 본 발명에 따라 코드 레이트를 고려하여 심볼의 에너지를 보상하는 시스템의 성능 곡선(RPD)은 종래 시스템의 성능 곡선(conventional SVD)보다 스루풋(throughput)이 높다는 것을 알 수 있다. 즉, 본 발명에 따르면 스펙트럼 효율 18 bps/Hz를 기준으로 약 4dB 정도 성능이 향상됨을 알 수 있다.

대부분의 실용적인 시스템은 모두 지원 가능한 최대 성상 크기에 제한을 두기 때문에, 높은 SNR 범위에서 전역 파워의 분배는 매우 일반적이다. 본 발명은 이를 보상한 비트 할당 기법을 제안함으로써 특히 높은 SNR 범위에서 큰 성능 이득을 얻을 수 있다.

바람직하게는, 본 발명의 전역 파워를 이용하는 비용 변조 방법을 컴퓨터에서 실행시키기 위한 프로그램을 컴퓨터로 읽을 수 있는 기록매체에 기록할 수 있다.

본 발명은 소프트웨어를 통해 실행될 수 있다. 소프트웨어로 실행될 때, 본 발명의 구성 수단들은 필요한 작업을 실행하는 코드 세그먼트들이다. 프로그램 또는 코드 세그먼트들은 프로세서 판독 가능 매체에 저장되거나 전송 매체 또는 통신망에서 반송파와 결합된 컴퓨터 데이터 신호에 의하여 전송될 수 있다.

본 발명은 도면에 도시된 일 실시예를 참고로 하여 설명하였으나 이는 예시적인 것에 불과하며 당해 분야에서 통상의 지식을 가진 자라면 이로부터 다양한 변형 및 실시예의 변형이 가능하다는 점을 이해할 것이다. 그러나, 이와 같은 변형은 본 발명의 기술적 보호범위내에 있다고 보아야 한다. 따라서, 본 발명의 진정한 기술적 보호범위는 첨부된 특허청구범위의 기술적 사상에 의해서 정해져야 할 것이다.

발명의 효과

상술한 바와 같이, 본 발명에 의하면, 스트림마다 다른 코드 레이트를 고려하여 심볼의 에너지 레벨을 결정하고, 같은 파워를 코딩율이 낮은 스트림에 추가로 할당함으로써, 잔여 파워의 낭비를 막고, 목표 비트 에러율의 제약조건이 있는 경우나 성상의 최대크기가 제한된 경우에도 시스템의 성능을 향상시킬 수 있으며, 구현 과정에 추가적인 계산량을 요구하지 않는 효과가 있다.

도면의 간단한 설명

도 1a 및 도 1b는 본 발명의 일 실시예에 따른 MIMO BIC-OFDM 시스템을 도시한 것이다.

도 2는 본 발명의 블럭도이다.

도 3은 본 발명의 흐름도이다.

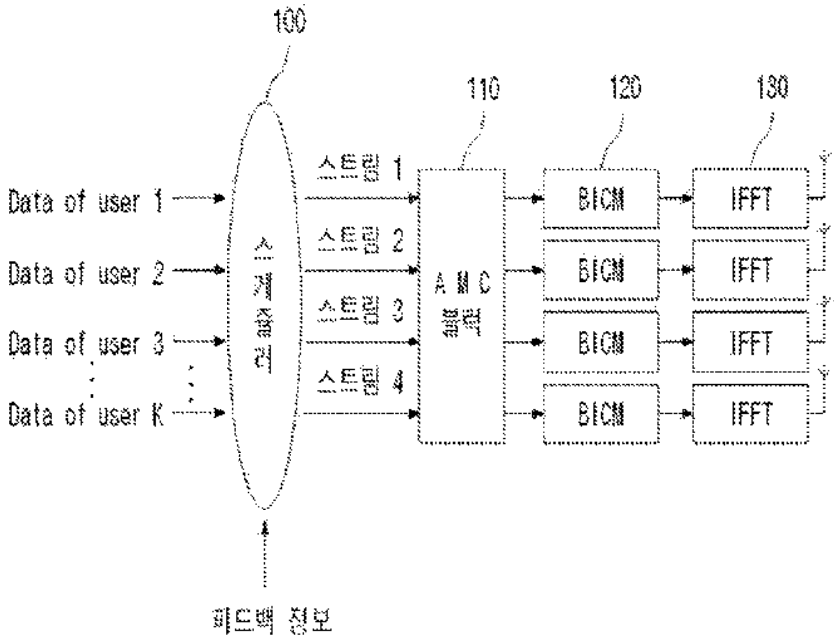
도 4는 도 3의 에너지 레벨 결정 과정의 상세 흐름도이다.

도 5는 도 3의 에너지 레벨 결정 과정에서 사용되는 코딩 오프셋 테이블의 일 예를 도시한 것이다.

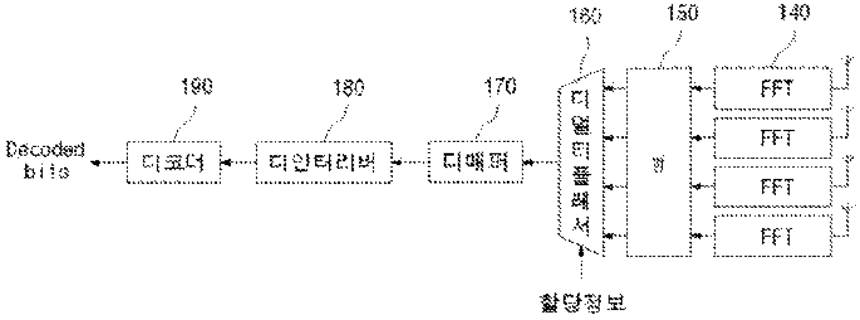
도 6은 종래의 변조 방법 및 본 발명에 따른 성능을 도시한 그래프이다.

도면

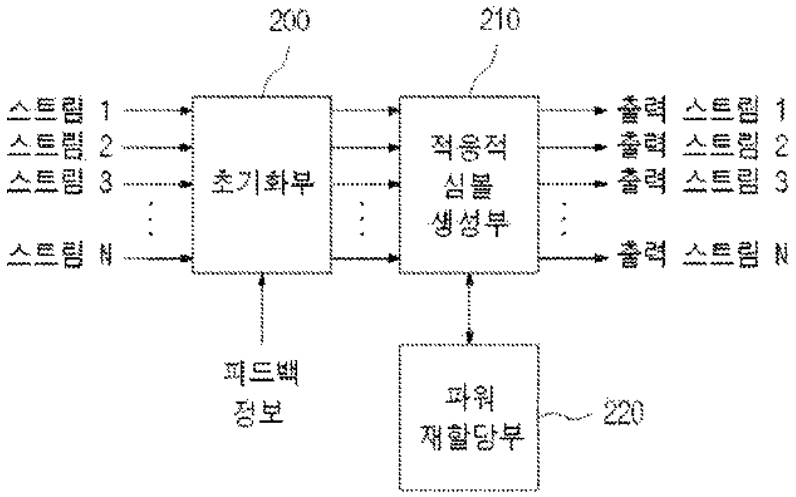
도면1a


100
110
120
130
Data of user 1
Data of user 2
Data of user 3
Data of user K
스케줄링
스트림 1
스트림 2
스트림 3
스트림 4
AMC 블록
BICM
BICM
BICM
BICM
IFFT
IFFT
IFFT
IFFT
피드백 정보

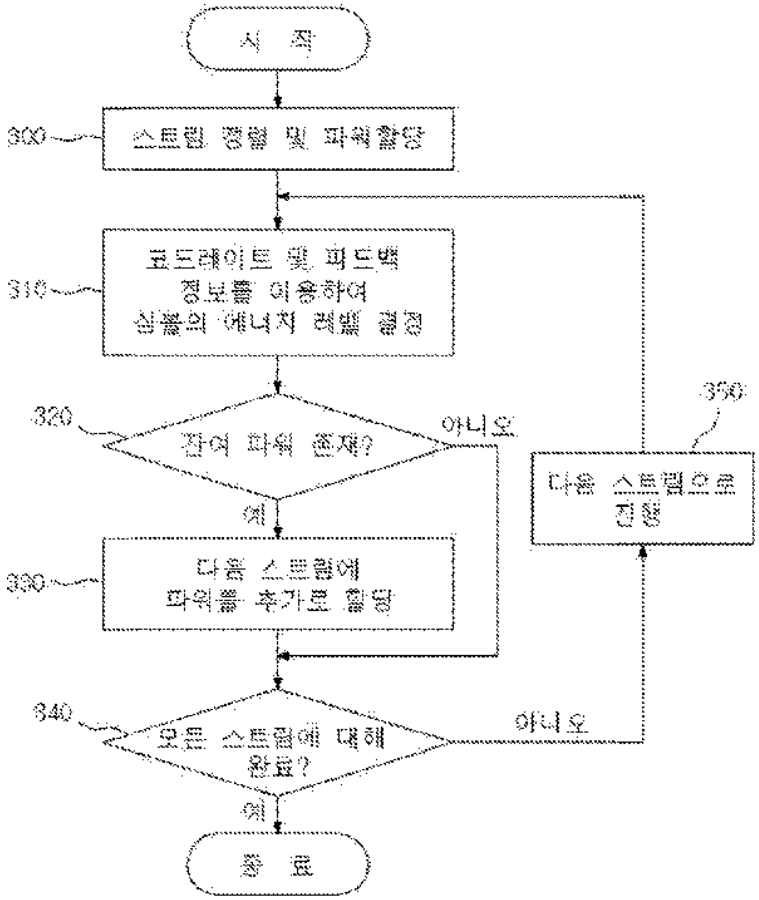
도면1b


160
150
140
190
180
170
Decoded bits
디코더
디인터리버
디매퍼
디멀티플렉서
FFT
FFT
FFT
FFT
할당정보

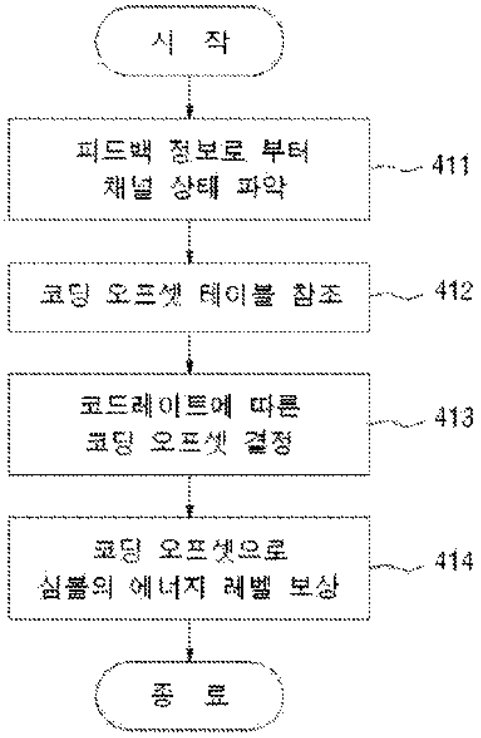
도면2


200
210
스트림 1
스트림 2
스트림 3
스트림 N
초기화부
적응적 심볼 생성부
출력 스트림 1
출력 스트림 2
출력 스트림 3
출력 스트림 N
피드백 정보
파워 재할당부
220

도면3


시 작
300
스트림 정렬 및 파워할당
310
코드레이트 및 피드백 정보를 이용하여 심볼의 에너지 레벨 결정
320
잔여 파워 존재?
아니오
예
330
다음 스트림에 파워를 추가로 할당
340
모든 스트림에 대해 완료?
아니오
예
350
다음 스트림으로 진행
종 료

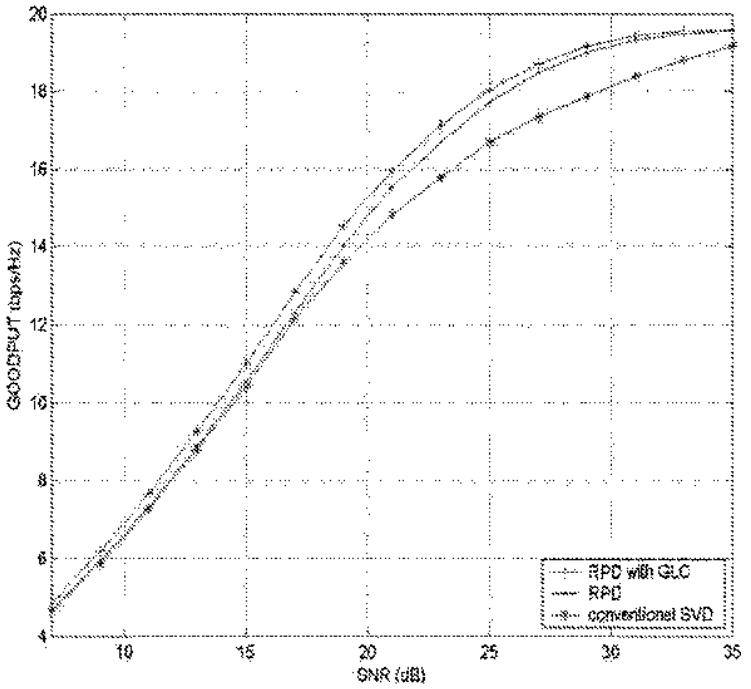
도면4



도면5

	BPSK	4QAM	8PSK	16QAM	32QAM	64QAM
1	8.40	6.62	7.98	6.329	6.32	6.30
5/6	4.5	2.86	3.98	2.40	2.31	2.06
4/5	4.15	2.43	3.53	2.01	1.89	1.65
3/4	2.97	1.40	2.25	0.94	1.00	0.37
2/3	2.09	0.42	1.00	-0.39	-0.46	-1.09
1/2	0.37	-1.34	-1.35	-2.61	-2.61	-3.96

도면6



Method and apparatus for multi-carrier transmission

Abstract

The present invention provides a method and apparatus for multi-carrier transmission of data. The method comprising the steps of: providing a stream of data, encoding the stream of data to create a plurality of complex values, assigning each of the plurality of complex values to one of a plurality of sub-channels, assigning a separate value to each of the plurality of sub-channels, multiplying each of the plurality of sub-channels with the assigned separate value to generate a multiplied value for each of the plurality of sub-channels, modulating the multiplied value of each of the plurality of sub-channels to a sub-carrier to generate a modulated signal for each of the sub-channels, and simultaneously transmitting the modulated signals.

Classifications

H04L27/2602 Signal structure

CN100471191C
China
Download PDF Find Prior Art Similar
Other languages
Chinese
Inventor
安德列·诺尔巴雷托

Worldwide applications
2002 WO ~~AU~~ ~~CA~~ ~~JP~~ ~~CN~~ ~~KR~~ 2004 ~~US~~

Application CNB028251210A events

2001-12-17

Priority to EP01811232

2001-12-17

Priority to EP01811232.6

2002-11-21

Application filed by 国际商业机器公司

2005-04-27

Publication of CN1611048A

2009-03-18

Application granted

2009-03-18

Publication of CN100471191C

Info

Patent citations (7)

Cited by (5)

Legal events

Similar documents

Priority and Related Applications

External links

Espacenet
Global Dossier
Discuss

Description

Be used for the method and apparatus that multicarrier sends

Technical field

The present invention relates to the method and apparatus that a kind of multicarrier that is used for data sends.Particularly, the present invention relates to a kind of effective transmit diversity scheme that is particularly suitable for wireless transmission.

Background technology

Proposed to be used for the multi-carrier modulation of wireless environment, it not only is used for broadcasted application, as at European Digital Video Broadcasting (DVB, the Europe digital video broadcasting) defines in the standard, and be used for high-rate wireless local area network (LAN) (W-LAN), as defining at North America IEEE 802.11a with in European HIPERLAN-2 standard, above-mentioned application all depends on the OFDM (OFDM) of coding.These standards are supported the high-data-rate wireless transmission up to 54Mbps.

Being hidden in OFDM thought afterwards is that the data flow that will arrive is divided into several parallel data streams than low rate (so symbol period T_sLonger) and in different subchannels, send each data flow.The use 1/T of being separated by $_s$Different sub carrier send these data flow.By such chooser carrier spacing, subchannel is a quadrature when suitably being sampled, and allows the spectrum overlapping of subchannel, thus maximization transmission frequency spectra efficient.

The advantage of OFDM is its adaptive faculty to intersymbol interference (ISI), and this ISI is caused by the multipath transmisstion that often has in the wireless channel.This adaptive faculty can come the cyclic extensions signal to realize by using protection at interval, and this protection should be longer than the maximum delay of channel at interval.

Usually, frequency selects decline (promptly observing different declines on different frequencies) to show the characteristics of broadband wireless system.In coded OFDM, coding is striden the data bit on the different subcarriers, and this provides some protection for opposing frequency selective channel.But because side frequency is height correlation most probably, therefore this protection also is limited, thereby the deep fading often influences several subchannels.

An optional method of opposing decline is to use a plurality of antennas to obtain space diversity.In order to obtain enough diversity, the channel of different antennae must have low correlation, this means that these channels are apartly enough far away.In addition, each antenna also needs independent wireless front end, thereby has increased the cost of transceiver.These problems make more may only use a plurality of antennas

in the base station, therefore, in down link, must use diversity technique at the emission pusher side.

Static applications or low speed that high speed W-LAN system is positioned in the indoor environment move application.Use for this class, channel variation ground is very slow, for example, and with walking speed (1m/s), for carrier frequency f_c=5GHz, be T coherence time $_c$=25ms, this is corresponding to the mac frame of 12 2ms among the HIPERLAN/2.For static (portable) terminal, decline may continue the hundreds of millisecond.Use for data, repeat requests (ARQ) scheme or simple packet retransmissions present can be used to guarantee low packet loss and intimate error-free transmission automatically.But under above-mentioned channel condition, when this divided into groups by error free reception, grouping may be had to repeatedly be retransmitted or between retransmitting big delay be arranged, thereby has reduced throughput of system and increased transmission delay.

At United States Patent (USP) the 5th, 914, advised a kind of ofdm system (clustered OFDM) of so-called diversity in No. 933, wherein, the different subclass of subcarrier are assigned to each antenna continuously.The shortcoming of this system is: owing to adjacent subcarrier sends from the same antenna, and therefore be correlated with, so can obtain little frequency diversity.

United States Patent (USP) the 6th, 005 has been described the high-speed radio transmission system No. 876, and wherein, subclass is such, so that subcarrier is expanded equably on whole bandwidth.This can be counted as the antenna frequency hopping in the frequency domain.Consider the throughput of using repeat mechanism, this system has shortcoming.With regard to frequency diversity, this method has been represented a kind of progress, but considers the time diversity of using ARQ, even change subcarrier, benefits also seldom.

From above-mentioned clearly as can be known: be starved of a kind of existing standard that can be applicable to, such as standard based on OFDM, effective transmit diversity scheme.And, in order to improve transmission performances and to obtain higher reliability,, the reduction error rate should realize thereby also improving the throughput of data.

Summary of the invention

According to an aspect of the present invention, the method that provides a kind of multicarrier that is used for data to send, the method comprising the steps of:

Data flow is provided;

This data flow of encoding is to create a plurality of complex values;

In a plurality of complex values each distributed in a plurality of subchannels of one that form in two or more channels;

To be worth (separate value) separately and distribute in these a plurality of subchannels each;

Each of this a plurality of subchannels be multiply by each the product value (multiplied value) that produces these a plurality of subchannels mutually with the independent value of being distributed;

Each product value of these a plurality of subchannels is modulated to subcarrier, to produce each modulation signal of these two or more channels; With

Send each this modulation signal of these two or more channels simultaneously.

This method provides a kind of effective transmit diversity scheme, as long as use a plurality of antennas, this scheme just has low additional complexity, thereby it can be applied to such as the existing standard based on the W-LAN standard of OFDM, and does not need or only need seldom revise just to can be applicable to existing standard.And, can realize the substantial reduction of the error rate.Therefore, higher throughput can realize.Therefore can provide the raising of transmission performance and higher reliability.

Basically, this method provides the frequency domain predistortion and has used a plurality of transmitting antennas to improve the frequency diversity of multicarrier system.This method also can be used to provide a kind of system with time diversity, can utilize this system with time diversity by high-

rise wrong controlled function (for example, automatic repeat requests (ARQ)), to improve data throughout.

The step that multiplies each other with the independent value of being distributed provides phase shift and/or the changes in amplitude in the subcarrier.By doing like this, the auto-correlation in the frequency domain becomes littler.In addition, can more effectively use applied sign indicating number.

If the difference from a subcarrier to the phase shift of next subcarrier is a constant, this is favourable.This influences the delay in the channel.Thereby, can more effectively carry out channel estimating at receiver side.

Each the step of value separately being distributed to a plurality of subchannels can comprise: be provided for the stochastic variable of value separately.Use stochastic variable to improve the frequency selectivity in the channel, simultaneously, employed sign indicating number becomes more effective.

Each the step of value separately being distributed to a plurality of subchannels can comprise: be provided for the constant amplitude with out of phase value of value separately.Owing to kept power division between the subcarrier, and can the appreciable impact transmission performance, so this is favourable.

Different phase values may belong to one group of possible fixed value, and this is owing to can simplify complex multiplication thereafter.

Data flow comprises grouping and is used for an independent value of each grouping that promptly, for each grouping, value is different separately.By doing like this, can realize dividing into groups to each with being worth fixed allocation separately, this has caused time diversity.

During in knowing a plurality of subchannels one channel gain, change the phase value of value separately, thereby separately value provide with a plurality of subchannels in the corresponding phase shift of phase reversal of one phase place, be favourable, this be since thereafter following advantage occur: but be coherent reception from the signal of different antennae.

When channel gain is known, promptly, when channel estimating is successful, revise the range value of value separately, thereby one amplitude in range value and a plurality of subchannel is proportional, be more favourable, this occurs owing to following advantage thereafter: but signal is coherent reception, and signal to noise ratio (snr) can be maximized.

Modulation step can comprise the OFDM modulation.This shows: proposed scheme can be applied to standard modulation techniques.

According to a second aspect of the invention, the device that provides a kind of multicarrier that is used for data to send, this device comprises:

Cell encoder is used for receiving data stream and creates a plurality of complex values;

Remove multiplexer, be used for each of a plurality of complex values is distributed in a plurality of subchannels of one that form two or more channels one;

Multiplication unit is used for each of a plurality of subchannels and independent value be multiply by each the product value that produces a plurality of subchannels mutually;

Modulator is used for each product value of a plurality of subchannels is modulated to subcarrier, to produce each modulation signal of two or more channels; With

Transmitter is used for launching modulation signal simultaneously through transmit antenna, and each of two or more channels has the transmit antenna of distributing to it.

Therefore, the embodiment of this aspect of the present invention uses and above-mentioned similar principle.

Description of drawings

By below with reference to exemplary accompanying drawing, to the preferred embodiments of the present invention carried out detailed, only as the description of example, wherein:

Fig. 1 shows the exemplary illustration according to multicarrier transmitter apparatus of the present invention;

Fig. 2 shows exemplary illustration to multicarrier transmitter apparatus in more abstract mode;

Fig. 3 shows the exemplary illustration of corresponding receiver;

Fig. 4 shows the figure of demonstration with respect to the data throughout of different transmission schemes.

Accompanying drawing provides for illustrative purposes, and accompanying drawing needn't be represented actual example of the present invention pari passu.

Embodiment

Though can be used for multiple multicarrier, the present invention sends application, but emphasis is described wireless system, be the application of WLAN (wireless local area network) (W-LAN), it is as using OFDM using OFDM (OFDM) among W-LAN standard IEEE 802.11a and the HIPERLAN-2.Before describing embodiments of the invention, some rudimentary knowledge are described according to the present invention.

Usually, the transmit diversity scheme that is proposed uses the symbol that sends on k subchannel on each subcarrier on the antenna A1 (to be also referred to as complex values $x_k(i)$) (be also referred to as independent value a with coefficient $_{L,k}$) multiplication.Expression formula i is corresponding to i OFDM symbol.Each is value a separately $_{L,k}$Comprise range value $\alpha_{L,k}$With phase value $\varphi_{L,k}$, as described in more detail below.Be worth a separately $_{L,k}$Can be counted as complex values.Use has at least two antenna A_1The system of (this means and have at least two channels 1) can obtain best result.Consider the situation of single receive antenna 52 as shown in Figure 3, k subchannel, through fast Fourier transform (FFT) received signal afterwards will be:

$$r_k(i) = h_{eq,k} x_k(i)$$

H wherein $_{Eq,k}$Be the gain of the equivalent channel that constitutes by all channels 1, be also referred to as equivalent channel gain $h_{Eq,k}$

This gain provides by following formula:

$h_{eq,k}=\sum_{l} a_{l,k} h_{l,k}$

Wherein, $h_{L,k}$Be the 1st antenna A_1Channel gain with k subchannel.Transmitting antenna A_1Quantity and independent value $a_{L,k}$Selection be transparent to receiver, and do not need extra signaling.Receiver receives and uses equivalent channel gain $h_{Eq,k}$The transmission signal x that revises $_k$(i), send from individual antenna A just as it.Therefore, receiver is only seen equivalent channel gain $h_{Eq,k}$If, and be worth a separately $_{L,k}$Also can be applicable to train leading (preamble), use traditional channel estimation technique as be known in the art, just can obtain equivalent channel gain $h_{Eq,k}$

For time diversity is provided, be worth a separately $_{L,k}$Should change in each grouping place.There are several diverse ways to select and n the corresponding independent value a of grouping $_{L,k}$(n).In first example, advise: by making

$a_{l,k}(n)=\alpha_{l,k}?\exp(j\varphi_{l,k}(n))$,

Then all independent values all have amplitude alpha $_{L,k}$And random phase, wherein, phase value $\varphi_{L,k}$(n) comprise the interval [0,2 π) in independent unified (independent uniform) stochastic variable.As the same transmitted power in the individual antenna system, amplitude can be chosen as if desired α l , k = (L) , Wherein L is antenna A_1Sum.Can illustrate: system compares with individual antenna, and the frequency diversity of this system is along with this is selected, and the correlation between the channel gain of promptly different subchannel k reduces and improves.This has caused the substance of the error rate to reduce.Perhaps, in second example, selecting range α at random $_{L,k}$Performance classes according to method first example, the use random phase is similar to second example.

As mentioned above, when the packet retransmissions present scheme used such as automatic repeat requests (ARQ), the time-varying characteristics of proposed transmit diversity scheme provide time diversity.This technology can merge with grouping at the receiver place to be used, and improves to realize further performance.Having the wrong grouping that is received should not be dropped.Not the vicious grouping of storage, but its retransmission version afterwards with same grouping made

up that ideal situation is to use high specific to merge.The grouping relevant with transmit diversity scheme merges the throughput that can improve the OFDM wireless system.This has caused the raising of capacity and the reduction of transmission delay, and can use in existing system.

Fig. 1 shows the exemplary illustration of multicarrier transmitter apparatus 2.Cell encoder 10 is at its input receiving data stream b, and provides a plurality of complex values x in its output place.Cell encoder 10 also is counted as Bit Interleave coded modulation (BICM) unit 10, and it comprises encoder 11 and mapper 12 here, itself or use phase shift keying (PSK) or use quadrature amplitude modulation (qam).For for simplicity, the interleaver unit between not shown encoder 11 and the mapper 12.The output of cell encoder 10 goes multiplexer 14 to link to each other with 2, and wherein, each goes multiplexer corresponding to a channel 1.The quantity of channel 1 can be higher than two, as shown in Figure 2.Below, only regard a channel as the function of each unit be identical.Go multiplexer 14 with a plurality of complex values x_kEach distribute to each of a plurality of subchannel k.Multiplication unit 16 links to each other with each of a plurality of subchannel k.Be worth a separately $_{L,k}$Be provided for multiplication unit, and as mentioned above, it is assignable.In each channel 1, a plurality of subchannel k link to each other with modulator 20.Modulator 20 comprises inverse fast fourier transformed (IFFT) unit 22, and it links to each other with multiplexer 24.Multiplexer 24 serializations are 22 signal flows that receive from inverse fast fourier transformed (IFFT) unit.Serialized signal flow is fed to cyclic extension unit 26.The output of cyclic extension unit 26, the output of modulator 20 just is fed to transmitter 30.Such transmitter 30 generally includes and sends or TX filter and RF (radio frequency) front end, and is not shown for simplification.Modulation signal s_1Can be through transmitting antenna A_1Send.Each channel 1 has its transmitting antenna A_1, A_2

Multicarrier transmitter apparatus 2 following operations.By cell encoder 10 data flow b is encoded into a plurality of complex values x.With a plurality of complex values x_kEach distribute among a plurality of subchannel k one.In addition, be worth a separately with one $_{L,k}$Distribute to each of a plurality of subchannel k.Although the possibility of several variations is arranged, can create each value a separately as mentioned above $_{L,k}$Simultaneously, can make independent value a $_{L,k}$Be fit to channel condition.As shown in Figure 1, with each of a plurality of subchannel k and the independent value a that is distributed $_{L,k}$Multiply each other, to produce each the product value m of a plurality of subchannel k $_{L,k}$It is illustrated by the multiplication symbol in

the multiplication unit 16.In modulator 20, the product value m of each of a plurality of subchannel $k_{L,k}$Be fed to inverse fast fourier transformed (IFFT) unit 22.After the processing that the serialization of using multiplexer 24 to carry out and use cyclic extension unit 26 are carried out, with modulation signal s_1Offer transmitter 30.Through distributing to the transmitting antenna A1 of each channel 1, A2 launches the modulation signal s of each channel 1 simultaneously $_1$

Fig. 2 shows the exemplary illustration of another embodiment of the multicarrier transmitter apparatus 2 with a plurality of channels 1.Adopt the vector representation data, shown in the letter of underscore.General 26S Proteasome Structure and Function is similar to Fig. 1's.Identical reference number is used for representing same or analogous parts.Using encoder 11, is N with length $_{Pack}$Data flow b (n) (being also referred to as input data sequence) to be encoded into code check be R_CN $_{Pack,c}$=N $_{Pack}$/ R_cIndividual sign indicating number bit, being divided into each then has N_cThe N of individual bit $_{Pack,c}$/ N_cIndividual piece c (i), it is corresponding to i OFDM symbol.Then, by using mapper 12 that it is mapped to K_d=N_c/ $\log_2$(M) individual QAM or Quadrature Phase Shift Keying (QPSK) symbol, it is also referred to as complex value vector $\underline{x}$(i), wherein M is a constellation size.For simplified representation, remove time index i, consider single OFDM symbol or complex value vector simultaneously $\underline{x}$Complex value vector $\underline{x}$Corresponding to the ofdm signal in the frequency domain.Introduced the K relevant with each subchannel $_p$Pilot tone and K_zZero subcarrier, and signal process K point inverse fast fourier transformed (IDFT), wherein, K=K_d+ K_p+ K_z, as in modulator 20 (not shown), realizing.For the time-domain signal of acquisition like this, a Cyclic Prefix that adds G sampling is as what realize in 26 (not shown) of the cyclic extension unit in being also included within modulator 20, so that eliminate up to T_G=GT_sThe multipath that postpones expansion disturbs, and wherein Ts is the sampling interval.The modulation signal s of gained $_1$Filtered, use transmitter 30 to convert thereof into radio frequency, and through transmitting antenna A_1Through the multipath channel emission.

Multicarrier transmitter apparatus 2 uses predistortion in frequency domain, this predistortion is as shown in the symbol of taking advantage of each subchannel k in the multiplexed unit 16.By with complex value vector $\underline{x}$Element and independent value vector $\underline{a}_l$Element multiply each other and carry out this predistortion.At k subcarrier and the 1st antenna A_1On the transmission signal be:

$x_{l,k}=a_{l,k}x_k$

Receiver is carried out reverse operating.Received signal is filtered, is switched to base band and with $1/T_s$Speed sampled.Remove cyclic extensions and carry out discrete Fourier transform (DFT) (DFT).Remove the zero-sum pilot sub-carrier, and the signal at k subchannel is after this operation:

$y_k=h_kx_k+v_k$

Wherein, h_kBe equivalent channel gain, and v_kBe that variance is N_0The complex noise component.Based on channel estimating Can balanced received signal estimate with picked up signal:

$\hat{x}_k = y_k h\hat{}_k$

Use symbol and channel vector to estimate respectively With Can obtain a yard bit Log-likelihood ratio (log-likehood ratio), can use for example soft input Viterbi decoder that it is decoded.

Before each packet, send known leading so that allow receiver to carry out synchronously and channel estimating, and the initial acquisition of frequency shift (FS).Also available independent value $a_{L,k}$Revise leading.Because ofdm system is very sensitive to estimated frequency error, introduce a plurality of pilot sub-carriers to improve the estimation and the correction of the frequency shift (FS) in a grouping.IEEE 802.11a supports variable bit rate, and it can be realized with different encoding rates by different modulation schemes.

In receiver, the frequency-region signal of each reception antenna can multiply each other by element with vector, and is accumulated in together from the signal of all reception antennas.Can resemble the high specific merging that for example is used to maximize signal to noise ratio (snr) according to Merge Scenarios, select weight vectors.

Fig. 3 shows the exemplary illustration that can combine the receiver of using 50 with the multicarrier transmitter apparatus 2 shown in Fig. 1 and Fig. 2.Receiver 50 comprises single receive antenna 52, demodulator unit 54 and 56 and decoder 58, and above-mentioned each unit connects being aligned.Demodulator unit 54 and 56 known technologies that use such as relevant or Differential Detection come demodulated received signal, for

example ofdm signal.Decoder 58 is used as error correction decoder.Should be understood that: available a plurality of receivers 50 receive and send signal s_1Predistortion is transparent to receiver 50 in theory, and receiver 50 needn't know whether and use transmission diversity, and attempt estimating equivalent channel gain h simply $_{Eq, k}$

In Fig. 4, shown performance improvement by the proposed transmit diversity scheme of using random phase.Consider to have the system of 4 transmitting antennas, and and proposed transmit diversity scheme that will be shown in curve IV compare with individual antenna system shown in curve I and the known transmission diversity scheme shown in curve II and III.At length, curve II shows the delay diversity scheme, and curve III shows the antenna frequency hopping in the frequency domain.Come measurement performance according to throughput, described throughput is defined as the correct packet count that receives divided by the sum that sends grouping.In all 4 kinds of situations, all considered automatic repeat requests (ARQ).As can be known, curve IV shows optimum performance from 4 figures.

Any one the disclosed embodiments can with one or several combination among shown and/or described other embodiment.One or more combination of features for embodiment also are possible.

The present invention can realize with the combination of hardware, software or hardware and software.The computer system of any type or other device that is suitable for carrying out herein the method for describing all are suitable for.The typical combination of hardware and software can be the general-purpose computing system with computer program, and this computer program is controlled computer system when being loaded into and carry out, thereby carries out the method for describing herein.The present invention also can be embedded in the computer program, and this product comprises all features that the feasible method of describing can realize herein, and when this product was loaded into computer system, it can carry out these methods.

Computer program instrument herein or calculation procedure mean any expression of one group of instruction being carried out with any language, code or note, described instruction is intended to make after system's any or all directly or in carrying out following operation with information processing capability, carries out specific function: a) convert another kind of language, code or note to; B) reproduce with different material form.

Claims (10)

Hide Dependent

1. method that the multicarrier that is used for data sends comprises step:
Data flow (b) is provided;
This data flow of encoding is to create a plurality of complex values (x);
With a plurality of complex values (x_k) each distribute in a plurality of subchannels (k) of one that form in two or more channels (l) one;
To be worth (a separately $_{L, k}$) distribute to each in these a plurality of subchannels (k), wherein provide to be used for this value (a separately $_{L, k}$) stochastic variable, each described independent value is a plural number, described stochastic variable is used to described independent value that in random phase value or the random magnitude value one is provided;
Should a plurality of complex values (x_k) each and the independent value (a that distributed $_{L, k}$) multiply by each product value (m of these a plurality of subchannels (k) of generation mutually $_{L, k}$);
Use inverse fast fourier transformed (IFFT), the product value (m of each that will these a plurality of subchannels (k) $_{L, k}$) be modulated to different subcarriers, to produce each modulation signal (s of these two or more channels (l) $_1$); With
Send each this modulation signal (s of these two or more channels (l) simultaneously $_1$).

2. the method for claim 1 is wherein with the independent value (a that is distributed $_{L, k}$) step that multiplies each other provides phase shift and/or the changes in amplitude in the subcarrier.

3. method as claimed in claim 2, wherein the difference from a subcarrier to the phase shift of next subcarrier is a constant.

4. the method for claim 1, wherein with the interval [0,2 π) in variable as value (a separately $_{L, k}$) in stochastic variable.

5. the method for claim 1 wherein will be worth (a separately $_{L, k}$) each the step of distributing to a plurality of subchannels (k) comprises: be provided for independent value (a $_{L, k}$) the constant amplitude with out of phase value.

6. the method for claim 1 also comprises: during in knowing a plurality of subchannels (k) one channel gain, change value (a separately $_{L, k}$) phase

value ($\varphi_{L,k}$), thereby be worth (a separately $_{L,k}$) provide with a plurality of subchannels (k) in the corresponding phase shift of phase reversal of one phase place.

7. method as claimed in claim 6 also comprises: adapt to value (a separately $_{L,k}$) range value (α), thereby one amplitude in range value (α) and a plurality of subchannels (k) is proportional.

8. the method for claim 1, wherein modulation step comprises the OFDM modulation.

9. method as claimed in claim 1 or 2, wherein data flow (b) comprises an independent value (a who divides into groups and be used for each grouping $_{L,k}$).

10. device (2) that the multicarrier that is used for data sends comprising:
Cell encoder (10) is used for receiving data stream (b) and creates a plurality of complex values (x);
Remove multiplexer (14), be used for a plurality of complex values (x_k) each distribute in a plurality of subchannels (k) of one that form in two or more channels (1) one;
Multiplication unit (16) is used for a plurality of complex values (x_k) each and independent value ($a_{L,k}$) multiply by each the product value (m that produces a plurality of subchannels (k) mutually $_{L,k}$), wherein provide to be used for this value (a separately $_{L,k}$) stochastic variable, each described independent value is a plural number, described stochastic variable is used to described independent value that in random phase value or the random magnitude value one is provided;
Modulator (20) is used to use inverse fast fourier transformed (IFFT), the product value (m of each that will these a plurality of subchannels (k) $_{L,k}$) be modulated to different subcarriers, to produce each modulation signal (s of these two or more channels (l) $_1$); With
Transmitter (30) is used for through transmit antenna (A_1) launch modulation signal (s simultaneously $_1$), each of two or more channels (1) has the transmit antenna (A that distributes to it $_1$).

Patent Citations (7)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
EP0660559B1 *1993-12-222005-04-27Philips Electronics N.V.Multicarrier frequency-hopping communication system
CA2214934C *1996-09-242001-10-30At&T Corp.Method and apparatus for mobile data communication
US6005876A *1996-03-081999-12-21At&T CorpMethod and apparatus for mobile data communication
JPH1051418A *1996-08-061998-02-20Mitsubishi Electric CorpDigital receiver
US5973642A *1998-04-011999-10-26At&T Corp.Adaptive antenna arrays for orthogonal frequency division multiplexing systems with co-channel interference
EP1148659B1 *2000-04-182005-07-27Sony International (Europe) GmbHOFDM diversity transmission
EP1267513A3 *2001-06-112006-07-26Unique Broadband Systems, Inc.Multiplexing of multicarrier signals
* Cited by examiner, † Cited by third party

Cited By (5)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
JP2006196978A *2005-01-112006-07-27Kddi CorpBeam control apparatus, array antenna system, and wireless device
US8760994B22005-10-282014-06-24Qualcomm IncorporatedUnitary precoding based on randomized FFT matrices
US8862394B2 *2013-01-222014-10-14Ensco, Inc.System and method for tracking and locating a person, animal, or machine
EP2835926B1 *2013-08-052019-06-12Alcatel LucentTransmitter apparatus for conditioning a multicarrier signal, network node, method for conditioning a multicarrier signal, and computer program thereof
CN107356921A *2017-08-112017-11-17桂林电子科技大学A kind of method that frequency diversity array radar is positioned based on a frequency deviation target
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
US9973365B22018-05-15Method and system for combining DFT-transformed OFDM and non-transformed OFDM
US10439784B22019-10-08Systems and methods for use with orthogonal frequency division multiplexing
US20170195097A12017-07-06Methods and Systems for Orthogonal Frequency Division Multiplexing (OFDM) Multiple Zone Partitioning
US10356805B22019-07-16Methods and systems for scheduling in a virtual MIMO communication environment
US9871617B22018-01-16Method of optimizing portions of a frame
US9572172B22017-02-14Adaptive scheduling of voice traffic in a multi-carrier communication environment
US9225575B22015-12-29Precoding for single transmission streams in multiple antenna systems
US9191092B22015-11-17Method and system for compromise greenfield preambles for 802.11n
US8588276B22013-11-19Methods and systems for OFDM using code division multiplexing
US20170070325A12017-03-09Method and apparatus for generating pilot tone in orthogonal frequency division multiplexing access system, and method and apparatus for estimating channel using it
US9100076B22015-08-04Time varying delay diversity of OFDM
US8774151B22014-07-08Closed-loop MIMO systems and methods
JP6088574B22017-03-01Physical broadcast channel (PBCH) transmission for reliable antenna configuration detection
EP3005637B12019-07-24Methods and devices for processing a data frame having variable guard interval
US8532082B22013-09-10System and method for frequency division multiple access communications
US8363696B22013-01-29Low-rate long-range mode for OFDM wireless LAN
US8406114B22013-03-26Base station and method for transmitting downlink reference signals for MIMO channel estimation
CN101675620B2013-07-24Method and apparatus for packet wireless telecommunications
US8526529B22013-09-03Radio access method for reduced PAPR
CN102143119B2016-03-16The transmission method of OFDM multi-channel multiplexing symbols and system

CN101103552B2013-05-15Methods and apparatus for mitigating multi-antenna correlation effect in communication systems
RU2289210C22006-12-10Device and method for transferring/receiving data in communication system, using multi-access layout
CN102523192B2014-09-17Method and apparatus for transmitting uplink acknowledgement information in an OFDMA communication system
US7388847B22008-06-17Channel quality indicator for OFDM
US6842487B12005-01-11Cyclic delay diversity for mitigating intersymbol interference in OFDM systems

Priority And Related Applications

Priority Applications (2)

ApplicationPriority dateFiling dateTitle
EP018112322001-12-17
EP01811232.62001-12-17

Legal Events

DateCodeTitleDescription
2005-04-27C06Publication
2005-04-27PB01Publication
2005-06-29C10Entry into substantive examination
2005-06-29SE01Entry into force of request for substantive examination
2009-03-18C14Grant of patent or utility model
2009-03-18GR01Patent grant
2011-02-02C17Cessation of patent right
2011-02-02CF01Termination of patent right due to non-payment of annual fee
Granted publication date: 20090318

Termination date: 20091221

[19] 中华人民共和国国家知识产权局

[51] Int. Cl.
H04L 27/26 (2006.01)

[12] 发明专利说明书

专利号 ZL 02825121.0

[45] 授权公告日 2009 年 3 月 18 日

[11] 授权公告号 CN 100471191C

[22] 申请日 2002.11.21 [21] 申请号 02825121.0

[30] 优先权

[32] 2001.12.17 [33] EP [31] 01811232.6

[86] 国际申请 PCT/IB2002/004843 2002.11.21

[87] 国际公布 WO2003/053020 英 2003.6.26

[85] 进入国家阶段日期 2004.6.15

[73] 专利权人 国际商业机器公司

地址 美国纽约州

[72] 发明人 安德列·诺尔巴雷托

[56] 参考文献

EP0831627A2 1998.3.25

EP1148659A1 2001.10.24

CN1318920A 2001.10.24

审查员 赵晶晶

[74] 专利代理机构 北京市柳沈律师事务所

代理人 黄小临 王志森

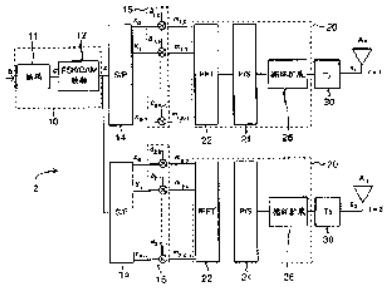
权利要求书 2 页 说明书 8 页 附图 3 页

[54] 发明名称

用于多载波发送的方法和装置

[57] 摘要

本发明提供了一种用于数据的多载波发送的方法和装置。该方法包括步骤：提供数据流；编码该数据流以创建多个复数值；将多个复数值的每一个分配给多个子信道中的一个；将单独值分配给该多个子信道的每一个；将该多个子信道的每一个与所分配的单独值相乘以产生该多个子信道的每一个的乘积值；将该多个子信道的每一个的乘积值调制到子载波，以产生该子信道的每一个的调制信号；和同时发送该调制信号。



1. 一种用于数据的多载波发送的方法，包括步骤：

提供数据流(b)；

编码该数据流以创建多个复数值(x)；

将多个复数值(x_k)的每一个分配给形成两个或多个信道(l)中的一个的多个子信道(k)中的一个；

将单独值($a_{l,k}$)分配给该多个子信道(k)中的每一个，其中提供了用于该单独值($a_{l,k}$)的随机变量，每个所述单独值为复数，所述随机变量用于为所述单独值提供随机相位值或随机幅度值中的一个；

将该多个复数值(x_k)的每一个与所分配的单独值($a_{l,k}$)相乘以产生该多个子信道(k)的每一个的乘积值($m_{l,k}$)；

使用反向快速傅立叶变换(IFFT)，将该多个子信道(k)的每一个的乘积值($m_{l,k}$)调制到不同的子载波，以产生该两个或多个信道(l)的每一个的调制信号(s_l)；和

同时发送该两个或多个信道(l)的每一个的该调制信号(s_l)。

2. 如权利要求 1 所述的方法，其中与所分配的单独值($a_{l,k}$)相乘的步骤提供了子载波中的相移和/或幅度变化。

3. 如权利要求 2 所述的方法，其中从一个子载波到下一个子载波的相移之差是常数。

4. 如权利要求 1 所述的方法，其中将区间$[0,2\pi)$中的变量用作该单独值($a_{l,k}$)中的随机变量。

5. 如权利要求 1 所述的方法，其中将单独值($a_{l,k}$)分配给多个子信道(k)的每一个的步骤包括：提供用于单独值($a_{l,k}$)的具有不同相位值的常数幅度值。

6. 如权利要求 1 所述的方法，还包括：当知道多个子信道(k)中的一个的信道增益时，改变单独值($a_{l,k}$)的相位值($\phi_{l,k}$)，从而单独值($a_{l,k}$)提供与多个子信道(k)中的一个的相位的逆相位相对应的相移。

7. 如权利要求 6 所述的方法，还包括：适应单独值($a_{l,k}$)的幅度值(α)，从而幅度值(α)与多个子信道(k)中的一个的幅度成比例。

8. 如权利要求 1 所述的方法，其中调制步骤包括 OFDM 调制。

9. 如权利要求 1 或 2 所述的方法，其中数据流(b)包括分组和用于每个

分组的一个单独值($a_{l,k}$)。

10. 一种用于数据的多载波发送的装置(2)，包括：

编码器单元(10)，用于接收数据流(b)和创建多个复数值(x)；

去多路复用器(14)，用于将多个复数值(x_k)的每一个分配给形成两个或多个信道(l)中的一个的多个子信道(k)中的一个；

乘法单元(16)，用于将多个复数值(x_k)的每一个与单独值($a_{l,k}$)相乘以产生多个子信道(k)的每一个的乘积值($m_{l,k}$)，其中提供了用于该单独值($a_{l,k}$)的随机变量，每个所述单独值为复数，所述随机变量用于为所述单独值提供随机相位值或随机幅度值中的一个；

调制器(20)，用于使用反向快速傅立叶变换(IFFT)，将该多个子信道(k)的每一个的乘积值($m_{l,k}$)调制到不同的子载波，以产生该两个或多个信道(l)的每一个的调制信号(s_l)；和

发射机(30)，用于经传输天线(A_l)同时发射调制信号(s_l)，两个或多个信道(l)的每一个具有分配给它的传输天线(A_l)。

用于多载波发送的方法和装置

技术领域

本发明涉及一种用于数据的多载波发送的方法和装置。特别是，本发明涉及一种尤其适合于无线传输的有效传输分集方案。

背景技术

已经提出了用于无线环境的多载波调制，其不但用于广播应用，如在European Digital Video Broadcasting（DVB，欧洲数字视频广播）标准中定义的，而且用于高速率无线局域网（W-LAN），如在北美 IEEE 802.11a 和在欧洲 HIPERLAN-2 标准中定义的，上述应用都依赖于编码的正交频分复用（OFDM）。这些标准支持高达 54Mbps 的高数据率无线传输。

隐藏在 OFDM 之后的思想是将到来的数据流分割成几个较低速率的并行数据流（因此符号周期 T_s 更长）并在不同的子信道中发送每个数据流。使用相隔 $1/T_s$ 的不同子载波发送这些数据流。通过这样选择子载波间隔，子信道在被适当地抽样时是正交的，并且允许子信道的频谱重叠，从而最大化传输的频谱效率。

OFDM 的优点在于其对符号间干扰（ISI）的适应力，该 ISI 是由无线信道中常有的多径传播造成的。该适应力可通过使用保护间隔来循环扩展信号来实现，该保护间隔应当长于信道的最大延迟。

通常，频率选择衰落（即在不同的频率上观测到不同的衰落）表现了宽带无线系统的特点。在编码的 OFDM 中，编码跨不同的子载波上的数据比特，这给抵抗频率选择信道提供了某些保护。但是，由于相邻频率极可能是高度相关的，因此该保护也是受限的，从而深度衰落往往影响几个子信道。

抵抗衰落的一个可选方法是使用多个天线以获得空间分集。为了获得足够的分集，不同天线的信道必须具有低相关性，这意味着这些信道彼此相隔地足够远。除此之外，每个天线还需要单独的无线前端，从而增加了收发信机的成本。这些问题使得更可能仅在基站使用多个天线，因此，在下行链路

中，必须在发射机侧使用分集技术。

高速 W-LAN 系统定位于室内环境中的静止应用或低速移动应用。对于这类应用，信道变化地很慢，例如，以步行速度(1m/s)，对于载波频率 f_c=5GHz，相干时间为 T_c=25ms，这对应于 HIPERLAN/2 中 12 个 2ms 的 MAC 帧。对于静止（便携）终端，衰落可能持续几百毫秒。对于数据应用，自动重传请求（ARQ）方案或简单的分组重传可用来保证低分组丢失和近乎无错传输。但是在上述信道条件下，直到该分组被无错接收时，分组可能不得不被多次重传或者在重传之间有大的延迟，从而降低了系统吞吐量且增加了传输延迟。

在美国专利第 5,914,933 号中建议了一种所谓的分集的 OFDM 系统 (clustered OFDM)，其中，连续子载波的不同子集被分配给每个天线。该系统的缺点在于：由于相邻的子载波是从同一天线发送的，并因此而相关，所以能够获得小的频率分集。

美国专利第 6,005,876 号描述了高速无线传输系统，其中，子集是这样的，以至子载波被均匀地扩展在整个带宽上。这可以被看作是频域中的天线跳频。考虑到使用重复机制的吞吐量，该系统具有缺点。就频率分集而言，该方法代表了一种进步，但是考虑到使用 ARQ 的时间分集，即使变化子载波，获益也很少。

从上述清楚地可知：非常需要一种可应用于现有标准，诸如基于 OFDM 的标准，的有效传输分集方案。而且，为了提高传输的性能和获得更高的可靠性，降低误码率并从而提高数据的吞吐量应该是可以实现的。

发明内容

根据本发明的一个方面，提供了一种用于数据的多载波发送的方法，该方法包括步骤：

提供数据流；

编码该数据流以创建多个复数值；

将多个复数值中的每一个分配给形成两个或多个信道中的一个的多个子信道中的一个；

将单独值(separate value)分配给该多个子信道中的每一个；

将该多个子信道的每一个与所分配的单独值相乘以产生该多个子信道的每一个的乘积值(multiplied value)；

将该多个子信道的每一个的乘积值调制到子载波，以产生该两个或多个信道的每一个的调制信号；和

同时发送该两个或多个信道的每一个的该调制信号。

该方法提供了一种有效的传输分集方案，只要使用多个天线，该方案就具有低的额外复杂度，从而其可以应用于诸如基于 OFDM 的 W-LAN 标准的现有标准，而不需或仅需极少地修改就可应用于现有标准。而且，可以实现误码率的实质性的降低。因此，较高的吞吐量是可以实现的。因此可以提供传输性能的提高和更高的可靠性。

基本上，该方法提供了频域预失真且使用了多个发射天线以提高多载波系统的频率分集。该方法也可以用来提供一种具有时间分集的系统，可以通过高层的错误控制功能（例如，自动重传请求(ARQ)）来利用该具有时间分集的系统，以提高数据吞吐量。

与所分配的单独值相乘的步骤提供了子载波中的相移和/或幅度变化。通过这样做，频域中的自相关变得更小。此外，可以更有效地使用所应用的码。

如果从一个子载波到下一个子载波的相移之差是常数，这是有利的。这影响信道中的延迟。从而在接收机侧，可以更有效地执行信道估计。

将单独值分配给多个子信道的每一个的步骤可以包括：提供用于单独值的随机变量。使用随机变量提高了信道中的频率选择性，同时，所使用的码变得更有效。

将单独值分配给多个子信道的每一个的步骤可以包括：提供用于单独值的具有不同相位值的常数幅度值。由于保持了子载波之间的功率分配，而不会显著影响传输性能，所以这是有利的。

不同的相位值可能属于一组可能的固定值，这是由于其后可以简化复数乘法。

数据流包括分组和用于每个分组的一个单独值，即，对于每个分组，单独值是不同的。通过这样做，可以实现将单独值固定分配给各个分组，这导致了时间分集。

当知道多个子信道中的一个的信道增益时，改变单独值的相位值，从而单独值提供与多个子信道中的一个的相位的逆相位相对应的相移，是有利的，这是由于其后如下优点出现：来自不同天线的信号是可相干接收的。

当信道增益是已知的，即，信道估计是成功的时，修改单独值的幅度值，从而幅度值与多个子信道中的一个的幅度成比例，是更加有利的，这是由于

其后如下优点出现：信号是可相干接收的，并且信噪比（SNR）可被最大化。

调制步骤可以包括OFDM调制。这示出了：所提议的方案可被应用于标准调制技术。

根据本发明的第二方面，提供了一种用于数据的多载波发送的装置，该装置包括：

编码器单元，用于接收数据流和创建多个复数值；

去多路复用器，用于将多个复数值的每一个分配给形成两个或多个信道中的一个的多个子信道中的一个；

乘法单元，用于将多个子信道的每一个与单独值相乘以产生多个子信道的每一个的乘积值；

调制器，用于将多个子信道的每一个的乘积值调制到子载波，以产生两个或多个信道的每一个的调制信号；和

发射机，用于经传输天线同时发射调制信号，两个或多个信道的每一个具有分配给它的传输天线。

因此，本发明的这个方面的实施例使用与上述相类似的原理。

附图说明

通过下面参照示例性的附图，对本发明的优选实施例进行了详细的、仅作为示例的描述，其中：

图1示出了对根据本发明的多载波发送装置的示例性说明；

图2以更抽象的方式示出了对多载波发送装置的示例性说明；

图3示出了相应接收机的示例性说明；

图4示出了显示相对于不同传输方案的数据吞吐量的图。

附图是为了说明的目的提供的，并且附图不必按比例地表示本发明的实际示例。

具体实施方式

虽然本发明可用于多种多载波发送应用，但是重点描述对无线系统，即无线局域网（W-LAN）的应用，其如在W-LAN标准IEEE 802.11a和HIPERLAN-2中使用正交频分复用(OFDM)那样使用OFDM。在描述本发明

的实施例之前，根据本发明说明一些基础知识。

通常，所提出的传输分集方案使用在天线 A_l 上各个子载波上的第 k 个子信道上发送的符号（也被称为复数值 $x_k(i)$）与系数（也被称为单独值 $a_{l,k}$）的乘法。表达式 i 对应于第 i 个 OFDM 符号。每个单独值 $a_{l,k}$ 包括幅度值 $\alpha_{l,k}$ 和相位值 $\phi_{l,k}$，如下面更详细描述的。单独值 $a_{l,k}$ 可以被看作是复数值。使用具有至少两个天线 A_l(这意味着具有至少两个信道 l)的系统可以得到最好的结果。考虑如图 3 所示的单个接收天线 52 的情况，第 k 个子信道的、经过快速傅立叶变换（FFT）之后的接收信号将是:

$$r_k(i) = h_{eq,k} x_k(i)$$

其中 $h_{eq,k}$ 是由所有信道 l 构成的等效信道的增益，也被称为等效信道增益 $h_{eq,k}$。

该增益通过下式给出:

$$h_{eq,k} = \sum_l a_{l,k} h_{l,k}$$

其中，$h_{l,k}$ 是第 l 个天线 A_l 和第 k 个子信道的信道增益。发射天线 A_l 的数量和单独值 $a_{l,k}$ 的选择对接收机是透明的，并且不需要额外的信令。接收机接收使用等效信道增益 $h_{eq,k}$ 修正的发送信号 $x_k(i)$，就像它是从单个天线 A 发送的。因此，接收机仅仅看到等效信道增益 $h_{eq,k}$，并且如果单独值 $a_{l,k}$ 也可应用于训练前导(preamble)，使用如现有技术中已知的传统的信道估计技术，就可以获得等效信道增益 $h_{eq,k}$。

为了提供时间分集，单独值 $a_{l,k}$ 应当在每个分组处改变。有几种不同的方法来选择与第 n 个分组相对应的单独值 $a_{l,k}(n)$。在第一个示例中建议：通过使

$$a_{l,k}(n) = \alpha_{l,k} \exp(j\phi_{l,k}(n)),$$

则所有的单独值都具有幅度 $\alpha_{l,k}$ 和随机相位，其中，相位值 $\phi_{l,k}(n)$ 包括区间 $[0,2\pi)$ 中的独立统一(independent uniform)随机变量。如果需要如在单个天线系统中那样的同一发送功率，可以将幅度选择为 $\alpha_{l,k} = \sqrt{(L)}$，其中 L 是天线 A_l 的总数。可以示出：与单个天线系统相比，该系统的频率分集随着该选择，即不同子信道 k 的信道增益之间的相关性降低而提高。这导致了误码率的实质性降低。或者，在第二示例中，可以随机选择幅度 $\alpha_{l,k}$。根据第一示例的、使用随机相位的方法的性能类似于第二示例。

如上所述，当使用诸如自动重传请求（ARQ）的分组重传方案时，所提议的传输分集方案的时变特性提供了时间分集。该技术在接收机处可以与分组合并一起使用，以实现进一步的性能提高。具有错误的所接收的分组不应

该被丢弃。不是存储有错误的分组，而是将其与同一分组的后来的重传版本组合，理想的情况是使用最大比进行合并。与传输分集方案相关的分组合并可以提高OFDM无线系统的吞吐量。这导致了容量的提高和传输延迟的降低，并且还可以在现有系统中使用。

图 1 示出了多载波发送装置 2 的示例性说明。编码器单元 10 在其输入处接收数据流 b，并在其输出处提供多个复数值 x。编码器单元 10 还被看作是比特交织编码调制(BICM)单元 10，其在这里包括编码器 11 和映射器 12，其或者使用相移键控（PSK）或者使用正交调幅（QAM）。为了简便起见，未示出编码器 11 和映射器 12 之间的交织器单元。编码器单元 10 的输出与 2 个去多路复用器 14 相连，其中，每个去多路复用器对应于一个信道 l。信道 l 的数量可以高于两个，如图 2 所示的。下面，仅将一个信道看作是各单元的功能是相同的。去多路复用器 14 将多个复数值 x_k 的每一个分配给多个子信道 k 的每一个。乘法单元 16 与多个子信道 k 的每一个相连。单独值 $a_{l,k}$ 被提供给乘法单元，并且如上所述，其是可指定的。在每个信道 l 中，多个子信道 k 与调制器 20 相连。调制器 20 包括反向快速傅立叶变换（IFFT）单元 22，其与多路复用器 24 相连。多路复用器 24 串行化从反向快速傅立叶变换（IFFT）单元 22 接收的信号流。串行化的信号流被馈送给循环扩展单元 26。循环扩展单元 26 的输出，也就是调制器 20 的输出，被馈送给发射机 30。这样的发射机 30 通常包括发送或 TX 滤波器和 RF（射频）前端，为了简化而未示出。调制信号 s_l 可经发射天线 A_l 发送。每个信道 l 具有其发射天线 A_1,A_2。

多载波发送装置 2 如下进行操作。由编码器单元 10 将数据流 b 编码成多个复数值 x。将多个复数值 x_k 的每一个分配给多个子信道 k 中的一个。此外，将一个单独值 $a_{l,k}$ 分配给多个子信道 k 的每一个。尽管有几种变化的可能性，都可以如上所述创建每个单独值 $a_{l,k}$。同时，能够使单独值 $a_{l,k}$ 适合信道条件。如图 1 所示，将多个子信道 k 的每一个与所分配的单独值 $a_{l,k}$ 相乘，以产生多个子信道 k 的每一个的乘积值 $m_{l,k}$。其由乘法单元 16 中的乘法符号示出。在调制器 20 中，多个子信道 k 的每一个的乘积值 $m_{l,k}$ 被馈送到反向快速傅立叶变换（IFFT）单元 22。在使用多路复用器 24 进行的串行化和使用循环扩展单元 26 进行的处理之后，将调制信号 s_l 提供给发射机 30。经分配给各个信道 l 的发射天线 A1,A2 同时发射每个信道 l 的调制信号 s_l。

图 2 示出了具有多个信道 l 的多载波发送装置 2 的另一实施例的示例性

说明。采用矢量表示数据，如有下划线的字母所示。通用的结构和功能与图1的相似。相同的参考数字用来表示相同或相似的部件。使用编码器11，将长度为 N_{pack} 的数据流 b(n)（也被称为输入数据序列）编码成码率为 R_c 的 $N_{pack,c}=N_{pack}/R_c$ 个码比特，然后将其分割成每个有 N_c 个比特的 $N_{pack,c}/N_c$ 个块 c(i)，其对应于第 i 个 OFDM 符号。然后，通过使用映射器 12 将其映射成 $K_d=N_c/\log_2(M)$ 个 QAM 或正交相移键控（QPSK）符号，其也被称为复数值矢量 $\underline{x}(i)$，其中 M 是星座大小。为了简化表示法，去掉时间索引 i，同时考虑单个 OFDM 符号或复数值矢量 $\underline{x}$。复数值矢量 $\underline{x}$ 对应于频域中的 OFDM 信号。引入了与各个子信道相关的 K_p 导频和 K_z 零子载波，并且信号经过 K 点反向快速傅立叶变换（IDFT），其中，$K=K_d+K_p+K_z$，如在调制器 20（未示出）中实现的。对于如此获得的时域信号，一个加上 G 个抽样的循环前缀，如在也包括在调制器 20 中的循环扩展单元 26（未示出）中实现的，以便消除高达 $T_G=GT_s$ 延迟扩展的多径干扰，其中 T_s 是抽样间隔。所得的调制信号 s_l 被滤波，使用发射机 30 将其转换成射频，并经发射天线 A_l 经过多径信道发射。

多载波发送装置 2 在频域中使用预失真，该预失真如在多路复用单元 16 中乘每个子信道 k 的符号所示。通过将复数值矢量 $\underline{x}$ 的元素与单独值矢量 $\underline{a_l}$ 的元素相乘来执行该预失真。在第 k 个子载波和第 l 个天线 A_l 上的发送信号是：

$$x_{l,k}=a_{l,k}x_k$$

接收机执行反向操作。接收信号被滤波，被转换到基带并以 $1/T_s$ 的速率被抽样。去除循环扩展且执行离散傅立叶变换（DFT）。去除零和导频子载波，并且该操作之后在第 k 个子信道的信号是：

$$y_k=h_kx_k+v_k$$

其中，h_k 是等效信道增益，而 v_k 是方差为 N_0 的复数噪声分量。基于信道估计 $\hat{h}_k$，可以均衡接收信号以获得信号估计：

$$\hat{x}_k=\frac{y_k}{\hat{h}_k}$$

分别使用符号和信道矢量估计 $\hat{x}$ 和 $\hat{h}$，可以获得码比特 $\hat{c}$ 的对数似然比 (log-likehood ratio)，可以使用例如软输入维特比解码器对其进行解码。

在每个数据分组之前发送已知的前导，以便允许接收机进行同步和信道估计，以及频率偏移的初始获得。也可用单独值 $a_{l,k}$ 来修正前导。由于 OFDM 系统对频率估计误差很敏感，引入多个导频子载波以改善在一个分组内的频

率偏移的估计和校正。IEEE 802.11a 支持可变比特率，其可以通过不同的调制方案和不同的编码率来实现。

在接收机中，每个接收天线的频域信号可以与矢量按元素相乘，并且来自所有接收天线的信号被累加在一起。可以根据合并方案，象例如用于最大化信噪比（SNR）的最大比合并，来选择加权矢量。

图 3 示出了可与图 1 和图 2 中示出的多载波发送装置 2 结合应用的接收机 50 的示例性说明。接收机 50 包括单个接收天线 52、解调器单元 54 和 56、和解码器 58，上述各单元连成一线。解调器单元 54 和 56 使用诸如相干或差分检测的已知技术，来解调接收信号，例如 OFDM 信号。解码器 58 被用作纠错解码器。应该明白：可用多个接收机 50 来接收发送信号 s_l。预失真在理论上对接收机 50 是透明的，接收机 50 不必知道是否使用了传输分集，并且简单地尝试估计等效信道增益 $h_{eq,k}$。

在图 4 中显示了通过使用随机相位的所提议的传输分集方案的性能改善。考虑具有 4 个发射天线的系统，并且并将如曲线 IV 所示的所提议的传输分集方案与如曲线 I 所示的单个天线系统和如曲线 II 和 III 所示的已知传输分集方案相比较。详细地，曲线 II 示出了延迟分集方案，而曲线 III 示出了频域中的天线跳频。按照吞吐量来测量性能，所述吞吐量被定义为正确接收的分组数除以发送分组的总数。在所有 4 种情况中都考虑了自动重传请求（ARQ）。从 4 个图形中可知，曲线 IV 示出了最佳性能。

任意一个所公开的实施例都可以与所示出的和/或所描述的其它实施例中的一个或几个组合。对于实施例的一个或多个特征的组合也是可能。

本发明可以以硬件、软件或硬件和软件的组合来实现。任意类型的计算机系统或适于执行本文中描述的方法的其它装置都是适用的。硬件和软件的典型组合可以是具有计算机程序的通用计算机系统，该计算机程序在被装入和执行时，控制计算机系统，从而执行本文中描述的方法。本发明也可以被嵌入到计算机程序产品中，该产品包括使得本文中描述的方法能够实现的所有特征，并且当该产品被装入计算机系统时，其能够执行这些方法。

本文中的计算机程序工具或计算程序意指以任何语言、代码或注释对一组指令进行的任何表示，所述指令旨在使具有信息处理能力的系统直接或者在执行下列操作中的任一个或全部之后，执行特定功能：a)转换成另一种语言、代码或注释；b)以不同的物质形式再现。

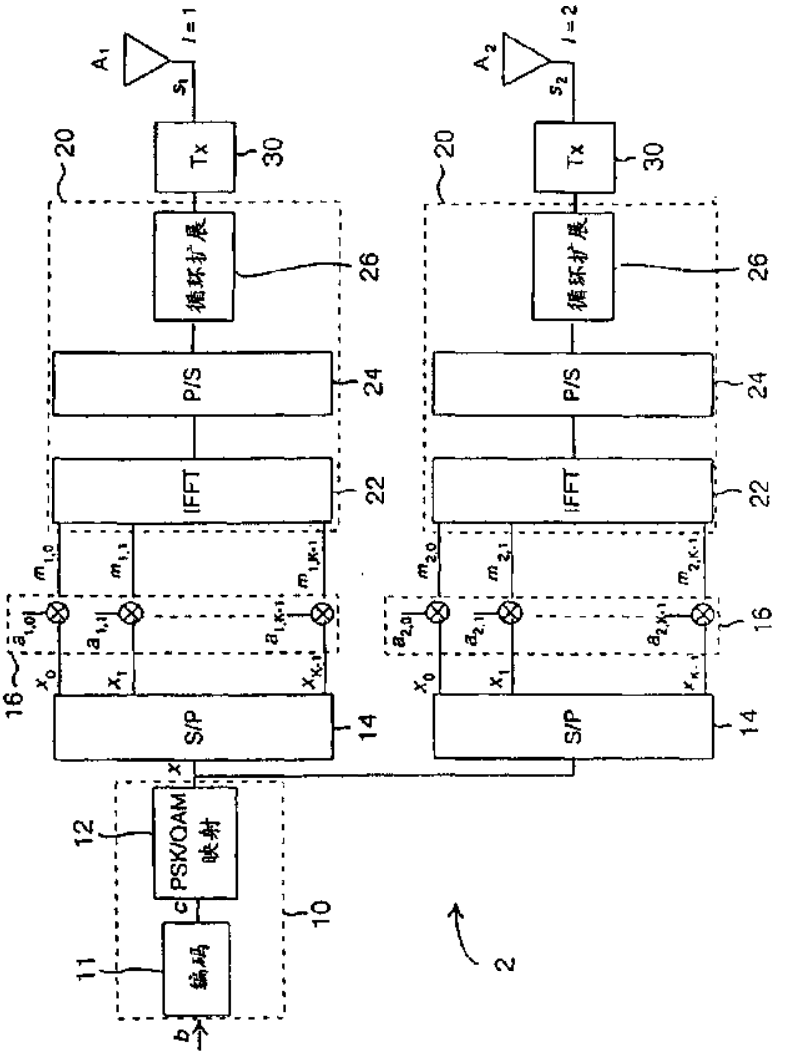

A1
l = 1
s1
A2
l = 2
s2
20
Tx
30
循环扩展
26
P/S
24
IFFT
22
S/P
14
16
PSK/QAM 映射
12
编码
11
10
b
c
x
2

图 1

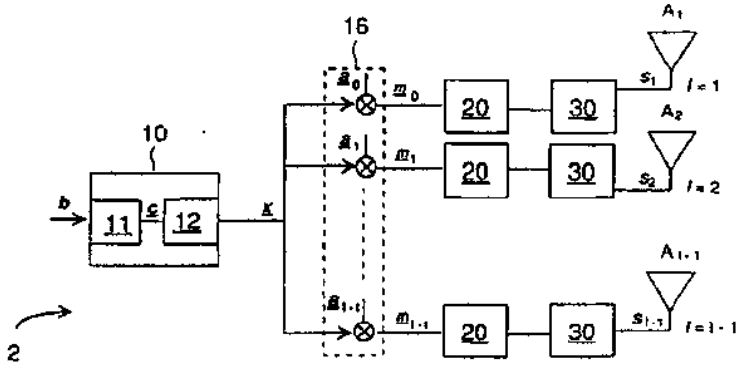


图 2

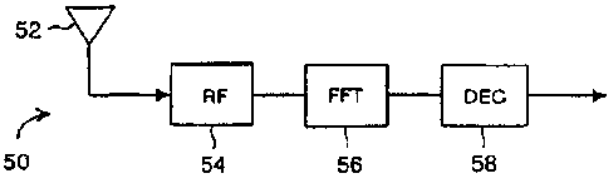


图 3

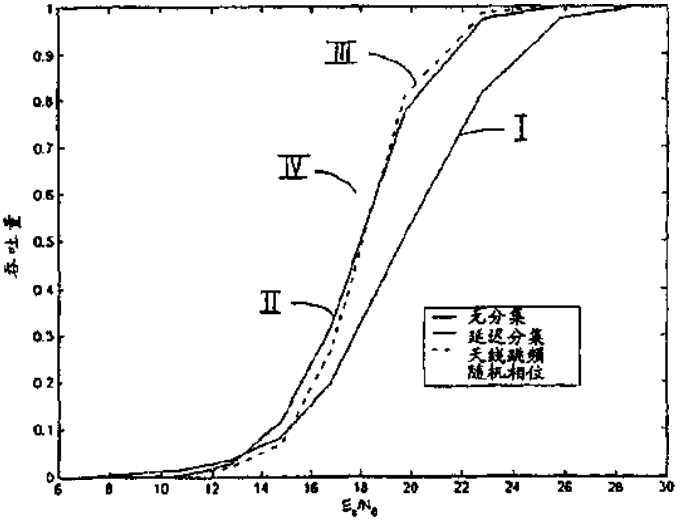


图 4

Correction coding method for ground digital television broadcast

Abstract

This invention relates to earth digital television broadcast coding method in digital information transmission technique field, which comprises the following steps: diving transmission flow into 752 bit set and adding 261 zeros to get 1013 bit for BCH coding and removing front 261 bit to get BCH(762,752); forming one set of BCH(762,752) with certain number to generate matrix of LDPC(7493,3048),LDPC(7493,4572) or LDPC(7493,6096); deleting output 7493 bit front five correction bit to get final 7488 output bit.

CN100584011C
China

Download PDF Find Prior Art Similar

Other languages

Chinese

Inventor

杨林

杨知行

张晓林

门爱东

Worldwide applications
2005 CN

Application CN 200510086336 events

2005-09-02

Application filed by 清华大学

2005-09-02

Priority to CN 200510086336

2007-03-07

Publication of CN1925615A

2010-01-20

Application granted

2010-01-20

Publication of CN100584011C

Info

Cited by (14)

Legal events

Similar documents

Priority and Related Applications
External links
Espacenet
Global Dossier
Discuss

Description

The error correction/encoding method that is used for ground digital television broadcast

Technical field

The invention belongs to digital information transmission technical field, particularly a kind of error correction/encoding method that is used for ground digital multimedia TV broad cast.

Background technology

Experienced the mechanical television epoch, electronics black-and-white TV and color TV be after the epoch, Japan Broadcasting Corporation NHK had many countries and International Standards Organization also to actively develop the research and development of this respect from the initial

stage seventies high definition TV that begins one's study afterwards.But before nineteen ninety, the technical scheme that is adopted also all is a scheme full simulation or that analog digital is mixed.In June 1 nineteen ninety, U.S. GI Company Inc. has submitted digital ground HDTV standard to U.S. advanced television consultative committee (ACATS), and from then on TV has entered a New Times---the Digital Television epoch.

Through insistent research and development for many years, ground digital television broadcast (Digital Television Terrestrial Broadcasting has been formulated in many countries and regions, DTTB) standard mainly contains three international terrestrial DTV standards (U.S. ATSC, European DVB-T and Japanese ISDB-T).At home, Tsing-Hua University has also proposed ground digital multimedia TV broad cast (DMB-T) transmission plan of oneself.These standards and scheme system's formation, design parameter and target etc. many aspect, exist difference more or less, exist various different requirements.

The business need of different ground system of digital television broadcast.In the process of digitization of terrestrial television broadcasts, need provide different business for the user:

√ for example provides multimedia broadcasting and the integrated data service business that comprises HDTV;

√ supports receiving equipments such as fixing, portable, hand-held;

The frequency planning of √ high flexible and broadcast coverage areas.

Different design parameters and environmental requirement.According to the concrete condition of China's terrestrial broadcast television, the ground digital multimedia TV broad cast transmission system face designing requirement have:

√ radio-frequency region: less than 900 megahertzes

√ base band frequency bandwidth: 8 megahertzes

The speed per hour of √ mobile receiver: less than per hour 130 kilometers

√ single transmit machine coverage: radius is less than 75 kilometers

√ area of application: city, suburb, cities and towns

√ is suitable for landforms: Plain, hills, coastal, lakeside

√ environment for use: indoor, outdoor, vehicle-mounted, portable

Can handle different channel model requirements.The channel characteristics of these ground digital multimedia TV broad cast transmission systems has:

The √ frequency selectivity: time delay distribution (Delay spread) is about 50 microseconds

The √ time selectivity: Doppler's distribution (Doppler spread) is about 100 hertz

The √ channel model: broad sense steadily is not dependent scattering (WSSUS)

The √ noise model: different modulation and error correcting system are adopted in additive white Gaussian noise (AWGN), pulsed interference, single-frequency interference etc., adapt to various code rate and require:

√ code check: 5Mb/s～30Mb/s adapts to from moving to fixing different reception conditions;

√ modulation system: QPSK, 16QAM and 64QAM

√ error correcting system: RS sign indicating number, BCH code, convolution code, TCM sign indicating number, TPC sign indicating number, Turbo code, SCCC sign indicating number, LDPC sign indicating number etc.

Require to be consistent with the TS flow structure.Each of MPEG TS stream is surrounded by 188byte and forms, the payload information of the synchronization character of 1byte and 187byte wherein, this just requires the frame structure of terrestrial digital television system preferably to transmit an integer TS code block, and promptly effectively frame length is the integral multiple of 188byte, is convenient to like this and the source encoding interface.

Require to be consistent with OFDM length.In the terrestrial digital television system of multi-carrier modulation, adopted OFDM (OFDM) modulation of different length.Europe DVB-T has used the OFDM modulation of 1705 (2k) or the individual subcarrier of 6817 (8k), and the DMB-T OFDM of Tsing-Hua University has used 3744 effective subcarriers.

Satisfy the requirement of travelling performance.In mobile receiving course, because the influence of multipath reflection and other propagation path, fast slow fading appears in received signal, is a time varying channel.For the ease of handling, need treatment limits in an OFDM frame, make it constant channel when an OFDM frame proximate is.So certain restriction has been proposed the OFDM frame length.

Owing to will satisfy above-mentioned these requirements, and the error rate of ground digital television receiver operate as normal requires 10^{-11}Below, therefore, different ground system of digital television broadcast has adopted different error correction coding modes, different code length and different cascade system etc., so that satisfy different requirements.

1948, Shannon proposes famous channel theory: for the given channel of thanksing for your hospitality, if channel capacity is C, as long as transmitting terminal sends information with the speed R that is lower than channel capacity C, then necessarily there is a kind of coding method, makes the error rate along with the increase of code length becomes arbitrarily small.This is theoretical for realizing that highly reliable transfer of data has provided the direction of research.Infinitely close to shannon limit, become the strong impetus that countless scientific and technical personnel work hard.

Existing three international terrestrial DTV standards (U.S. ATSC, European DVB-T and Japanese ISDB-T) have all adopted by ISN, have interweaved, the basic structure of outer sign indicating number cascade.Their outer sign indicating number has all adopted RS to shorten sign indicating number, and ISN has bigger difference.The ISN of U.S. ATSC error correction coding has adopted the lattice type sign indicating number (TCM) of cbr (constant bit rate), though performance is fine, only adapts to and is used for high code check.Europe DVB-T and Japanese ISDB have adopted the ISN of the convolution code of various code rate as error correction coding, though (64QAM) performance is applicable to all code checks (64QAM, 16QAM,

QPSK) simultaneously not as lattice type sign indicating number during high code check.

Along with the development of error correction coding, low-density error correcting code (Low Density Parity Check, LDPC) focus that is being called research and is using.

Satellite digital video broadcast standard DVB-S (EN 300 421) has been formulated in Europe in 1994, DVB-S standard code QPSK modulation, convolution and Reed-Solomon cascaded channel coding, it is adopted by most satellite television in the world wide and data broadcasting service merchant now.In the end of the year 2004, second generation satellite digital video broadcast standard DVB-S2 has been formulated in Europe again.DVB-S2 system definition realize the functional module of baseband digital signal, (finish, and code check comprises 1/4,1/3,1/2,3/5,2/3,3/4,4/5,5/6,8/9 and 9/10 by LDPC (LowDensity Parity Check) cascade by sign indicating number and ISN outside the BCH for wherein forward error correction (FEC) coding module.Depend on application, the length of FEC coding groups is nldpc=64800bits or 16200bits, and is as shown in table 1, can see that the code length of the BCH code of DVB-S2 and LDPC sign indicating number is all very long from table, and it realizes complicated.When using VCM and ACM, FEC and modulating mode can change between different frame, but remain unchanged in a frame.In the FEC of 8PSK, 16APSK and 32APSK coded-bit, adopted Bit Interleave.The sign map of modulation depends on application, can adopt QPSK, 8PSK, 16APSK and 32APSK constellation mapping.BCH that DVB-S2 adopts and LDPC cascaded code BER～the C/N performance curve as shown in Figure 1.

The BCH of table 1DVB-S2 and the coding parameter of LDPC cascaded code

??LDPC ??code ??BCH?Uncoded ??Block?K_{bch} ??BCH?coded?block?N_{bch}??LDPC?Uncoded?Block?k_{ldpc} ? ?BCH ??t-error?correction ??LDPC?Coded?Block ??n_{ldpc} ??1/4 ??16?008 ??16?200 ??12 ??64?800 ??1/3 ??21?408 ??21?600 ??12 ??64?800 ??2/5 ??25?728 ??25?920 ??12 ??64?800 ??1/2 ??32?208 ??32?400 ??12 ??64?800 ??3/5 ??38?688 ??38?880 ??12 ??64?800 ??2/3 ??43?040 ??43?200 ??10 ??64?800 ??3/4 ??48?408 ??48?600 ??12 ??64?800 ??4/5 ??51?648 ??51?840 ??12 ??64?800 ??5/6 ??53?84

0 ??54?000 ??10 ??64?800 ??8/9 ??57?472 ??57?600 ??8 ??64?800 ??9 /10 ??58?192 ??58?320 ??8 ??64?800

Having set forth terrestrial DTV above has diversified requirement, and in order to adapt to these different requirements, the error correction coding of employing also has nothing in common with each other, and wherein LDPC is the focus of present error correction coding research and application.Before introducing the LDPC sign indicating number, we at first describe several error correction coding notion or the definition relevant with the LDPC sign indicating number, comprise linear block codes (linear blockcode), system linear block code (systematic linear block code), syndrome (or syndrome syndrome), smallest hamming distance, cyclic code (cyclic code) and quasi-cyclic code (Quasi-cyclic code) etc.

The limited territory of element number is called finite field, and perhaps gal is patrolled magnificent territory (Galois Field, GF), the number of element is called the rank in territory in the territory, GF (q) expression q rank finite field.

For (n, k) the binary packet sign indicating number has 2^kIndividual code word.If there is not structural characteristics, then encoder remembers 2 with needs ^{k}Shine upon between individual message sequence and the code word, when the k value is very big, this will be unpractical.But linear characteristic makes encoder more simple.The mould 2 of two code words that and if only and if when remaining a code word, then the binary packet sign indicating number is linear.More formal being defined as follows:

(n, k) block code is the condition of linear block codes its 2 that is that and if only if ^{k}Individual code word constitutes the n dimensional linear SPACE V on the GF (q) $_n$In a k n-dimensional subspace n $V_{N, k}$

Because (n, k) linear block codes C can be expressed as independently code word g of k linearity $_0$, g_{1}...., g_{K-1}Linear combination, therefore, this k linearity independently vector can constitute a k * n matrix, and is as follows:

G = g 0 g 1 M g k - 1 = g 00 g 01 L g 0 , n - 1 g 10 g 11 L g 1 , n - 1 M M M M g k - 1,0 g k - 1,1 L g k - 1 , n - 1

Wherein, gi=(g_{I0}, g_{I1}...., $g_{I, n-1}$), 0≤i＜k.Make u=(u_0, u_1...., u_{K-1}) information encoded of indicating, then corresponding code word is:

$v=u \cdot G=u_0 g_0+u_1 g_1+....+u_{k-1} g_{k-1}$

Wherein matrix G is called the generator matrix of code word C.

Noticeable characteristic of linear block codes is that it can constitute a system configuration (Systematic Structure), the systematicness of so-called linear system block code is exactly that each code word can be divided into absolutely clear two parts, a part is k the original information data without any change, another part is (n-k) individual redundancy check data, and they are the linear combination of information data.

The linear system block code (n, k) can represent with the matrix G on k * n rank fully:

$$G = g_0 g_1 M g_{k-1} = 1\ 0\ L\ 0\ p_{0,0}\ p_{0,1}\ L\ p_{0,n-k-1}\ 0\ 1\ L\ 0\ p_{1,0}\ p_{1,1}\ L\ p_{1,n-k-1}\ M\ M\ M\ M\ M\ M\ p_{i,j}\ M\ 0\ 0\ L\ 1\ p_{k-1,0}\ p_{k-1,1}\ L\ p_{k-1,n-k-1}$$

Wherein, $p_{i,j}=0$ or 1.If make I_kExpression k * k rank unit square formation, then following formula is expressed as:

$G=[I_k ? P]$

Wherein P is the submatrix of G, and it is made up of the back n-k row of G matrix.

For (there are (n-k) * n rank matrix H in n, the k) k of linear block codes C * n rank generator matrix G, and H is independently gone by n-k linearity and forms, so that G and H are quadratures, promptly the row space of any row of H and G is a quadrature.This shows that n rank v is the code word of C, and if only if $vH^T=0$ o'clock.Matrix H is called the check matrix (parity-check matrix) of code word.V_nIn the subspace with C in code vector keep each vector of quadrature to constitute (n, n-k) a linear block codes C_d, C_dThe dual code that is called C.C_dGenerator matrix be the check matrix H of C.

If (n, k) generator matrix of linear block codes is a system form, i.e. $G=[I_k P]$, then its check matrix can be expressed as form:

$$H = P^T I_{n-k} = p_{0,0}\ p_{1,0}\ L\ p_{k-1,0}\ 1\ 0\ L\ 0\ p_{0,1}\ p_{1,1}\ L\ p_{k-1,1}\ 0\ 1\ L\ 0\ M\ M\ M\ M\ M\ M\ M\ M\ p_{0,n-k-1}\ p_{1,n-k-1}\ L\ p_{k-1,n-k-1}\ 0\ 0\ L\ 1$$

It has clearly illustrated that the orthogonality between G and the H, $GH^T=0$.

(n, k) in the codeword set of linear block codes, i information bit numeral is all code words of 0 before selecting, and forms a new subclass, and then (n-i k-i) is (n, k) Ma shortening sign indicating number or shortened code to block code.Because the preceding i of this subclass is for all getting 0, so can not transmit them when transmitting, the k-i position that only passes the back gets final product.This subclass is k-i dimension subgroup in the n-i dimensional linear space, and (n-i, k-i) sign indicating number is shortened in grouping to constitute one.In other words, for (n-i, k-i) sign indicating number is shortened in linear grouping, and can fill i before the information of k-i position is zero information, and the information that makes is that length becomes k, carry out (n then, k) linear block encoding, i information bit before removing at last, the i position of promptly filling previously zero information, (n-i, k-i) linear grouping shortening sign indicating number have just been obtained.This minimum range that shortens sign indicating number is identical with true form at least.The G matrix that shortens sign indicating number just removes i row in the left side in the G battle array of former [n, k] sign indicating number and top i is capable gets final product.Shorten the H matrix of sign indicating number or shortened code, can obtain as long as in true form H matrix, remove some row.The encoder bit rate R that shortens sign indicating number is littler than true form.

If one (n, k) generator matrix of linear block codes is G, check matrix is H, $v=(v_0, v_1\ldots, v_{N-1})$ code word of expression transmission, $r=(r_0, r_1\ldots, r_{N-1})$ hard decision vector that receives of expression demodulator output.Because interchannel noise, r may be different from v.Vector sum

$$e=r+v=(e_0, e_1, \ldots., e_{n-1})$$

Be called n rank error patterns (errorpattern).Among the e 1 is the error of transmission that interchannel noise causes.By following formula as can be known:

$$r=v+e$$

In order to realize error detection, decoder need calculate following (n-k) heavy vector:

$$s=r\cdot H^T=(s_0, s_1, \ldots., s_{n-k-1})$$

S is called the syndrome (or syndrome) that receives vectorial r.S=0 so is when only when r is correct code word.Therefore, if s ≠ 0, r is not a code word just, shows to detect wrong existence.When error pattern e just was a non-zero codeword, though there is mistake among the r, r might remain a code word.This error pattern is called the untestable fault pattern.Codeword detection random error and the ability of proofreading and correct random error depend on the minimum range (minimum distance) of code word.At first we define the Hamming weight (Hamming Weight) of the number of non-zero symbol among the heavy v of binary system n for it, are expressed as w (v).The different number of corresponding position value between two n heavy v, the u is called the Hamming distance between them, with d (v, u) expression.Hamming distance satisfies the triangle inequality, makes that v, u, x are the heavy vectors of three n, then

$d(v, u)+d(u, x)\geq d(v, x)$

Hamming distance between two n heavy v, the u equal they and Hamming distance, promptly d (v, u)=w (v+u).

For given block code C, can calculate the Hamming distance between any two code words, the smallest hamming distance of C is expressed as d_{Min}, it is defined as follows:

$d_{min}=\min\{d(v, u) : v, u\in C, v\neq u\}$

If C is a linear block codes, two code word sums remain a code word among the C, and therefore, the Hamming distance among the C between two code word v, the u just equals the Hamming weight of another one code word (v+u) among the C, promptly

d min = min { d (v , u) : v , u ∈ C , v ≠ u }

= min { w (v + u) : v , u ∈ C , v ≠ u }

= min { w (x) : v , x ∈ C , x ≠ 0 }

= Δ w min

W wherein $_{Min}$It is the minimum weight of linear code C.Therefore, in error correction coding just like next important theorems:

The minimum range of linear block codes equals the minimum weight of non-zero codeword, otherwise also sets up.

Following theorem has been explained the relation between linear block codes minimum range and its check matrix:

Minimum range with linear code C of check matrix H equals among the H it and is the minimal amount of 0 row.

An important subclass of linear block codes is cyclic code (Cyclic Code), and its coding and syndrome calculate and can realize by the shift register of band feedback.Cyclic code has rigorous Algebraic Structure, and its performance is easy to analyze; Cyclic code has cycle characteristics, and coding-decoding circuit is simple, is easy to realize.Most of linear code and the cyclic code found at present have substantial connection, and the major part in them all can ascribe cyclic code to.

Make $v=(v_0, v_1..., v_{N-1})$ be that length is the vector of n, if we are the code element ring shift right i position of v, $0<i<n$ will obtain the heavy v of another $n^{(i)}=(v_{N-i}, v_{N-i+1}...., v_{N-1}, v_0, v_1..., v_{N-i-1})$.Clearly, v ring shift right i position is equal to v ring shift left n-i position.

(n, k) linear code C is a cyclic code, when each cyclic shift of code word remains among the C code word among and if only if the C.Code word v=(the v of cyclic code $_0$, $v_1..., v_{N-1}$) can use polynomial repressentation, its coefficient is the combination of v:

$$v(X)=v_0+v_1X+v_2X^2...+v_{n-1}X^{n-1}$$

The non-zero code multinomial of minimum degree among the cyclic code C (degreen) is unique.If make $g(X)=g_0+g_1X+g_2X^2...+g_{R-1}X^{R-1}+X^r$Be (n, k) the non-zero code multinomial of minimum degree, constant term g so among the cyclic code C_0With perseverance is 1.Simultaneously, the binary system multinomial of n-1 or littler degree is the multinomial of a code word, and and if only if, and it is the product of g (x).(n k) in the cyclic code, has and has only the code polynomial of a n-k degree:

$$g(X)=1+g_1X+g_2X^2+...+g_{n-k-1}X^{n-k-1}+X^{n-k}$$

Because the non-zero code multinomial of minimum degree g (X) is unique, so it is called the generator polynomial of yard C.

The g of g (X) and its ring shift right k position $^{(k)}$(X) exist simple relation between:

$X^{k}g(X)=(X^{n}+1)+g^{(k)}(X)$

Because $g^{(k)}(X)$ be code polynomial, the multiplication of g (X) means that g (X) is X so ^{n}+ 1 factor.If g (x) is the multinomial of n-k degree, and is $X^{n}+1$ the factor, then g (x) just can generate (n, k) cyclic code.

Another kind of important error correcting code is quasi-cyclic code (Quasi-cyclic code, QC code).Each code word of cyclic code C moves to left or moves to right and remains the code word of C after one, and promptly cyclic code has consistency under cyclic shift.But,, do not have this character to some sign indicating number.Though what one of code word cyclic shift obtained is not the code word of this yard, if cyclic shift n_0(≠ 1) position, a code word that remains this yard that obtains.Therefore,

(mn_0, mk_0) linear block codes, if its any one a code word left side or right cyclic shift n_0Behind (≠ 1) position, a code word that is still this yard that obtains claims that then this class sign indicating number is quasi-cyclic code (Quasi-cyclic code, QC code).

(mn_0, mk_0) generator polynomial of quasi-cyclic code C has following form:

G = IP 0 0 P 1 L 0 P m - 1 0 P m - 1 IP 0 L 0 P m - 2 M M M M 0 P 1 0 P 2 L IP 0

Wherein, I and 0 represents k respectively $_0$* k_0The unit square formation and the null matrix on rank, P_iBe k_0* (n_0-k_0) the rank matrix.So each code word of quasi-cyclic code is made up of the m section, every section front position k_0Individual information word, back n_0-k_0Individual is verification unit.Corresponding check matrix is:

H = P 0 T I P T m - 1 0 L P 1 T 0 P 1 T 0 P 0 T I L P 2 T 0 M M M M P T m - 1 0 P T m - 2 0 L P 0 T I

Wherein, I and 0 represents (n respectively $_0$-k_0) * (n_0-k_0) the unit square formation and the null matrix on rank.We can be by being listed as G matrix and the H matrix that exchange obtains quasi-cyclic code systematic code form to the G matrix.

(mn $_0$, mk $_0$) more generally expression-form is as follows for the quasi-cyclic code generator polynomial:

G = G 0 G 1 L G m - 1 G m - 1 G 0 L G m - 2 M M M M G 1 G 2 L G 0

Each G wherein $_i$Be a k $_0$* n $_0$The rank submatrix can be seen, the generator matrix of this form is fully by its preceding k $_0$The row decision.

Circular matrix is a square formation, and each row is the cyclic shift (moving to right) of delegation above it, and first row is the cyclic shift of last column.For this circular matrix, each row is the downward cyclic shifts of its left side one row, and first row are cyclic shifts of last row.The row, column weight of circular matrix equates, i.e. w.Simpler, we are w with regard to the weight that claims circular matrix.If w=1, circular matrix is a permutation matrix so, is called cyclic permutation matrices (circulant permutationmatrix).For circular matrix, row set (reading from top to bottom) are identical with row set (reading from right to left).The characteristic of circular matrix can be embodied by its first row fully, and therefore, first row is called the generator polynomial of circular matrix.

We are expressed as the circular matrix form to the generator matrix G of quasi-cyclic code, and are as follows:

G=[M $_0$M $_1$... M $_{M-1}$],
wherein, M l = G l G l - 1 M G 0 G m - 1 M G l + 1 for 0 ≤ l < m

For 0≤j＜n $_0$, make Q $_j$For extracting M $_0$, M $_1$...., M $_{M-1}$In j row and a mk constituting $_0$* m rank matrix.Therefore, G can be expressed as follows form:

G ′ = [Q 0 Q 1 . . . . . Q n 0 - 1]

For 0≤i＜k $_0$, make Q $_{l, j}$For extracting Q $_j$In (lk $_0$+ i) be listed as and a m * m rank matrix of formation, 0≤l＜m.Therefore, G can be expressed as following circular matrix form:

G ′ ′ = Q 0,0 Q 0,1 L Q 0 , n 0 - 1 Q 1,0 Q 1,1 L Q 1 , n 0 - 1
M M M M Q k 0 - 1,0 Q k 0 - 1,1 L Q k 0 - 1 , n 0 - 1

Similarly, the check matrix of quasi-cyclic code also can be expressed as the circular matrix form.

If the generator matrix of quasi-cyclic code has following form:

G qc = G 0 G 1 M G k 0 - 1 = I O L O Q 0,0 Q 0,1 L Q 0 , n 0 - k 0 - 1 O I L O Q 1,0 Q 1,1 L Q 1 , n 0 - k 0 - 1 M M M M M M M M O O L I Q k 0 - 1,0 Q k 0 - 1,1 L Q k 0 - 1 , n 0 - k 0 - 1 = I k 0 P

Wherein, I is b * b unit matrix, and O is b * b rank zero battle arrays, $Q_{I,j}$Be b * b rank circular matrix, $0 \le i \le k_0-1, 0 \le j \le n_0-k_0-1$.The generator matrix G of this form $_{Qc}$(it is made up of two parts for Systematic Circulant, SC) form, and the left side is the P matrix, and it is corresponding to the redundancy check part of code word, and the right is I to be called systemic circulation $_{K0}$Matrix.Because P is the array of a circular matrix, therefore, we say G_{Qc}It is the SC form.

LDPC (Low Density Parity Check, loe-density parity-check code) sign indicating number is a kind of parity check code based on sparse matrix.The LDPC sign indicating number is the Robert of the Massachusetts Institute of Technology. a kind of good sign indicating number that brother's glug (Robert Gallagher) proposed in the thesis for the doctorate at him in 1963, its performance is limit near Shannon.He has not only proved the premium properties of LDPC sign indicating number: the typical minimum range of (1) these code words increases with the increase of code length is linear; (2) the typical probability of decoding error reduces with the code length index under the BSC channel, and has proposed two kinds of iterative algorithms.But, do not paid close attention at that time by the people because its realization, particularly hardware realize complicated (needing a large amount of multiplication and division computings).This sign indicating number was known as classic sign indicating number type afterwards.

In a very long time, be not subject to people's attention subsequently, live down already.After Frenchman Berrou in 1993 etc. have proposed the Turbo iterative decoding, nineteen ninety-five Mackay finds that with Neal the performance of LDPC sign indicating number can be to compare with Turbo code, and more have superiority in realization, so introduced LDPC again, this silence code word for many years become the error correction coding mode that a kind of introducing is attracted attention very soon, its design, structure, decoding, performance evaluation and application etc. become the focus of people research.1998, MacKay and Spielman invented irregular LDPC.Richardson and Urbanke have started the method for analyzing the design code type with decoding.

The LDPC sign indicating number is very long linear block codes, its check matrix $H_{(n-k)*n}$Be a sparse matrix, each code word satisfies the linear restriction of some, and the number of constraint is normally very little, is easy to decoding.

The information u={u that the LDPC sign indicating number will send $_1$, u_2...., u_mConvert the code word v={v that is transmitted to $_1$, v_2...., v_n}=u G, n＞m, n represent the length of dividing into groups, the span of n usually from thousands of to hundreds of thousands.Corresponding with generator matrix G is a check matrix H, and H satisfies Hv^T=0, H be one almost all by 0 sparse matrix of forming, in every row and the every row 1 number all seldom, for example 3,4 and 5 etc.

Gallagher definition (q) the LDPC sign indicating number is that code length is the code word of n for n, p, and in its check matrix H, 1 number (code weight) is fixed in each row and column, and wherein the number of each row 1 (weight) all is γ, and the number (weight) of row all is ρ, γ 〉 =3.1 overlapping number is smaller or equal to 1 between any two row (or two row), and this is called ranks constraint (Row-ColumnConstraint, RC constraint).If each row of check matrix H be linearity independently, code check is that (ρ-γ)/ρ, otherwise code check is (ρ-γ ')/ρ, wherein γ ' is a line linearity number independently in the check matrix H so.

Fig. 2 is the check matrix by (20,3,4) LDPC sign indicating number of Gallagher structure, its d_{Min}=6, the design code check is 1/4, and actual bit rate is 7/20.

1 the identical LDPC sign indicating number of number (Hamming weight) is called regular LDPC sign indicating number (Regular LDPC code) in the every row of this check matrix and the every row, obtains as shown in Figure 3 two-dimensional plot (BipartiteGraph) by the check matrix H of regular LDPC sign indicating number.Each node x below figure $_n$That represent is variable node (Variable Node), each node z of top $_m$That represent is check-node (Check Node).Certain row x_nZ with non-zero place in these row $_m$Link to each other, for example for x_2Row, three 1 correspond respectively to z in these row $_1$, z_7And z_{12}OK, so just x_2And z_1, z_7And z_{12}Couple together.Consider from the angle of row, certain z of delegation $_m$The x at middle non-zero points place $_n$Link to each other, obtain same two-dimensional plot.In regular LDPC sign indicating number, the number on the limit that links to each other with each information node is identical, and

check-node also has identical characteristics.At decoding end, with some check-node z_mThe x that links to each other $_n$Summation, if result 0, then inerrancy takes place.

The iteration probabilistic decoding algorithm that Gallagher adopts regular LDPC sign indicating number decoding is called reliability and propagates (BeliefPropagation).

Corresponding with regular LDPC sign indicating number is irregular LDPC codes (Irregular LDPC code), 1 number (weight) difference in every row or column in its check matrix H, and for example 3,4 and 5,1 number is also different in the row.Its coding method and regular LDPC sign indicating number are basic identical, in the non-regular two-dimensional plot between the information node, the degree between the check-node is by may be different.Therefore, for the LDPC sign indicating number of non-rule schema structure, the column weight amount of its check matrix H is inequality, is the value of a variation, and this is the important difference between non-regular code and the regular code.

The performance of non-regular code is better than regular code.Studies show that of recent years, for the non-regular code at GF (8) structure, its performance is not worse than Turbo code, can significantly improve codeword performance, and its performance is in close proximity to the Shannon limit.The decoding of non-rule can be adopted credible propagation iterative decoding algorithm, also can adopt sequential decoding and parallel decoding algorithm etc.

Quasi-cyclic low-density check code (Quasi-cyclic LDPC Code, QC-LDPC) generator matrix is sparse cyclic permutation matrices, QC-LDPC compares with the LDPC of other type, coding is realized simple, adopt simple feedback shift register to get final product, the number of its complexity and check bit is linear scale.About the QC-LDPC sign indicating number, please refer to

(1)Z.W.Li, L.Chen, S.Lin, W.Fong, and?P.S.Yeh, “Efficient?Encoding?of?Quasi-Cyclic?LDPCCodes, ”submitted?to?IEEE?Trans.On?Communications, 2003。

QC-LDPC sign indicating number generator matrix G_{Qc}With check matrix be sparse cyclic permutation matrices, wherein a check matrix H_{Qc}Be a γ * ρ

array, it constitutes matrix $B_{i,j}$Be (q-1) * (q-1) rank cyclic permutation matrices, for:

H qc (γ , ρ) = B 1,1 B 1,2 L B 1 , ρ B 2,1 B 1,2 L B 2 , ρ M M M M B γ , 1 B γ , 2 L B γ , ρ

It satisfies the RC constraint.(γ, kernel ρ) have provided QC-LDPC sign indicating number C to H_{Qc}, length is n=(q-1) ρ.

People have proposed the method for many structure cyclic permutation matrices.In the method in early days, mainly based on the basis territory, the cyclic permutation matrices of generation may satisfy or not satisfy the RS constraint, and these class methods have:

(2)R.J.G.Smith, “Easily?Decodable?Efficient?Self-Orthogonal?Block?Codes, ”ElectronicsLetters, vol.13, no.7, pp.173-174, March?1977。

(3)M.Blaum, P.Farrell, and?H.van?Tilborg, “Array?Codes, ”Handbook?of?Coding?Theory, V.S.Pless?and?W.C.Huffman?Eds., Elsevier?1998。

(4)R.M.Tanner, “Minimum-Distance?Bounds?by?Graph?Analysis”IEEE?Trans.on?InformTheory, vol?47, no.2, pp.808-821, February, 2001。

(5)J.-L.Kim, U.Peled, I.Perepelitsa, and?V.Pless, “Explicit?Construction?of?families?ofLDPC?Codes?with?Girth?at?Least?Six, ”the?40th?Annual?Allerton?Conference?on?Communications, Control?and?Computing, Oct.2002。

(6)R.M.Tanner, D.Sridhara, A.Sridharan, T.E.Fuja, and?D.J.Costello, jr, “LDPC?Blockand?Convolutional?Codes?based?on?Circulant?Matrices, ”submitted?to?IEEE?Trans.on?Inform.Theory, 2004。

(7)B.Vasic?and?O.Milenkovic, “Combinatorial?Construction?of?Low?Density?Parity?CheckCodes?for?Iterative?Decoding, ”IEEE?Trans.On?Inform.Theory, vol?50, no.6, pp.1156-1176, June2004。

Said method is basic identical, and the matrix of its structure is also similar.People had proposed the cyclic permutation matrices building method based on matrix decomposition again afterwards, and the list of references of these class methods has:

(8)L.Chen, J.Xu, I.Djurdjevic, and?S.Lin, “Near?Shannon?Limit?Quasi-Cyclic?LDPC?Codes, ”IEEE?Trans.on?Commun.Theory, vol.52, no.7, July?2004(also?in?Proc.2003?IEEE?GlobeCom., pp.2030-2035, San?Francisco, December?1-5, 2003)。

(9)B.Ammar, B.Honary, Y.Kou, J.Xu, and?S.Lin, “Construction?of?Low?Density?Parity?CheckCodes?based?on?Balanced?Incomplete?Block?Designs, ”IEEE?Trans.Inform.Theory, vol.50, no.6, pp.1257-1268, June, 2004。

In the said method, H decomposes by row and column circular matrix, the circulation submatrix on the γ that obtains successively decreasing * ρ rank:

H = B 1 (1) B 2 (1) L B ρ (1) B 1 (2) B 2 (2) L B ρ (2) M M M M B 1 (γ) B 2 (γ) L B ρ (γ)

If $B_i^{(k)}$Weight be 1, then H is a cyclic permutation matrices.Building method based on matrix decomposition has Euclidean geometry method and perspective geometry method, and the RS sign indicating number method with two information symbols, and wherein the list of references of RS sign indicating number method is:

(10)Ivana?Djurdjevic, Jun?Xu, Khaled?Abdel-Ghaffa?and?Shu?Lin, “A?Class?of?Low-DensityParity-Check?Codes?Constructed?Based?on?Reed-Solomon?Codes?With?Two?Information?Symbols”, IEEE?Comm.Letters , vol.7, no.7, pp317-318, July?2003。

The RS sign indicating number is to propose in the Reed of MIT Lincoln Lab and the paper that Solomon delivered in nineteen sixty " Polynomial Codesover Certain Finite Fields ".The RS sign indicating number is the very high block code of a kind of efficient, both has been applicable to and has entangled random error, also is specially adapted to entangle error burst.The RS sign indicating number is a class nonbinary BCH code, and each symbol

is made up of the m bit.After removing some information code element of RS sign indicating number, block length shortens, as long as the supervise code element number is constant, the minimum range of sign indicating number just can not reduce, i.e. the RS sign indicating number of any shortening is still a maximum code.Any k position in the code word of RS sign indicating number all can be used as ensemble of communication.Be on any one GF (q) (n, k) RS sign indicating number, to any k character position, with have only one with the relative code word in this k position.This is convenient to obtain the definite power distribution of any RS sign indicating number.The accurate error-correcting performance of RS sign indicating number is to be distributed by the minimum range of sign indicating number and power to determine, two indexs of this of RS sign indicating number are all fully by n and k decision.This is convenient to the sign indicating number according to index Design RS very much, is used widely.

(n, k) in the RS sign indicating number, input information is divided into one group of km bit, and every group comprises k symbol, and each symbol is made up of the m bit.The RS code parameters of correcting t symbol error is as follows:

Code length n=2m-1 symbol, or m (2m-1) bit

Message segment k symbol, or km bit

Supervision section n-k=2t symbol, or m (n-k) bit

Minimum distance d=2t+1 symbol, or m (2t+1) bit

The code weight distribution of structure, the generating mode of the performance of LDPC sign indicating number and sign indicating number, code length, smallest hamming distance, check matrix ranks, decoding algorithm, the girth of two-dimensional plot, the loop distribution of two-dimensional plot, the connection performance of two-dimensional plot, the node degree distribution of two-dimensional plot etc. are relevant.For example generate to be listed as the even mode of even mode and ranks that code check is 1/2 respectively, frame length is 250 regular LDPC sign indicating number, its E_b/N_0～BER performance as shown in Figure 4, the error rate is 10^{-4}The time, adopt row modes (evencol) that the signal to noise ratio of 3dB need be provided, and adopt linescan method (evenboth) only to need 2.7dB just passable.In this example, the effect of evenboth mode is

obviously more effective than evencol.Under the situation of given identical signal to noise ratio, the error rate of evenboth mode is lower than the error rate of evencol, and along with the increase of signal to noise ratio, it is faster that the error rate of evenboth descends.

For another example, 1/2 code check of the different frame lengths that generate in the even mode of ranks rule LDPC sign indicating number, frame length divide other performance relatively 250,1000,5000, the BER error rate and E_b/N_0The relation of signal to noise ratio as shown in Figure 5, therefrom as can be seen, under each frame length, the error rate all reduces fast along with the increase of signal to noise ratio.And frame length is long more, and this decrease speed is fast more, and it is just low more to reach the identical needed signal to noise ratio of the error rate.

The LDPC sign indicating number is made first appearance from the sixties, to yielding unusually brilliant results of today, has experienced nearly half a century.In these decades, development and perfect a whole set of theory and the correlation technique of LDPC sign indicating number.The extensive use background of LDPC sign indicating number becomes the focus that various countries, world today academia and IT industry circle are paid much attention to.

The technology that any one is good only obtains practical application in reality, just can embody its real value.The LDPC sign indicating number is undoubtedly a kind of outstanding error correcting code type, the close degree of it and shannon limit is that existing other any yard type is all incomparable at present, but be still waiting improvement aspect some, for example implementation complexity, system's time delay, code efficiency etc., perhaps in other words, need be according to the demand of real application systems, the for example specific demand of foregoing ground digital television broadcast, need carry out concrete analysis and design to the problem of the aspects such as design, structure, decoding, performance evaluation and application of LDPC sign indicating number:

1) design of LDPC sign indicating number, structure and optimization

At present, people have proposed various LDPC sign indicating number designs and building method.Method based on structure is classified, and the LDPC sign indicating number is divided into two classes: random code (Random Code) and constructive code (Structured Code),

The structure of LDPC sign indicating number adopts the computer search mode at random, but need satisfy certain restrictive condition, or the sign indicating number graph structure that needs, and for example the girth of node and degree distribute.The amount of calculation that this search generating mode needs is o (n^2), the performance of abnormal LDPC is just better under long code, but amount of calculation can be bigger.Therefore, need seek to reduce Calculation Method, seek the structure of check matrix H.Can reduce amount of calculation though had some algebraically modes at present, make to have the sign indicating number type of the H of special construction for some, encoding calculation amount can be with the increase of code length linear growth.But whether these structures are optimum structure, and still whether establishment was also unknown for this conclusion when the distribution of degree was more complicated.

The method of algebraical sum combination constructs and structurized LDPC sign indicating number is based on.The long sign indicating number of LDPC at random more approaches Shannon limit than equal structured LDPC code usually, but they do not have sufficient structuring, thereby simplifies coding.On the contrary, structured LDPC code has better coding advantage usually, particularly circulation or quasi-cyclic LDPC code (Quasi-CyclicLDPC, QC-LDPC), they can adopt simple shift register to realize coding, and for serial code, its complexity linearity is proportional to the number of check bit; For parallel encoding, its complexity is proportional to code length.For the code word size of a practical application, the structurized LDPC code performance (bit error rate, Block Error Rate and error floor) of a good design that people are verified is equal to LDPC sign indicating number at random.

Therefore, seeking better or be more suitable in coded system and the LDPC code structure of the LDPC of application-specific is one of problem that must solve.

2) research of LDPC performance under non-standard channel

Present existing document all is under standard channel to the research of LDPC, especially carries out under the symmetric channel.And in ground digital television broadcast, launching tower is typical non-standard channel to moving reception, and exists the very serious multipath effect and the influence of Doppler effect.The LDPC sign indicating number will be used in mobile reception environment, and whether must study its performance

under this channel still remarkable, and seeks to be exclusively used in the optimization method under this channel.

3) more effective decoding algorithm

Various LDPC decoding algorithm is arranged at present.The simplest is the majority logic that adopts hard decision

(Majority-logic, MLG) algorithm.LDPC decoding algorithm commonly used is bit reversal (Bit-flipping, BF) algorithm, iterative decoding (the Iterative Decoding Based onBeliefProgagation that increases the weight of bit reversal (Weighted BF) algorithm and transmit based on reliability, IDBP) algorithm etc. is three kinds, IDBP sum-product algorithm (Sum-Product Algorithm just wherein, SPA), at above-mentioned decoding algorithm with in realizing, the SPA algorithm has provided best performance.

Therefore, seeking more effective, the better decoding algorithm of performance also is the focus of current LDPC research.

4) hardware of LDPC is realized

Because the amount of calculation of LDPC is big, therefore, adopt special chip to realize that the Code And Decode of LDPC will improve system effectiveness greatly, the dsp chip of research and utilization special use realizes that the coding and decoding function of LDPC sign indicating number also is a big focus.

Certainly also have some other problems, in the ground digital television broadcast practical application, we will face these problems, have only these problems are provided one to take all factors into consideration and solution, just can obtain a suitable LDPC sign indicating number

Summary of the invention

The objective of the invention is various demands, for example different modulation systems at China's ground digital television broadcast transmission; Be convenient to bit stream interface, carry an integer TS code block with TS; Be consistent with the OFDM modulation length; For better travelling performance is provided, time-varying characteristics are limited in the OFDM frame, have proposed a kind of multi code Rate of Chinese character error correction/encoding method based on BCH code and LDPC

sign indicating number serially concatenated, by different LDPC sign indicating number structures, obtain multiple encoder bit rate, so that be suitable for different channel situation.The present invention has adopted short code length, and implementation structure is simple, has less time delay and good error-correcting performance, adapts to various application better.

The present invention proposes the error correction/encoding method of a kind of employing BCH (762,752) and LDPC (7488,3048) cascaded code, it is characterized in that, may further comprise the steps:

1) TS of input is transmitted per 2 of stream bag and be divided into one group, length is 3008 bits, i.e. 188 * 8 * 2=3008 bit, and then be further divided into 4 son groups, length of each son group is 752 bits;

2) 752 information bits to 4 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 4 * 762=3048 bit;

3) to 4 BCH (762,752) sign indicating number C_{BCH}3048 bits of output carry out LDPC (7493,3048) coding, promptly predetermined with LDPC (7493,3048) generator matrix G_{Qc1}Multiply each other, obtain QC-LDPC (7493,3048) sign indicating number C_{Qc1}, described generator matrix G_{Qc1}Structure based on predetermined check matrix H_{Qc1}

4) deletion LDPC (7493,3048) sign indicating number C_{Qc1}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:

1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;

2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+ x^{3}+ 1$, export 1023 error correction coding bits;

3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, the check matrix H of described LDPC (7493,3048) $_{Qc1}$Constitute by cyclic permutation matrices:

H qc 1 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1

Wherein, γ=35, ρ=59, B $_{I, j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1.

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) check matrix H $_{Qc1}$Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:

1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal $_{RS}$, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;

2) from C $_{RS}$Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$

3) structure codeword set C $_{b}$ $^{(i)}$: c is C $_{b}$In the code word that code weight is ρ, C $_{b}$Middle q-1 non-zero codeword constitutes the set C of one dimension code word $_{b}$ $^{(1)}$={ β c: β ∈ GF (p ^{s}), based on subcode C $_{b}$ $^{(1)}$, by one of ring shift right, C $_{b}$Be divided into q-1 coset, i.e. C $_{b}$ $^{(1)}$, C $_{b}$ $^{(2)}$, L C $_{b}$ $^{(q-1)}$, so any coset C $_{b}$ $^{(i)}$In two code words all be different in all positions;

4) by C $_{b}$ $^{(i)}$Structure character position set of vectors Z (C $_{b}$ $^{(i)}$): z=(z $_{0}$, z $_{1}$, L, z $_{Q-2}$) be (q-1)-heavily on the GF (2), corresponding to GF (p ^{s}) the field element α of q-1 non-zero ^{i}, and b=(b $_{0}$, b $_{1}$, L, b $_{ρ-1}$) be C $_{b}$In code word, with z (b $_{j}$) replace each component b of b $_{j}$, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_{0}$), z (b $_{1}$) ... .., z (b $_{ρ-1}$)), C so $_{b}$ $^{(1)}$I coset C $_{b}$ $^{(i)}$In the set of character position vector of q-1 code word be Z (C $_{b}$ $^{(i)}$)={ z (b): b ∈ C $_{b}$ $^{(i)}$, Z

($C_b^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";

5) structure Z ($C_b^{(i)}$) the character position matrix A_i: for $0 \le i \le q-2$, go up structure (q-1) * ρ (q-1) rank matrix A at GF $(2)_i$, its row is Z ($C_b^{(i)}$) in q-1 character position vector;

6) by the character position matrix A_iThe basic matrix H of a rule of structure $_{GA}(\gamma, \rho)$:

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1

Its row, column weight respectively is ρ and γ, and it has its member's matrix A_iDesign feature;

7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);

8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc1}$:

H qc 1 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

Wherein

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) generator matrix G_{Qc1}Constitute by cyclic permutation matrices:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{l,j}$Be (q-1) * (q-1) circular matrix, q=128, γ=35, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1.

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) generator matrix G_{Qc1}Structure based on described check matrix H_{Qc1}, step is as follows:

1) H_{Qc}The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1

D is square formation, full rank, nonsingular, i.e. inverse matrix D^{-1}There are its order and H_{Qc}Order identical;

2) according to coding theory $H_{Qc}G_{Qc}^{T}$=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation $M_i u^{T}+ Dz_i^{T}$=0, then

$z_i^{T}=D^{-1}M_i u^{T}$

Wherein

z_i=($g_{l,0}g_{l,1}Kg_{l,\gamma-1}$) expression G_{Qc}The i of P matrix is capable in the generator matrix

M_i=[$B^{T}_{0,i}$... ..$B^{T}_{\gamma-1,i}$]^{T}Expression H_{Qc}The i row of circular matrix;

U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;

Thereby obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

3) when the non-full rank situation of matrix D, to G_{Qc}The correction of structure:

A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;

B) by $z_i^T = D^{-1} M_i u^T$Try to achieve z_iMiddle r linearity be element independently;

C) at G_{Qc}Middle γ (the q-1)-r linearity of adding is independently gone g_0^*, $g_1^* \ldots \ldots, g^*_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{Qc}Expression formula:

G * qc = G - Q = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1

And g_i^*Obtain by finding the solution following matrix equality:

B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g * i , ρ (q - 1) - 1 = 0 .

The error correction/encoding method of described a kind of employing BCH (762,752) and LDPC (7488,4572) cascaded code is characterized in that, may further comprise the steps:

1) TS of input is transmitted per 3 of stream bag and be divided into one group, length is 4512 bits, i.e. 188 * 8 * 3=4512 bit, and then be further divided into 6 son groups, length of each son group is 752 bits;

2) 752 information bits to 6 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 6 * 762=4572 bit;

3) to 6 BCH (762,752) sign indicating number C_{BCH}4572 bits of output carry out LDPC (7493,4572) coding, promptly with the generator matrix G of LDPC (7493,4572) $_{Qc2}$Multiply each other, obtain QC-LDPC (7493,4572) sign indicating number C_{Qc2}, described generator matrix G_{Qc2}Structure based on predetermined check matrix H_{Qc2}

4) deletion LDPC (7493,4572) sign indicating number C_{Qc2}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,4572) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:

1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;

2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+ x^{3}+ 1$, export 1023 error correction coding bits;

3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,4572) cascaded code is characterized in that, the check matrix H of described LDPC (7493,4572) $_{Qc2}$Constitute by cyclic permutation matrices:

H qc 2 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1

Wherein, γ=23, ρ=59, $B_{I, j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1.Described check matrix H_{Qc2}Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:

1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal $_{RS}$, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;

2) from C_{RS}Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$

3) structure codeword set $C_b^{(i)}$: c is C_bIn the code word that code weight is ρ, C_bMiddle q-1 non-zero codeword constitutes the set C of one dimension code word $_b{}^{(1)}$={ β c: β ∈ GF (p^s), based on subcode $C_b^{(1)}$, by one of ring shift right, C_bBe divided into q-1 coset, i.e. $C_b^{(1)}$, $C_b^{(2)}$, L $C_b^{(q-1)}$, so any coset $C_b^{(i)}$In two code words all be different in all positions;

4) by $C_b^{(i)}$Structure character position set of vectors Z ($C_b^{(i)}$): z=(z_0, z_1, L, z_{Q-2}) be (q-1)-heavily on the GF (2), corresponding to GF (p^s) the field element α of q-1 non-zero ^{i}, and b=(b_0, b_1, L, $b_{\rho-1}$) be C_bIn code word, with z (b_j) replace each component b of b_j, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_0$), z (b_1) ... .., z ($b_{\rho-1}$)), C so $_b{}^{(1)}$I coset $C_b^{(i)}$In the set of character position vector of q-1 code word be Z ($C_b^{(i)}$)={ z (b): b ∈ $C_b^{(i)}$, Z ($C_b^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";

5) structure Z ($C_b^{(i)}$) the character position matrix A_i: for 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF (2) $_i$, its row is Z ($C_b^{(i)}$) in q-1 character position vector;

6) by the character position matrix A_iThe basic matrix H of a rule of structure $_{GA}$(γ, ρ):

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1

Its row, column weight respectively is ρ and γ, and it has its member's matrix A_iDesign feature;

7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);

8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc2}$:

H qc 2 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

Wherein

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,4572) cascaded code is characterized in that, described LDPC (7493,4572) generator matrix G_{Qc2}Constitute by cyclic permutation matrices:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G & rho; - γ - 1,1 L G ρ - γ - 1 , γ - 1

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{i, j}$Be (q-1) * (q-1) circular matrix, q=128, γ=23, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1.Described generator matrix G_{Qc2}Structure based on described check matrix H_{Qc2}, step is as follows:

1) H_{Qc}The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B & gamma; - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B & gamma; - 1 , ρ - 1

D is square formation, full rank, nonsingular, i.e. inverse matrix D^{-1}There are its order and H_{Qc}Order identical;

2) according to coding theory $H_{Qc}G_{Qc}^{T}$=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation $M_i u^{T}$+ Dz_i^{T}=0, then

z_i^{T}＝$D^{-1}M_i u^{T}$

Wherein

z_i=($g_{i, 0}g_{i, 1}$K$g_{i, \gamma-1}$) expression G_{Qc}The i of P matrix is capable in the generator matrix

M_i=[$B^{T}_{0, i}$… ..$B^{T}_{\gamma-1, i}$]^{T}Expression H_{Qc}The i row of circular matrix;

U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;

Thereby obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

3) when the non-full rank situation of matrix D, to G_{Qc}The correction of structure:

A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;

B) by $z_i^T=D^{-1}M_i u^T$Try to achieve z_iMiddle r linearity be element independently;

C) at G_{Qc}Middle γ (the q-1)-r linearity of adding is independently gone g_0^*, $g_1^*\ldots\ldots, g^*_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{Qc}Expression formula:

G * qc = G - Q = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1

And g_i^*Obtain by finding the solution following matrix equality:

B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g * i , ρ (q - 1) - 1 = 0 .

The error correction/encoding method of described a kind of employing BCH (762,752) and LDPC (7488,6096) cascaded code is characterized in that, may further comprise the steps:

1) TS of input is transmitted per 4 of stream bag and be divided into one group, length is 6016 bits, i.e. 188 * 8 * 4=6016 bit, and then be further divided into 8 son groups, length of each son group is 752 bits;

2) 752 information bits to 8 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 8 * 762=6096 bit;

3) to 8 BCH (762,752) sign indicating number C_{BCH}6096 bits of output carry out LDPC (7493,6096) coding, promptly with the generator matrix G of LDPC (7493,6096) $_{Qc3}$Multiply each other, obtain QC-LDPC (7493,6096) sign indicating number C_{Qc3}, described generator matrix G_{Qc3}Structure based on predetermined check matrix H_{Qc3}

4) deletion LDPC (7493,6096) sign indicating number C_{Qc3}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,6096) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:

1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;

2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+ x^{3}+ 1$, export 1023 error correction coding bits;

3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,6096) cascaded code is characterized in that, the check matrix H of described LDPC (7493,6096) $_{Qc3}$Constitute by cyclic permutation matrices:

H qc 3 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1

Wherein, γ=11, ρ=59, $B_{I, j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1.Described check matrix H_{Qc3}Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:

1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal $_{RS}$, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;

2) from C $_{RS}$Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$

3) structure codeword set C $_{b}$ $^{(i)}$: c is C $_{b}$In the code word that code weight is ρ, C $_{b}$Middle q-1 non-zero codeword constitutes the set C of one dimension code word $_{b}$ $^{(1)}$={ β c: β ∈ GF (p ^{s}), based on subcode C $_{b}$ $^{(1)}$, by one of ring shift right, C $_{b}$Be divided into q-1 coset, i.e. C $_{b}$ $^{(1)}$, C $_{b}$ $^{(2)}$, L C $_{b}$ $^{(q-1)}$, so any coset C $_{b}$ $^{(i)}$In two code words all be different in all positions;

4) by C $_{b}$ $^{(i)}$Structure character position set of vectors Z (C $_{b}$ $^{(i)}$): z=(z $_{0}$, z $_{1}$, L, z $_{Q-2}$) be (q-1)-heavily on the GF (2), corresponding to GF (p ^{s}) the field element α of q-1 non-zero ^{i}, and b=(b $_{0}$, b $_{1}$, L, b $_{ρ-1}$) be C $_{b}$In code word, with z (b $_{j}$) replace each component b of b $_{j}$, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_{0}$), z (b $_{1}$) ... .., z (b $_{ρ-1}$)), C so $_{b}$ $^{(1)}$I coset C $_{b}$ $^{(i)}$In the set of character position vector of q-1 code word be Z (C $_{b}$ $^{(i)}$)={ z (b): b ∈ C $_{b}$ $^{(i)}$, Z (C $_{b}$ $^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";

5) structure Z (C $_{b}$ $^{(i)}$) the character position matrix A $_{i}$: for 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF (2) $_{i}$, its row is Z (C $_{b}$ $^{(i)}$) in q-1 character position vector;

6) by the character position matrix A $_{i}$The basic matrix H of a rule of structure $_{GA}$(γ, ρ):

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1

Its row, column weight respectively is ρ and γ, and it has its member's matrix A $_{i}$Design feature;

7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);

8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc3}$:

H qc 3 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

Wherein

The error correction/encoding method of described employing BCH (762,752) and LDPC (7488,6096) cascaded code is characterized in that, described LDPC (7493,6096) generator matrix G $_{Qc3}$Constitute by cyclic permutation matrices:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and G $_{i, j}$Be (q-1) * (q-1) circular matrix, q=128, γ=11, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1.Described generator matrix G $_{Qc3}$Structure based on described check matrix H $_{Qc3}$, step is as follows:

1) H $_{Qc}$The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1

D is square formation, full rank, nonsingular, i.e. inverse matrix D $^{-1}$There are its order and H $_{Qc}$Order identical;

2) according to coding theory H $_{Qc}$G $_{Qc}$ ^{T}=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation M $_{i}$u ^{T}+ Dz $_{i}$ ^{T}=0, then

z $_{i}$ ^{T}＝D $^{-1}$M $_{i}$u ^{T}

Wherein

$z_i=(g_{l,0}g_{l,1}Kg_{l,\gamma-1})$ expression G_{Qc}The i of P matrix is capable in the generator matrix

$M_i=[B^T_{0,i}\ldots..B^T_{\gamma-1,i}]^T$Expression H_{Qc}The i row of circular matrix;

U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;

Thereby obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

3) when the non-full rank situation of matrix D, to G_{Qc}The correction of structure:

A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;

B) by $z_i^T=D^{-1}M_iu^T$Try to achieve z_iMiddle r linearity be element independently;

C) at G_{Qc}Middle γ (the q-1)-r linearity of adding is independently gone g_0^*, $g_1^*\ldots..$, $g^*_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{Qc}Expression formula:

G * qc = G - Q = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1

And g_i^*Obtain by finding the solution following matrix equality:

B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g * i , ρ (q - 1) - 1 = 0 .

Characteristics of the present invention:

The present invention be directed to various characteristics in the ground digital television broadcast to the requirement of channel error correction coding and proposed a kind of multi code Rate of Chinese character error correction/encoding method, the method adopts the mode of BCH code and LDPC sign indicating number serially concatenated, obtain multiple encoder bit rate, so that be suitable for different channel situation.The code length that the LDPC sign indicating number of very long (64800bit) that adopts in the systems such as DVB-S2, the LDPC sign indicating number tool that the present invention proposes are short, the encoding and decoding implementation structure is simple, and time delay is little, and (error floor) is low for error floor, and performance is better than the DVB-S2LDPC sign indicating number.

The present invention only needs the iteration of limited number of time just can obtain good error-correcting performance, improved the error-resilient performance of system, and be applicable to multiple transmission modulation system (QPSK, 16QAM and 64QAM etc.), can be applicable to digital television broadcastings such as ground, wired, satellite, and in the digital communication system.

Description of drawings

Fig. 1 is a DVB-S2 error correction coding performance curve

Fig. 2 is the check matrix of (20,3,4) LDPC sign indicating number of Gallagher structure

Fig. 3 represents for the two-dimensional plot of LDPC sign indicating number

Fig. 4 is identical code length, but the comparison of LDPC code performance during the different coding mode

Fig. 5 is the same-code mode, but the comparison of LDPC code performance during different code length

Fig. 6 is the frame structure of TDS-OFDM system

The error correction/encoding method coding block diagram that Fig. 7 proposes for the present invention

Fig. 8 is LDPC sign indicating number and long-pending decoding algorithm signal flow graph

Fig. 9 is for adopting the transmitting terminal block diagram of ground digital multimedia TV broad cast system of the present invention

Figure 10 is for adopting the receiving terminal block diagram of ground digital multimedia TV broad cast system of the present invention.

Figure 11 is for adopting the test performance of ground digital multimedia TV broad cast system of the present invention

Embodiment

Below in conjunction with accompanying drawing specific embodiments of the invention are described in detail.

The time-domain synchronization OFDM that adopts among the DMB-T of Tsing-Hua University (TDS-OFDM) modulation belongs to multi-transceiver technology; but the OFDM (COFDM) of the coding that adopts with European DVB-T is different; in TDS-OFDM, do not insert the pilot tone signal; but in the protection of OFDM at interval, inserted pseudorandom (PN) sequence in the mode of time domain, be used for frame synchronization, Frequency Synchronization, regularly synchronously, channel transfer characteristic is estimated and follow the tracks of phase noise etc.

For realize quick and stable synchronously, Tsing-Hua University's TDS-OFDM transmission system has adopted hierarchical frame structure.The elementary cell of frame structure is called signal frame, as shown in Figure 6.200/225 signal frame is defined as a frame group, and 512 frame groups are defined as a superframe.The top layer of frame structure is called a day frame, is made up of superframe.Each signal frame among the frame group has unique frame number, and it is coded in the PN sequence of frame head.

The signal frame of TDS-OFDM transmission system uses the OFDM modulation of Domain Synchronous, and perhaps being called with the PN sequence is protection OFDM modulation at interval.A signal frame is made up of frame synchronization and frame two parts, and they have identical baseband signalling rate 7.56MS/s (1/T).A signal frame can be used as an OFDM (OFDM) piece.An OFDM piece further is divided into a protection interval and an inverse discrete Fourier transform (IDFT) piece.For TDS-

OFDM; frame synchronization PN sequence as the protection of OFDM at interval, and frame is as the IDFT piece, length is 3780 points; wherein have be used for system parameterss such as PN length, error correcting system, modulation system at 36, the payload length that really is used for transmitting user data is 3744 points.Therefore, in order to handle conveniently, the length of error correction/encoding method and 3744 has substantial connection, its integral multiple preferably, but also to satisfy the error-correcting performance requirement simultaneously, in the present invention, the length of the error correction/encoding method that we proposed is 3744 2 times, is 7488.

Seeing grant number for details about the correlation circumstance of DMB-T, TDS-OFDM is that 00123597.4 " ground digital multimedia TV broad cast system " by name, grant number are that 01115520.5 " time-domain synchronous orthogonal frequency division multiplex modulation method " by name, grant number are ZL01130659.9 " frame-synchronization generation method in the ground digital multimedia TV broad cast system " by name, and grant number is the Chinese invention patent that 01124144.6 " protection fill method at interval in the orthogonal FDM modulation system " by name waits Tsing-Hua University to apply for.

Below we just explain the principle and the step of the Code And Decode of the LDPC sign indicating number that the present invention proposes, it comprises the FEC forward error correction coding of three kinds of code checks: code check is 0.4 FEC (7488,3008), code check is that 0.6 FEC (7488,4512) and code check are 0.8 FEC (7488,6016).

In the present embodiment, the FEC forward error correction coding has adopted BCH (762,752) and the serial concatenation of codes of LDPC sign indicating number, wherein the LDPC sign indicating number has adopted quasi-cyclic low-density check code (Quasi-cyclic LDPC Code, QC-LDPC), the notion of accurate circulation error correction coding is as described in the previous technique background.The QC-LDPC that the present invention proposes has three kinds: LDPC (7493,3048), LDPC (7493,4752) and LDPC (7493,6096).After deleting preceding 5 check bits of these three kinds of QC-LDPC cascaded codes, make the code length of the LDPC that carries become 7488 by 7493.The complete BCH proposed by the invention and the construction process of LDPC cascaded code are as follows:

1) encoder bit rate is 0.4 FEC

Data code flow → BCH (762,752) → LDPC (7493,3048) → preceding 5 the check bit → LDPC of deletion (7488,3048);

2) encoder bit rate is 0.6 FEC

Data code flow → BCH (762,752) → LDPC (7493,4572) → preceding 5 the check bit → LDPC of deletion (7488,4512);

3) encoder bit rate is 0.8 FEC

Data code flow → BCH (762,752) → LDPC (7493,6096) → preceding 5 the check bit → LDPC of deletion (7488,6096).

From the front about the description of error correction coding as can be known, generator matrix G or check matrix H are one of key factors of decision LDPC performance, the error-correcting performance that different generator matrix G obtains is just different, therefore, seek or design the generator matrix G of LDPC sign indicating number or the major issue that check matrix H just becomes the LDPC sign indicating number.QC-LDPC is like this equally, and the QC-LDPC error performance of a good design can be compared with at random rule or irregular LDPC codes.Therefore, below we at first describe the check matrix H and the generator matrix G of three QC-LDPC sign indicating numbers proposed by the invention, and then three kinds of cascade implementations of error correcting codes steps in the present embodiment are described.

As previously mentioned, circular matrix is a square formation, and its each row and each row all have the cyclic shift characteristic, and its capable weight equals the column weight amount, is w.If w=1 so just obtains cyclic permutation matrices (circulant permutationmatrix).

The check matrix H of QC-LDPC sign indicating number $_{Qc}$Be a sparse cyclic permutation matrices, in the present embodiment, H $_{Qc}$Structure adopted RS code method based on two information symbols, its process is as described below:

The first step generates RS (q-1, q-ρ, ρ-1) sign indicating number C $_{RS}$

Patrol magnificent territory GF (p at a gal ^{s}) in, wherein p is a prime number, s is a positive integer, makes q=p ^{s}, α represents GF (p ^{s}) primitive element, then a α $^{\infty}$=0, α ^{0}=1, α ^{1}, α ^{2}, α ^{3}... .., α $^{Q-2}$Constituted GF (p ^{s}) all

elements.Make that ρ is a positive integer, and $2 \le \rho < q$, the RS that circulates so (q-1, q-ρ, ρ-1) sign indicating number C_{RS}Length be q-1, information symbol length is q-ρ-1, smallest hamming distance is ρ-1, its generator polynomial is:

$g(X) = (X-\alpha)(X-\alpha^{2})L(X-\alpha^{\rho-2})$

$= g_{0} + g_{1}X + g_{2}X^{2} + L + X^{\rho-2}$

G wherein $_{i} \in GF(p^{s})$

Preceding q-ρ-1 information symbol of second step deletion from C obtains shortening RS (ρ, 2, ρ-1) sign indicating number, structure base sign indicating number C_{b}Each code word C_{RS}In the deletion of preceding q-ρ-1 information symbol, code word C_{RS}Contraction in length obtains shortening RS (ρ, 2, ρ-1) sign indicating number C_{b}, it has only 2 information symbols, and its generator matrix is:

G b = g 0 g 1 g 2 L 1 0 0 g 0 g 1 g 2 L 1

C so $_{b}$Non-zero codeword has two different code weights, ρ-1 and ρ.We claim C_{b}Be base sign indicating number (Base Code).

The 3rd step structure codeword set $C_{b}^{(i)}$

Because C_{b}Smallest hamming distance be ρ-1, so C_{b}In any two code words have identical code sign a position at the most.We need utilize C_{b}Some architectural characteristics so that construct a rule-like LDPC sign indicating number, making the two-dimensional plot of LDPC avoid length is 4 circulation, this is the basic principle of regular LDPC code check matrix H design.

Make that c is C_{b}In a code word, its code weight is ρ, then C_{b}Middle q-1 non-zero codeword set $C_{b}^{(1)} = \{ \beta c : \beta \in GF(p^{s}) \}$ constituted C_{b}The one dimension subcode, $C_{b}^{(1)}$In each non-zero codeword have code weight ρ.$C_{b}^{(1)}$In two code words all be different in each position.Based on subcode $C_{b}^{(1)}$, by one of ring shift right, C_{b}Be divided into q-1 coset (coset), $C_{b}^{(1)}$, $C_{b}^{(2)}$, L $C_{b}^{(q-1)}$, so any coset $C_{b}^{(i)}$In two code words all be different in all positions.

If coset $C_b^{(i)}$In q-1 non-zero codeword be expressed as (q-1) * ρ rank matrix, then in the matrix arbitrarily all q-1 element of row all be inequality.

The 4th step is by $C_b^{(i)}$Structure character position set of vectors $Z(C_b^{(i)})$

Make $z=(z_0, z_1, L, z_{Q-2})$ be (q-1)-heavy (array) on the GF (2), its component composition is corresponding to GF (p^s) the element of q-1 non-zero, i.e. z_iField element α corresponding to non-zero ^{i}We claim α^iBe z_iLocation number (LocationNumber).For i=0,1, L, q-2, we define α^iPosition vector (Location Vector) be (q-1)-heavy on the GF (2), i component z wherein $_i$Equal 1, other all q-2 component equals 0.

Make $b=(b_0, b_1, L, b_{\rho-1})$ be C_bIn code word, for 0≤j≤ρ-1, with its position vector $z(b_j)$ replace each component b of b_j, then obtain ρ (the q-1)-weight on the GF (2):

$$z(b)=(z(b_0),\ z(b_1),\ .....,\ z(b_{\rho-1}))$$

Its code weight is ρ, is called the character position vector of b.Because C_bIn any two code words have identical code sign a position at the most, thereby their character position vector z (b) has 1 " 1-component " to have at the most.

Make $Z(C_b^{(i)})=\{ z(b): b \in C_b^{(i)}$Be $C_b^{(1)}$I coset $C_b^{(i)}$In the set of character position vector of q-1 code word, according to front $C_b^{(1)}$Coset $C_b^{(i)}$Architectural characteristic as can be known, $Z(C_b^{(i)})$ in any two character position vectors be that have or identical without any " 1-component ".

The 5th step structure $Z(C_b^{(i)})$ the character position matrix A_i

For 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF $(2)_i$, its row is $Z(C_b^{(i)})$ in q-1 character position vector.

Because Z is $(C_b^{(i)})$ in the weight of each vector be ρ, so A_iIn " 1-clauses and subclauses (entries) " add up to ρ (q-1).Because Z is $(C_b^{(i)})$ in two character position vectors be identical without any " 1-component ", so A_iThe weight of every row is 1.Therefore, A_i(1, ρ)-regular matrix, its row, column weight respectively is ρ and 1 to be one.According to C_bThe definition of the character position vector of middle code word, and each coset $C_b^{(i)}$The design feature of middle code word, A_iBe (q-1) * ρ (q-1) rank permutation

matrixes, matrix A $_i$Be called character position matrix (Symbol Loacation Matrix).

The character position matrix A $_i$The class (class) that constitutes, A={A $_0$, A $_1$... .., A $_{Q-2}$, have following architectural characteristic: the matrix A that (1) is identical $_i$In do not have two row to have total " 1-component "; (2) from two different matrix A $_i$And A $_j$Total " the 1-components " of two row can not surpass 1.

The basic matrix H of a LDPC code check matrix of the 6th step structure $_{GA}$(γ, ρ)

Make that γ is a positive integer, and 1≤γ≤(q-1), construct the basic matrix (BaseMatrix) on γ (q-1) * ρ (q-1) rank of following GF (2):

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , &rh o; - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A &gamm a; - 1,1 L A γ - 1 , ρ - 1

This matrix be one (γ, ρ)-regular matrix, its row, column weight respectively is ρ and γ, it has its member's matrix A $_i$Design feature, i.e. H $_{GA}$(γ, ρ) any two row (or two row) have 1 " 1-component " to have at the most.This represents H $_{GA}$(γ ρ) can not have 4 each " 1 " on middle four angles of rectangle, and this will be avoided H $_{GA}$(it is 4 circulation that γ, two-dimensional plot ρ) have length, and its girth is at least 6.

This H $_{GA}$(γ, ρ) matrix is the canonical form of Regulation G allager LDPC generator matrix, and the LDPC code length of its generation is n=ρ (q-1), and code check is at least (ρ-γ)/ρ.

Because H $_{GA}$(γ, ρ) the total quantity of " 1-component " is no more than 1 in two row, and the weight of the every row of matrix is γ, so H $_{GA}$(γ, minimum distance d ρ) $_{Min}$(γ) be at least γ+1.If consideration check matrix H $_{GA}$(γ, structure ρ), minimum distance lower limit d $_{Min}$(γ) that change of possibility:

Masking matrix W of the 7th step structure (γ, ρ)

The basic matrix of above-mentioned generation be one (γ, ρ)-check matrix of regular QC-LDPC sign indicating number, by sheltering selected H $_{GA}$(γ, the ρ) set of cyclic permutation matrices can obtain the check matrix H of

various QC-LDPC sign indicating numbers $_{Qc}$(γ, ρ).Masked operation can pass through masking matrix W (γ, ρ) and H $_{GA}$(γ, ρ) matrix multiple obtains.

If (γ ρ) is regular matrix to masking matrix W, and row and/or row have identical code weight, then H $_{Qc}$(γ ρ) is regular matrix, and the code word of acquisition is regular QC-LDPC sign indicating number; If (γ ρ) has variable column weight amount and/or variable capable weight, then H to W $_{Qc}$(γ is non-regular matrix ρ), the code word right and wrong rule QC-LDPC sign indicating number that obtains.

(γ ρ) is γ * ρ rank matrix on the GF (2) to masking matrix W, and γ, ρ are less usually, are generally less than 500.

Masking matrix with very big girth can be constructed by computer search, participates in following document:

(11)X.-Y?Hu, E.Eleftheriou, and?D.-M.Arnold, “Progressive?edge-growth?Tanner?graphs, ”Proc.2001?GlobeCom, pp.995-1001, Nov.25-29, 2002.

(12)L.Lan, Y.Y.Tai, L.Chen, S.Lin?and?K.Abdel-Ghaffar, “A?trellis-based?method?forremoving?cycles?from?bipartite?graphs?and?constructi on?of?LDPC?codes, ”IEEE?Communicat?ionLetters, vol.8, no.7, July?2004.

(γ ρ) can construct by algebraic approach the masking matrix W of rule, the minimum code weight code word of for example circular matrix decomposition, how much decomposition, the non-complete design of balance, stack, product graph, Hamming code etc.

(γ, ρ), (γ, row, column ρ) has the weight that does not wait respectively to non-regular masking matrix W and present embodiment has adopted the masking matrix W of non-rule.Thereby its two-dimensional plot has the variable node degree of variation and/or the check-node degree of variation.The degree distribution table of this two category node is shown two polynomial forms:

v (X) = Σ i = 1 dv v i X i - 1

c (X) = Σ i = 1 dc c i X i - 1

V wherein $_{i}$And c $_{i}$Degree of a representation is the decimal of variable node and check-node in the two-dimensional plot of i respectively, d $_{v}$And d $_{c}$The maximal degree of representing variable node and check-node respectively.

Because the variable node of two-dimensional plot and check-node are corresponding to the row and the row of check matrix H, therefore, v (X) and c (X) have also just provided the row and the row distribution of weight of check matrix H.People have proved that the error performance of irregular LDPC codes depends on that the degree of variable node and check-node distributes in its two-dimensional plot, by the probability density distribution that message is transmitted between this two category node in the emulation BP algorithm, optimize these two degree distribution v (X) and c (X), can design the LDPC sign indicating number that approaches celestial farming limit.

According to desired code word size n and encoder bit rate R, select suitable γ, ρ and q.

In the sign indicating number structure, in case designed degree distribution v (X) and c (X), (γ ρ), makes that (γ, the distribution of weight of row ρ) and row equals (or approaching) degree distribution v (X) and c (X) to W just can to determine W by computer with random fashion according to the degree distribution.

The 8th step generated the check matrix H of QC-LDPC $_{Qc}$=W * H $_{GA}$

The masking matrix W of above-mentioned structure (γ, ρ)=[w $_{I, j}$] be the γ * ρ rank matrix on the GF (2), the check matrix H of QC-LDPC sign indicating number $_{Qc}$(γ ρ) obtains by following matrix multiplication:

H qc (γ , ρ) = W (γ , ρ) × H GA (&gamma ; , ρ)

= w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

= B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B & gamma; - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1

Wherein

Promptly by (γ, ρ) multiplying each other of " 0 " item masks H simply with W_{GA}(γ, some permutation matrix set A ρ) $_{I, j}$H $_{Qc}$(γ, ρ) middle displacement matrix A $_{I, j}$Distribution be equal to W (γ, ρ) in the distribution of " 1 ".Reduced H because shelter $_{GA}$(γ, ρ) in the density of " 1 ", promptly masked operation has kept the constraint of RC ranks, thereby H $_{Qc}$(γ ρ) also satisfies the ranks constraint, and H $_{Qc}$(γ, the number of the short circulation of two-dimensional plot ρ) will be less than H $_{GA}$(γ, ρ).If (γ, ρ) girth of two-dimensional plot is more than or equal to 6, then H for W $_{Qc}$(γ, girth ρ) is at least 6.

In the present embodiment, the QC-LDPC check matrix H that generates based on the RS sign indicating number of 2 information symbols $_{GA}$(γ, ρ) be the rule, but masking matrix W (γ ρ) is non-regular matrix.So, the check matrix right and wrong rule that the QC-LDPC that both obtain behind the matrix multiple is final, the QC-LDPC sign indicating number right and wrong rule that promptly adopts in the present embodiment.

Comprehensive above-mentioned steps, the check matrix H that the QC-LDPC sign indicating number is final $_{Qc}$(γ, it is as follows ρ) to have a form:

H qc (γ , ρ) = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1 - - - (1)

Wherein, B $_{I, j}$Be (q-1) * (q-1) rank matrix, 0≤i≤γ-1 and 0≤j≤ρ-1.Than its big or small q-1, weight w is very little.For the QC-LDPC sign indicating number, adopted w=1 among the present invention.As above-mentioned definition, B $_{I, j}$It is cyclic permutation matrices.

Order is for the B in the matrix (1) $_{I, j}$=k, 0≤k≤q-2, it means that from left to right each row is designated 0 to q-1, wherein has only k to classify " 1 " as in the 0th row of this cyclic permutation matrices, other q-2 row all are " 0 ".In the remaining row (the 1st row is capable to q-2), each row is the cyclic shift (moving right) of its lastrow, and the 0th row is actually the capable cyclic shift of q-2.Therefore, this cyclic permutation matrices B $_{I, j}$Structure can describe by " k " value fully.For example, if k=3, q=8, then B $_{I, j}$=3 cyclic permutation matrices are as follows:

B ij = 0 0 0 1 0 0 0 0 0 0 0 1 0 0 0 0 0 0 0 1 0 0 0 0 0 0 0 1 1 0 0 0 0 0 0 0 1 0 0 0 0 0 0 0 1 0 0 0 0

The parameter of the check matrix of three kinds of QC-LDPC sign indicating numbers that propose in the present embodiment is respectively:

1) check matrix H of LDPC (7493,3048) sign indicating number $_{Qc1}$Form shown in following formula (1), γ=35 wherein, ρ=59 and q=p ^{s}=2 ^{7}=128, code length n=ρ * (q-1)=59 * 127=7493, information bit length k=(ρ-γ) * (q-1)=24 * 127=3048.

2) check matrix H of LDPC (7493,4572) sign indicating number $_{Qc2}$Form shown in following formula (1), γ=23 wherein, ρ=59 and q=128, code length n=ρ * (q-1)=59 * 127=7493, information bit length k=(ρ-γ) * (q-1)=4572.

3) check matrix H of LDPC (7493,6096) sign indicating number $_{Qc3}$Form shown in following formula (1), γ=11 wherein, ρ=59 and q=128, code length n=ρ * (q-1)=59 * 127=7493, information bit length k=(ρ-γ) * (q-1)=6096.

QC-LDPC sign indicating number receiving end that present embodiment the proposed needed check matrix H of decoding has been described above $_{Qc}$Generative process, but at transmitting terminal, QC-LDPC sign indicating number coding needs a generator matrix G $_{Qc}$, below the check matrix H how circulation form is arranged is just described $_{Qc}$Obtain the generator matrix G of systemic circulation form $_{Qc}$

By check matrix H $_{Qc}$The QC-LDPC code table that provides is shown C $_{Qc}$, suppose H $_{Qc}$Order r equal H $_{Qc}$The line number of matrix, i.e. (ρ-γ) * (q-1) the bit correspondence information bit for r=γ (q-1), and Hqc preceding.We are H $_{Qc}$The column weight of circular matrix newly arrange, make following H $_{Qc}$The order of γ * ρ rank submatrix be full rank γ (q-1), with H $_{Qc}$Order r identical:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 - - - (2)

The QC-LDPC sign indicating number C that requires $_{Qc}$Generator matrix G $_{Qc}$Have following systemic circulation permutation matrix form:

G qc = G 0 G 1 M G ρ - γ - 1

= I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1

= I (ρ - γ) (q - 1) P

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{i,j}$Be (q-1) * (q-1) rank circular matrixes, 0≤i≤ρ-γ-1,0≤j≤γ-1.

The generator matrix G of following formula form $_{Qc}$Be so-called systemic circulation (Systematic Circulant, SC) form, it is made up of two parts, and left-hand component I is that leading diagonal is that ((q-1) * (q-1) rank unit matrix of ρ-γ), right-hand component P are that (q-1) * (q-1) rank circular matrix constitutes (ρ-γ) * γ order array.By coding theory as can be known, G $_{Qc}$Be the systematic code form, its left-hand component P is called the P matrix, corresponding to the check digit of systematic code.The P matrix is the array of circular matrix, has the SC form.

G $_{Qc}$Be C $_{Qc}$The necessary and sufficient condition of generator matrix is H $_{Qc}$G $_{Qc}$ ^{T}=| O|, here | O| is γ (q-1) * ((q-1) rank of ρ-γ) zero battle array.For 0≤i≤ρ-γ-1 and 0≤j≤γ-1, make $g_{i,j}$Expression circular matrix $G_{i,j}$Generator polynomial.In a single day we known these $g_{i,j}$, just can construct G $_{Qc}$All circular matrix $G_{i,j}$Therefore, the characteristic of Gqc fully can be by γ * (ρ-γ) set of rank circular matrix production is characterized, and is called QC-LDPC sign indicating number C $_{Qc}$Generator matrix.

Make u=(1 0 K 0) represent that " 1 " is positioned at unit (the q-1)-weight of beginning, and 0=(0 0 K 0) expression complete zero (q-1)-heavy.For 0≤i≤ρ-γ-1, G_{i}The i of submatrix capablely be:

Wherein, unit (q-1)-weight u is positioned at g_{i}The i position.

If G $_{Qc}$Be C $_{Qc}$Generator matrix, then H must be arranged $_{Qc}g_{i}$ ^{T}=0.Make $z_{i}=(g_{i,0}g_{i,1}K\ g_{i,\gamma-1})$ expression G $_{Qc}$The i of P matrix is capable in the generator matrix, $M_{i}=[B^{T}_{0,i}\ldots\ldots B^{T}_{\gamma-1,i}]$ ^{T}Expression H $_{Qc}$The i row of circular matrix.H so $_{Qc}g_{i}$ ^{T}=0 becomes following equation:

$M_{i}u^{T}+Dz_{i}^{T}=0$

Because D is previously defined square formation in the following formula, it is a full rank, nonsingular, and inverse matrix D^{-1}Exist.

Therefore, get by following formula:

$z_i^T = D^{-1} M_i u^T$???????????(3)

D, M and u are known quantities in this formula, find the solution this equation, obtain $z_0, z_1 \ldots, z_{\gamma-1}$Because $z_i=(g_{l,0} g_{l,1} K g_{l,\gamma-1})$, therefore, from $z_0, z_1 \ldots, z_{\gamma-1}$Can obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of QC-LDPC sign indicating number $_{Qc}$

Describe in the above by H_{Qc}Find the solution G_{Qc}Process in, we suppose H_{Qc}Order r equal H_{Qc}The line number of matrix, i.e. $r=\gamma(q-1)$, but also may have the situation of $r<\gamma(q-1)$, we also need to handle how to find the solution G in this case $_{Qc}$

In this case, still make $\gamma * \gamma$ rank D rank of matrix equal H_{Qc}Order r, but $r<\gamma(q-1)$, this shows that D is not a full rank, has linear dependence between some row.In order to make above-mentioned formula (3)

$z_i^T = D^{-1} M_i u^T$

Unique separating arranged, and we are z_iThe element of middle $\gamma(q-1)-r$ linear correlation is made as 0, and these elements are corresponding to the relevant row of D matrix neutral line.We just can obtain z by following formula so $_i$Middle r linearity be element independently.

Thus, we have obtained γ (the individual circular matrix G of $\rho-\gamma)_{l,j}$γ (the individual generator polynomial g of $\rho-\gamma)_{l,j}$Set out by these productions, can construct the generator matrix G of a systemic circulation structure, this matrix generates C_{Qc}Subcode, dimension is $(\rho-\gamma)(q-1)$.In order to generate whole code word C_{Qc}, must add $\gamma(q-1)-r$ linearity at G and independently go, to constitute C_{Qc}Complete generator matrix G_{Qc}

Make $g_0^*, g_1^* \ldots \ldots, g^*_{\gamma(q-1)-r-1}$Those row that expression is added.For $0 \le i < \gamma(q-1)$, i the form of the composition that adds line is:

$g_i^* = (0 K ? 0 K ? 0 ? g^*_{i,(\rho-\gamma)(q-1)} g^*_{i,(\rho-\gamma)(q-1)+1} ? K ? g^*_{i,\rho(q-1)-1})$

It has following architectural characteristic:

It is (1) preceding that (ρ-γ) (q-1) bit is 0;

(2) $z_i=(g^*_{l,(\rho-\gamma)(q-1)}g^*_{l,((q-1)+1\ of\ \rho-\gamma)}K\ g^*_{l,\rho(q-1)-1})$ γ (q-1)-r bit corresponding to the relevant row of D neutral line, be called related bits (Dependent bit), related bits have (0 ... .., 0,1,0 ..., 0) form, the position of i related bits of " 1 " expression.

With top the same, the capable g that adds among the G_i^*Also must satisfy equation H qc g i * T = 0 , 0≤i＜γ (q-1)-r-1, thus column matrix obtained down:

B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g * i , ρ (q - 1) - 1 = 0 - - - (4)

Separate this matrix, obtain all g_i^*

In G, add g^*_0, g^*_1K $g^*_{\gamma(q-1)-r-1}$OK, obtain QC-LDPC sign indicating number C in such cases $_{Qc}$Generator matrix:

G * qc = G - Q = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1 - - - (5)

G^*_{Qc}All provisional capitals be linearity independently, thereby its order is ρ (q-1)-r.

Make the v information encoded bit of indicating, then corresponding QC-LDPC code word C_{Qc}For:

$C_{qc}=v\cdot G_{qc}$

Or

$C_{qc}=v\cdot G^*_{qc}$

In sum, the QC-LDPC sign indicating number C that is proposed in the present embodiment $_{Qc}$Generator matrix G $_{Qc}$Constitution step be:

The first step is by H $_{Qc}$Structural matrix D, D are square formation, full rank, nonsingular, i.e. inverse matrix D $^{-1}$There are its order and H $_{Qc}$Order identical;

Second step is according to coding theory, by equation M $_{i}$u ^{T}+ Dz $_{i}$ ^{T}=0 gets formula (3):

z $_{i}$ ^{T}=D $^{-1}$M $_{i}$u ^{T}

Wherein

z $_{i}$=(g $_{i, 0}$g $_{i, 1}$K?g $_{i, \gamma-1}$)

M $_{i}$=[B ^{T} $_{0, i}$.....B ^{T} $_{\gamma-1, i}$] ^{T}

u=(1?0?K?0)

Thereby obtain G $_{Qc}$Middle circular matrix G $_{I, j}$All generator polynomial g $_{I, j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

The 3rd step is when the non-full rank situation of matrix D, to G $_{Qc}$The correction of structure:

1) z $_{i}$In be made as 0 corresponding to the element of D matrix neutral line related column, the quantity of linear correlation row is γ (q-1)-r;

2) by z $_{i}$ ^{T}=D $^{-1}$M $_{i}$u ^{T}Try to achieve z $_{i}$Middle r linearity be element independently;

3) at G $_{Qc}$Middle γ (the q-1)-r linearity of adding is independently gone g $_{0}$ *, g $_{1}$ *... .., g * $_{\gamma (q-1)-r-1}$, to constitute complete generator matrix G $_{Qc}$Expression formula (5), and g $_{i}$ *Obtain by matrix equality (4).

Through above-mentioned steps, obtain QC-LDPC sign indicating number C at last $_{Q, c}$Generator matrix G $_{Qc}$Shown in the following matrix:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G & rho; - γ - 1,1 L G ρ - γ - 1 , γ - 1 - - - (6)

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{i,j}$Be (q-1) * (q-1) circular matrix, make m=ρ-γ, 0≤i≤m-1,0≤j≤γ-1.

The generator matrix G of three kinds of QC-LDPC sign indicating numbers that propose in the present embodiment $_{Qc}$Parameter be respectively:

1) the generator matrix G of LDPC (7493,3048) sign indicating number $_{Qc1}$Has the matrix form shown in the above-mentioned formula (6), parameter m=24 wherein, γ=35 and q=128.

2) the generator matrix G of LDPC (7493,4572) sign indicating number $_{Qc2}$Has the matrix form shown in the above-mentioned formula (6), parameter m=36 wherein, γ=23 and q=128.

3) the generator matrix G of LDPC (7493,6096) sign indicating number $_{Qc3}$Has the matrix form shown in the above-mentioned formula (6), parameter m=48 wherein, γ=11 and q=128.

The check matrix H of Miao Shuing in the above $_{Qc}$With generator matrix G $_{Qc}$On the basis of building method, the full implementation step of the three kinds of forward direction cascade error correcting codes (BCH+QC-LDPC) that propose in the present embodiment is described below.

Error correction/encoding method of the present invention is used for the embodiment one that ground digital television broadcast is made a start, and the FEC forward error correction coding has adopted the serially concatenated of BCH (762,752) sign indicating number and LDPC (7488,3048) sign indicating number, and as shown in Figure 7, implementation step is as follows:

1) per 2 TS of the information flow of input are transmitted stream packet and be divided into one group, length is 3008 bits, i.e. 188byte * 2=3008bit, and then be further divided into 4 son groups, length of each son group is 752 bits;

2) 752 information bits to 4 son groups carry out binary system BCH (762,752) error correction coding respectively, and the output length that obtains after the error correction coding is 4 * 762=3048 bit, wherein BCH (762,752) be the shortening sign indicating number of BCH (1023,1013), its generator polynomial is $x^{10}+ x^{3}+ 1$, shorten sign indicating number and generate according to the following steps:

A) before 752 data bits, increase by 261 " 0 " bits, form the information of 1013 bits;

B) these 1013 information bits are carried out BCH (1023,1013) error correction coding, obtain 1023 error correction coding bits;

C) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

3) to 4 BCH (762,752) sign indicating number C $_{BCH}$3048 bits of output carry out LDPC (7493,3048) coding, promptly with the generator matrix G of LDPC (7493,3048) $_{Qc1}$Multiply each other, obtain QC-LDPC (7493,3048) sign indicating number C $_{Qc1}$

4) deletion LDPC (7493,3048) sign indicating number C $_{Qc1}$Preceding 5 check bits of 7493 bits of output obtain 7488 output bits;

5) 7488 error correction coding bits that obtain are carried out corresponding QPSK or mQAM sign map;

6) carry out time-domain synchronization OFDM (TDS-OFDM) modulation, and then through giving Channel Transmission after the processing such as shaping filter, baseband signal frame up conversion.

Error correction/encoding method of the present invention is used for the embodiment two that ground digital television broadcast is made a start, and the FEC forward error correction coding has adopted BCH (762,752) sign indicating number and LDPC (7488,4572) Ma serially concatenated, its block diagram is similar to Fig. 7, and concrete parameter is variant, and implementation step is as follows:

1) per 3 TS of the information flow of input are transmitted stream packet and be divided into one group, length is 4512 bits, i.e. 188byte * 3=4512bit, and

then be further divided into 6 son groups, length of each son group is 752 bits;

2) 752 information bits to 6 son groups carry out binary system BCH (762,752) error correction coding respectively, and the output length that obtains after the error correction coding is 6 * 762=4572 bit, wherein BCH (762,752) be the shortening sign indicating number of BCH (1023,1013), its generator polynomial is $x^{10}+ x^{3}+ 1$, shorten sign indicating number and generate according to the following steps:

A) before 752 data bits, increase by 261 " 0 " bits, form the information of 1013 bits;

B) these 1013 information bits are carried out BCH (1023,1013) error correction coding, obtain 1023 error correction coding bits;

C) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

3) to 6 BCH (762,752) sign indicating number C $_{BCH}$4572 bits of output carry out LDPC (7493,4572) coding, promptly with the generator matrix G of LDPC (7493,4572) $_{Qc2}$Multiply each other, obtain QC-LDPC (7493,4572) sign indicating number C $_{Qc2}$

4) deletion LDPC (7493,4572) sign indicating number C $_{Qc2}$Preceding 5 check bits of 7493 bits of output obtain 7488 output bits;

5) 7488 error correction coding bits that obtain are carried out corresponding QPSK or mQAM sign map;

6) carry out time-domain synchronization OFDM (TDS-OFDM) modulation, and then through giving Channel Transmission after the processing such as shaping filter, baseband signal frame up conversion.

Except above-mentioned two embodiment, error correction/encoding method of the present invention is used for the embodiment three that ground digital television broadcast is made a start, the FEC forward error correction coding has adopted BCH (762,752) sign indicating number and LDPC (7488,6096) Ma serially concatenated, its block diagram is similar to Fig. 7, and concrete parameter is variant, and implementation step is as follows:

1) per 4 TS of the information flow of input are transmitted stream packet and be divided into one group, length is 6016 bits, i.e. 188byte * 4=6016bit, and then be further divided into 8 son groups, length of each son group is 752 bits;

2) 752 information bits to 6 son groups carry out binary system BCH (762,752) error correction coding respectively, and the output length that obtains after the error correction coding is 8 * 762=6096 bit, wherein BCH (762,752) be the shortening sign indicating number of BCH (1023,1013), its generator polynomial is $x^{10}+ x^{3}+ 1$, shorten sign indicating number and generate according to the following steps:

A) before 752 data bits, increase by 261 " 0 " bits, form the information of 1013 bits;

B) these 1013 information bits are carried out BCH (1023,1013) error correction coding, obtain 1023 error correction coding bits;

C) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

3) to 8 BCH (762,752) sign indicating number C $_{BCH}$6096 bits of output carry out LDPC (7493,6096) coding, promptly with the generator matrix G of LDPC (7493,6096) $_{Qc3}$Multiply each other, obtain QC-LDPC (7493,6096) sign indicating number C $_{Qc3}$

4) deletion LDPC (7493,6096) sign indicating number C $_{Qc3}$Preceding 5 check bits of 7493 bits of output obtain 7488 output bits;

5) 7488 error correction coding bits that obtain are carried out corresponding QPSK or mQAM sign map;

6) carry out time-domain synchronization OFDM (TDS-OFDM) modulation, and then through giving Channel Transmission after the processing such as shaping filter, baseband signal frame up conversion.

At receiving terminal, the signal that receives at first carries out the decoding of LDPC sign indicating number through after the demodulation, gives BCH (762,752) decoding then, obtains the transmission code stream after the error correction.Wherein BCH decoding is very ripe, does not adopt special

algorithm in the present embodiment, therefore, has just repeated no more at this.

Various LDPC decoding algorithm is arranged at present.LDPC decoding algorithm commonly used is bit reversal (Bit-flipping, BF) algorithm, iterative decoding (the Iterative DecodingBased on BeliefProgagation that increases the weight of bit reversal (Weighted BF) algorithm and transmit based on reliability, IDBP) algorithm etc. is three kinds, wherein IDBP just sum-product algorithm (Sum-ProductAlgorithm, SPA).

LDPC sign indicating number for the present invention proposes according to different requirements, can adopt different LDPC decoding algorithms.In above-mentioned three embodiment, we have adopted the iterative decoding algorithm that transmits based on reliability, i.e. sum-product algorithm.

BP (belief-propagation, reliability propagate) algorithm is combined closely with two-dimensional plot, and as shown in Figure 3, message transmitted is divided into two kinds of check to variable and variable to check on each limit.Suppose bit ' 0 ' is mapped as+1, bit ' 1 ' is mapped as-1.These two kinds of message are defined as follows:

√ check to variable:R_{Mn}^{a}(a=+1 or a=-1) is check-node z_mTo variable node x_nThe check information that transmits, expression z_mAccording to removing x_nOuter other adjacent variable node $\{ x_k : k \neq n, k \in N(m) \}$ current state is to x_n" the x that declares $_n$=a makes z_mSatisfy " confidence level;

√ variable to check:Q_{Mn}^{a}(a=+1 or a=-1) is variable node x_nTo check-node z_mThe variable information that transmits, expression x_nAccording to removing z_mOuter other adjacent check-node $\{ z_l : l \neq m, l \in M(n) \}$ and initial probability messages P_n^{a}To z_m" the x that declares $_n$=a " confidence level.

Process based on the iterative algorithm of " reliability propagation " is as follows:

1. initialization: calculate initial message $\{ P_n^{a}$And initialization $\{ R_{Mn}^{a}$Value;

2. message is transmitted iteration and renewal:

Stage 1: each P_nAnd z_mNode is all adjacent variable node x to it $_n$Difference pass-along message P_n^aAnd R_{Mn}^a, and each x_nNode is according to receiving information updating Q_{Mn}^a

Stage 2: each x_nNode is all adjacent check-node z to it $_m$Transmit the message Q that has upgraded $_{Mn}^a$, and each z_mNode is according to receiving information updating R_{Mn}^a

3. iteration stops: x_nNode receives message according to all and calculates final message Q_n^aThe iteration of taking stops strategy:

√ is every to take turns iteration when finishing, and calculates final message and measures judgement according to different message Calculate syndrome

If the √ syndrome is a null vector, then declare successfully termination of iterations; Otherwise the value that once calculates before getting forwards above-mentioned second step to and continues iteration.If iterations reaches predefined maximum times I_{Max}Success yet, then according to final message adjudicate x, termination of iterations, decoding is unsuccessful.

The sum-product algorithm based on above-mentioned " reliability propagation " iteration theorem has been used in LDPC decoding in the present embodiment.LDPC and sum-product algorithm have a kind of natural corresponding, and each iteration of decoding all is once " summation " and the realization of " product " computing.

According to the description of above-mentioned reliability propagation algorithm squadron's message transmission and renewal process, under probability measure, variable message is defined as:

$$Q_{mn}^a = P(x_n = a \mid z_m , \{ z_l : l \in M(n) \setminus m \})$$

The definition verification message is:

$$R_{mn}^a = P(z_m \mid x_n = a) ,$$

If code word priori etc. are general, then initial probability messages P_n^aExpression x_nThe prior information of=a, R_{Mn}^aBe initialized as 1, $a \in +1 ,-1\}$.The concrete steps of sum-product algorithm are as shown in Figure 8:

1. initialization

With x_nLikelihood value compose to P_n^aFor example, under additivity additive white Gaussian channel AWGN, likelihood value is:

P n a = 1 1 + exp (- 2 a y n σ 2)

2. check to variable (from the check-node to the variable node) information:

R mn a = P (z m | x n = a)

= Σ x : x n = a P (z m , x | x n = a)

= Σ x ; x n = a P (z m | x n = a , x) P (x | x n = a)

= Σ x : x n = a P (z m | x) P (x | x n = a)

= Σ x : x n = a P (z m | x) Π k ∈ N (m) \ n P (x k)

= Σ x : x n = a P (z m | x) Π k ∈ N (m) \ n Q mk x k

For example suppose that m check equations is $x_2 * x_3 * x_5 = +1$.Calculate x_2Satisfied the probability R of this check equations at=+ 1 o'clock $_{M2}^{0}$:

R m 2 + 1 = P { x 2 × x 3 × x 5 = + 1 | x 2 = + 1 , x 3 = + 1 , x 5 = + 1 } · Q + 1 m 3 · Q + 1 m 5

+ P { x 2 × x 3 × x 5 = + 1 | x 2 = + 1 , x 3 = + 1 , x 5 = - 1 } &C enterDot; Q + 1 m 3 · Q - 1 m 5

+ P { x 2 × x 3 × x 5 = + 1 | x 2 = + 1 , x 3 = - 1 , x 5 = + 1 } &C enterDot; Q - 1 m 3 · Q + 1 m 5

+ P { x 2 × x 3 × x 5 = + 1 | x 2 = + 1 , x 3 = - 1 , x 5 = - 1 } &Ce nterDot; Q - 1 m 3 · Q - 1 m 5

Conditional probability P{} wherein, if satisfy check equations then be 1, not satisfying is 0.In like manner, also can calculate R_{M2}^{-1}

3. variable to check (from the variable node to the check-node) information:

Q mn a = P (x n = a | y n , { z l : l ∈ M (n) \ m })

= P (x n = a , y n , { z l : l ∈ M (n) \ m }) P (y n , { z l : l &Element ; M (n) \ m })

= P (x n = a | y n) P ({ z l : l ∈ M (n) \ m } | y n , x n = a) P ({ z l : l ∈ M (n) \ m } | y n)

= P (x n = a | y n) Π l ∈ M (n) \ m P (z l | x n = a) P ({ z l : l & Element; M (n) \ m })

= α mn P n a Π l ∈ M (n) \ m R ln a

Wherein, α_{Mn}Be normalized parameter, make Q mn + 1 + Q mn - 1 = 1 .

Obtain posterior probability at last:

Q n a = α mn P n a Π l ∈ M (n) \ m R ln a

4. judgement

If Q n + 1 < Q n - 1 , Judge x_n=+1, X then $_n$=0, judge that transmitting terminal has transmitted ' 0 ';

If Q n + 1 > Q n - 1 , Judge x_n=-1, X then $_n$=1, judge that transmitting terminal has transmitted ' 1 '.

Up to XH^T=0, otherwise forward above-mentioned second step to.

As required, can set iterations, if it is not successfully decoded yet to have passed through the iterations of setting, termination of iterations then, decoding failure.

It is to be noted different decoding algorithms and realization different error correction coding performance to be obtained one of algorithm that LDPC that adopts in the present embodiment and long-pending decoding algorithm just generally adopt at present.

The error correction/encoding method that is used for ground digital television broadcast proposed by the invention has used field programmable device (FPGA) to obtain the hardware realization through after the Computer Simulation.One is adopted the ground digital multimedia TV broad cast emission system of the method for the invention to form as shown in Figure

9.The MPEG TS code stream of input can be a video, audio frequency, figure, multimedia messagess such as data, in order to resist the error code that produces in the transmission course, after the cascade error correction coding of one of the BCH of TS stream process the method for the invention and three kinds of LDPC, give modulator and become digital QPSK/mQAM modulation signal, give the modulation of OFDM OFDM multi-carrier again, after the last and Domain Synchronous PN sequence multiple connection, constitute a time-domain synchronization OFDM (TDS-OFDM) signal, through D/A converter module, be converted to suitable analog signal, the radio frequency module receives this analog signal, and the result after the processing gives transmitting antenna or other signal transmitter.

One is adopted the ground digital multimedia TV broad cast receiving system of the method for the invention to form as shown in figure 10.Antenna or other signal receiver receive modulation signal, give after down conversion module carries out frequency translation, give analog-to-digital conversion and become digital signal, and the reinsertion of carrier, clock etc. pass through the demodulation of OFDM multicarrier synchronously then simultaneously.Digital signal after the OFDM demodulation is given back BCH decoder through behind the LDPC sign indicating number BP iterative decoding proposed by the invention, recovers MPEG TS code stream at last.

For a transmission system; the reliability index of transmission information is BER～SNR; the ground digital multimedia TV broad cast system that adopts the method for the invention 64QAM modulation, 1/9 protection at interval, 0.6 encoder bit rate and 4 situations of iteration are little, actual test obtained BER～the SNR performance as shown in figure 11.As we can see from the figure, the BER=10 that requires usually in video broadcasting $^{-11}$During performance, the needed SNR of system is 14.5dB, functional, and because the LDPC code length that adopts has only 7488 bits, than the LDPC sign indicating number of 64800 bits that European DVB-S2 adopted, LDPC sign indicating number time delay proposed by the invention is little, realizes simple, excellent performance satisfies the performance requirement of ground digital television broadcast to high code check.

In conjunction with the accompanying drawings specific embodiments of the invention are had been described in detail above, but the present invention is not restricted to the foregoing description, under the spirit and scope situation of the claim that does not break away from the application, those skilled in the art can make various modifications or remodeling.

Claims (14)

1, the error correction/encoding method of a kind of employing BCH (762,752) and LDPC (7488,3048) cascaded code is characterized in that, may further comprise the steps:
1) TS of input is transmitted per 2 of stream bag and be divided into one group, length is 3008 bits, i.e. 188 * 8 * 2=3008 bit, and then be further divided into 4 son groups, length of each son group is 752 bits;
2) 752 information bits to 4 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 4 * 762=3048 bit;
3) 3048 bits to 4 BCH (762,752) sign indicating number CBCH output carry out LDPC (7493,3048) coding, promptly predetermined with LDPC (7493,3048) generator matrix G_{Qc1}Multiply each other, obtain QC-LDPC (7493,3048) sign indicating number C_{Qc1}, described generator matrix G_{Qc1}Structure based on predetermined check matrix H_{Qc1}
4) deletion LDPC (7493,3048) sign indicating number C_{Qc1}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.
2, the error correction/encoding method of employing BCH as claimed in claim 1 (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:
1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;
2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+ x^{3}+ 1$, export 1023 error correction coding bits;
3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$
3, the error correction/encoding method of employing BCH as claimed in claim 1 (762,752) and LDPC (7488,3048) cascaded code is characterized in that, listens the check matrix H of the LDPC (7493,3048) that states $_{Qc1}$Constitute by cyclic permutation matrices:
H qc 1 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1
Wherein, γ=35, ρ=59, $B_{i,j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1.

4, the error correction/encoding method of employing BCH as claimed in claim 1 (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) check matrix H $_{Qc1}$Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:

1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal $_{RS}$, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;

2) from C $_{RS}$Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$

3) structure codeword set C $_{b}$ $^{(i)}$: c is C $_{b}$In the code word that code weight is ρ, C $_{b}$Middle q-1 non-zero codeword constitutes the set C of one dimension code word $_{b}$ $^{(1)}$={ β c: β ∈ GF (p ^{s}), based on subcode C $_{b}$ $^{(1)}$, by one of ring shift right, C $_{b}$Be divided into q-1 coset, i.e. C $_{b}$ $^{(1)}$, C $_{b}$ $^{(2)}$, L C $_{b}$ $^{(q-1)}$, so any coset C $_{b}$ $^{(i)}$In two code words all be different in all positions;

4) by C $_{b}$ $^{(i)}$Structure character position set of vectors Z (C $_{b}$ $^{(i)}$): z=(z $_{0}$, z $_{1}$, L, z $_{Q-2}$) be (q-1)-heavily on the GF (2), corresponding to GF (p ^{s}) the field element α of q-1 non-zero ^{i}, and b=(b $_{0}$, b $_{1}$, L, b $_{ρ-1}$) be C $_{b}$In code word, with z (b $_{j}$) replace each component b of b $_{j}$, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_{0}$), z (b $_{1}$) ... .., z (b $_{ρ-1}$)), C so $_{b}$ $^{(1)}$I coset C $_{b}$ $^{(i)}$In the set of character position vector of q-1 code word be Z (C $_{b}$ $^{(i)}$)={ z (b): b ∈ C $_{b}$ $^{(i)}$, Z (C $_{b}$ $^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";

5) structure Z (C $_{b}$ $^{(i)}$) the character position matrix A $_{i}$: for 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF (2) $_{i}$, its row is Z (C $_{b}$ $^{(i)}$) in q-1 character position vector;

6) by the character position matrix A $_{i}$The basic matrix H of a rule of structure $_{GA}$(γ, ρ):

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1

Its row, column weight respectively is ρ and γ, and it has its member's matrix A $_{i}$Design feature;

7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);

8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc1}$:

H qc 1 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

Wherein

5, the error correction/encoding method of employing BCH as claimed in claim 1 (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) generator matrix G_{Qc1}Constitute by cyclic permutation matrices:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{I, j}$Be (q-1) * (q-1) circular matrix, q=128, γ=35, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1.

6, the error correction/encoding method of employing BCH as claimed in claim 1 (762,752) and LDPC (7488,3048) cascaded code is characterized in that, described LDPC (7493,3048) generator matrix G_{Qc1}Structure based on described check matrix H_{Qc1}, step is as follows:

1) H_{Qc}The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1

D is square formation, full rank, nonsingular, i.e. inverse matrix D^{-1}There are its order and H_{Qc}Order identical;

2) according to coding theory $H_{Qc}G_{Qc}^{T}$=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation M i u T + Dz i T = 0 , Then

z i T = D - 1 M i u T

Wherein

z_i=($g_{I, 0}g_{I, 1}$K $g_{I, γ-1}$) expression G_{Qc}The i of P matrix is capable in the generator matrix

M_i=[$B^{T}_{0, i}$... ..$B^{T}_{γ-1, i}$] ^{T}Expression H_{Qc}The i row of circular matrix;

U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;

Thereby obtain G_{Qc}Middle circular matrix $G_{I, j}$All generator polynomial $g_{I, j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

3) when the non-full rank situation of matrix D, to G_{QC}The correction of structure:
A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;
B) by z i T = D - 1 M i u T Try to achieve z_iMiddle r linearity be element independently;
C) at G_{QC}Middle γ (the q-1)-r linearity of adding is independently gone g_0^*, g_1^*… .., $g^*_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{QC}Expression formula:
G * qc = G - Q = IOLO G 0,0 G 0,1 L G 0 , γ - 1 OILO G 1,0 G 1,1 L G 1 , γ - 1 MMOMMM G i , j M OOLI G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1
And g_i^*Obtain by finding the solution following matrix equality:
B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g i i , ρ (q - 1) - 1 = 0 .
7, the error correction/encoding method of a kind of employing BCH (762,752) and LDPC (7488,4572) cascaded code is characterized in that, may further comprise the steps:
1) TS of input is transmitted per 3 of stream bag and be divided into one group, length is 4512 bits, i.e. 188 * 8 * 3=4512 bit, and then be further divided into 6 son groups, length of each son group is 752 bits;
2) 752 information bits to 6 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 6 * 762=4572 bit;
3) to 6 BCH (762,752) sign indicating number C_{BCH}4572 bits of output carry out LDPC (7493,4572) coding, promptly with the generator matrix G of LDPC (7493,4572) $_{Qc2}$Multiply each other, obtain QC-LDPC (7493,4572) sign indicating number C_{Qc2}, described generator matrix G_{Qc2}Structure based on predetermined check matrix H_{Qc2}
4) deletion LDPC (7493,4572) sign indicating number C_{Qc2}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.
8, the error correction/encoding method of employing BCH as claimed in claim 7 (762,752) and LDPC (7488,4572) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:

1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;
2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+x^{3}+1$, export 1023 error correction coding bits;
3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$
9, the error correction/encoding method of employing BCH as claimed in claim 7 (762,752) and LDPC (7488,4572) cascaded code is characterized in that the check matrix H of described LDPC (7493,4572) $_{Qc2}$Constitute by cyclic permutation matrices:
H qc 2 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1
Wherein, γ=23, ρ=59, $B_{I,j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1, described check matrix H $_{Qc2}$Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:
1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal $_{RS}$, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;
2) from C $_{RS}$Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$
3) structure codeword set $C_b^{(i)}$: c is C_bIn the code word that code weight is ρ, C_bMiddle q-1 non-zero codeword constitutes the set C of one dimension code word $_{b}^{(1)}$={ β c: β ∈ GF (p^{s}), based on subcode $C_b^{(1)}$, by one of ring shift right, C_bBe divided into q-1 coset, i.e. $C_b^{(1)}$, $C_b^{(2)}$, L $C_b^{(q-1)}$, so any coset $C_b^{(i)}$In two code words all be different in all positions;
4) by $C_b^{(i)}$Structure character position set of vectors Z ($C_b^{(i)}$): z=(z_0, z_1, L, z_{Q-2}) be (q-1)-heavily on the GF (2), corresponding to GF (p^{s}) the field element α of q-1 non-zero ^{i}, and b=(b_0, b_1, L, $b_{\rho-1}$) be C_bIn code word, with z (b_j) replace each component b of b_j, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_{0}$), z (b_1) ... .., z ($b_{\rho-1}$)), C so $_{b}^{(1)}$I coset $C_b^{(i)}$In the set of character position vector of q-1 code word be Z ($C_b^{(i)}$)={ z (b): b ∈ $C_b^{(i)}$, Z ($C_b^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";
5) structure Z ($C_b^{(i)}$) the character position matrix A_i: for 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF (2) $_{i}$, its row is Z ($C_b^{(i)}$) in q-1 character position vector;

6) by the character position matrix A $_i$The basic matrix H of a rule of structure $_{GA}$(γ, ρ):
H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1
Its row, column weight respectively is ρ and γ, and it has its member's matrix A $_i$Design feature;
7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);
8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc2}$:
H qc 2 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ
Wherein
10, the error correction/encoding method of employing BCH as claimed in claim 7 (762,752) and LDPC (7488,4572) cascaded code is characterized in that, described LDPC (7493,4572) generator matrix G $_{Qc2}$Constitute by cyclic permutation matrices:
G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1
Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and G $_{I, j}$Be (q-1) * (q-1) circular matrix, q=128, γ=23, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1 Described generator matrix G $_{Qc2}$Structure based on described check matrix H $_{Qc2}$, step is as follows:
1) H $_{Qc}$The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:
D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1
D is square formation, full rank, nonsingular, i.e. inverse matrix D $^{-1}$There are its order and H $_{Qc}$Order identical;
2) according to coding theory H $_{Qc}$G $_{Qc}$ ^{T}=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation M i u T + Dz i T = 0 , Then
z i T = D - 1 M i u T

Wherein
$z_i=(g_{i,0}\ g_{l,1}K\ g_{l,\gamma-1})$ expression G_{Qc}The i of P matrix is capable in the generator matrix
$M_i=[B^T_{0,i}\ldots\ldots B^T_{\gamma-1,i}]^T$Expression H_{Qc}The i row of circular matrix;
U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;
Thereby obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$
3) when the non-full rank situation of matrix D, to G_{Qc}The correction of structure:
A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;
B) by z i T = D - 1 M i u T Try to achieve z_iMiddle r linearity be element independently;
C) at G_{Qc}Middle γ (the q-1)-r linearity of adding is independently gone g_0^*, $g_1^*\ldots\ldots, g^*_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{Qc}Expression formula:
G * qc = G - Q = IOLO G 0,0 G 0,1 L G 0 , γ - 1 OILO G 1,0 G 1,1 L G 1 , γ - 1 MMOMMM G i , j M OOLI G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1
And g_i^*Obtain by finding the solution following matrix equality:
B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g i i , ρ (q - 1) - 1 = 0 .
11, the error correction/encoding method of a kind of employing BCH (762,752) and LDPC (7488,6096) cascaded code is characterized in that, may further comprise the steps:
1) TS of input is transmitted per 4 of stream bag and be divided into one group, length is 6016 bits, i.e. 188 * 8 * 4=6016 bit, and then be further divided into 8 son groups, length of each son group is 752 bits;
2) 752 information bits to 8 son groups carry out binary system BCH (762,752) error correction coding respectively, and the total output length that obtains after the error correction coding is 8 * 762=6096 bit;
3) to 8 BCH (762,752) sign indicating number C_{BCH}6096 bits of output carry out LDPC (7493,6096) coding, promptly with the generator matrix G of LDPC (7493,6096) $_{Qc3}$Multiply each other, obtain QC-LDPC (7493,6096) sign

indicating number C_{Qc3}, described generator matrix G_{Qc3}Structure based on predetermined check matrix H_{Qc3}

4) deletion LDPC (7493,6096) sign indicating number C_{Qc3}Preceding 5 check bits of 7493 bits of output obtain 7488 output error correction coding bits.

12, the error correction/encoding method of employing BCH as claimed in claim 11 (762,752) and LDPC (7488,6096) cascaded code is characterized in that, described BCH (762,752) is the shortening sign indicating number of BCH (1023,1013), generates according to the following steps:

1) before 752 data bits, increases by 261 " 0 " bits, form the information of 1013 bits;

2) these 1013 information bits are carried out BCH (1023,1013) error correction coding, its generator polynomial is $x^{10}+ x^{3}+ 1$, export 1023 error correction coding bits;

3) remove preceding 261 bits of error correction coding bit, because BCH code is a systematic code, 261 bits that remove are " 0 " bits, obtain BCH (762,752) sign indicating number C at last $_{BCH}$

13, the error correction/encoding method of employing BCH as claimed in claim 11 (762,752) and LDPC (7488,6096) cascaded code is characterized in that the check matrix H of described LDPC (7493,6096) $_{Qc3}$Constitute by cyclic permutation matrices:

H qc 3 = B 0,0 B 0,1 L B 0 , ρ - 1 B 1,0 B 1,1 L B 1 , ρ - 1 M M B i , j M B γ - 1,0 B γ - 1,1 L B γ - 1 , ρ - 1

Wherein, γ=11, ρ=59, $B_{I, j}$Be (q-1) * (q-1) rank circular matrix, q=128,0≤i≤γ-1 and 0≤j≤ρ-1, described check matrix H_{Qc3}Structure based on the Reed Solomon code of two information symbols, i.e. RS sign indicating number, step is as follows:

1) patrols generation RS (q-1, q-ρ, ρ-1) sign indicating number C among the magnificent territory GF (q) at gal_{RS}, wherein code length is q-1, and information symbol length is q-ρ-1, and smallest hamming distance is ρ-1;

2) from C_{RS}Q-ρ-1 information symbol obtains shortening RS (ρ, 2, ρ-1) sign indicating number C before the middle deletion $_{b}$

3) structure codeword set $C_{b}^{(i)}$: c is C_{b}In the code word that code weight is ρ, C_{b}Middle q-1 non-zero codeword constitutes the set C of one dimension code word $_{b}^{(1)}$={ β c: β ∈ GF (p^{s}), based on subcode $C_{b}^{(1)}$, by one of ring shift right, C_{b}Be divided into q-1 coset, i.e. $C_{b}^{(1)}$, $C_{b}^{(2)}$, L $C_{b}^{(q-1)}$, so any coset $C_{b}^{(i)}$In two code words all be different in all positions;

4) by $C_{b}^{(i)}$Structure character position set of vectors Z ($C_{b}^{(i)}$): z=(z_{0}, z_{1}, L, z_{Q-2}) be (q-1)-heavily on the GF (2), corresponding to GF (p^{s}) the field element α of q-1 non-zero ^{i}, and b=(b_{0}, b_{1}, L, $b_{\rho-1}$) be C_{b}In code word, with

z (b_j) replace each component b of b_j, then obtain ρ (q-1)-weight: z (b)=(z (b on the GF (2) $_0$), z (b_1) ... .., z ($b_{ρ-1}$)), C so $b^{(1)}$I coset $C_b^{(i)}$In the set of character position vector of q-1 code word be Z ($C_b^{(i)}$)={ z (b): b ∈ $C_b^{(i)}$, Z ($C_b^{(i)}$) in any two character position vectors be that have or identical without any " 1-component ";

5) structure Z ($C_b^{(i)}$) the character position matrix A_i: for 0≤i≤q-2, go up structure (q-1) * ρ (q-1) rank matrix A at GF (2) $_i$, its row is Z ($C_b^{(i)}$) in q-1 character position vector;

6) by the character position matrix A_iThe basic matrix H of a rule of structure $_{GA}$(γ, ρ):

H GA (γ , ρ) = A 0 M A γ - 1 = A 0,0 A 0,1 L A 0 , ρ - 1 A 1,0 A 1,1 L A 1 , ρ - 1 M M A i , j M A γ - 1,0 A γ - 1,1 L A γ - 1 , ρ - 1

Its row, column weight respectively is ρ and γ, and it has its member's matrix A_iDesign feature;

7) (γ ρ), makes its row, row distribution of weight equal to calculate in advance definite variable node degree distribution v (X) and check-node degree distribution c (X) respectively to go up the γ * ρ rank masking matrix W of non-rule by a GF of computer random structure (2);

8) obtain the check matrix H of QC-LDPC by following matrix multiplication $_{Qc3}$:

H qc 3 = w 1,1 A 1,1 w 1,2 A 1,2 L w 1 , ρ A 1 , ρ w 2,1 A 2,1 w 2,2 A 2,2 L w 2 , ρ A 2 , ρ M M w i , j A i , j M w γ , 1 A γ , 1 w γ , 2 A γ , 2 L w γ , ρ A γ , ρ

Wherein

14, the error correction/encoding method of employing BCH as claimed in claim 11 (762,752) and LDPC (7488,6096) cascaded code is characterized in that, described LDPC (7493,6096) generator matrix G_{Qc3}Constitute by cyclic permutation matrices:

G qc = I O L O G 0,0 G 0,1 L G 0 , γ - 1 O I L O G 1,0 G 1,1 L G 1 , γ - 1 M M O M M M G i , j M O O L I G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1

Wherein, I is (q-1) * (q-1) rank unit matrixs, and O is (q-1) * (q-1) rank zero battle arrays, and $G_{I, j}$Be (q-1) * (q-1) circular matrix, q=128, γ=11, ρ=59,0≤i≤ρ-γ-1,0≤j≤γ-1, described generator matrix G_{Qc3}Structure based on described check matrix H_{Qc3}, step is as follows:

1) H_{Qc}The column weight of circular matrix newly arrange, construct a γ * ρ rank matrix D:

D = B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B & gamma; - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B & gamma; - 1 , ρ - 1

D is square formation, full rank, nonsingular, i.e. inverse matrix D^{-1}There are its order and H_{Qc}Order identical;

2) according to coding theory $H_{Qc}G_{Qc}^{T}$=| O|, | O| be γ (q-1) * ((q-1) rank of ρ-γ) zero battle array, equation M i u T + Dz i T = 0 , Then

z i T = D - 1 M i u T

Wherein

z_i=($g_{l,0}g_{l,1}$K $g_{l,\gamma-1}$) expression G_{Qc}The i of P matrix is capable in the generator matrix

M_i=[$B^{T}_{0,i}$... ..$B^{T}_{\gamma-1,i}$] ^{T}Expression H_{Qc}The i row of circular matrix;

U=(1 0 K 0) expression " 1 " is positioned at unit (the q-1)-weight of beginning;

Thereby obtain G_{Qc}Middle circular matrix $G_{l,j}$All generator polynomial $g_{l,j}$, promptly obtained the generator matrix G of the QC-LDPC sign indicating number under the matrix D full rank situation $_{Qc}$

3) when the non-full rank situation of matrix D, to G_{Qc}The correction of structure:

A) z_iIn be made as 0 corresponding to the element of D matrix neutral line related column, its quantity is γ (q-1)-r;

B) by z i T = D - 1 M i u T Try to achieve z_iMiddle r linearity be element independently;

C) at G_{Qc}Middle γ (the q-1)-r linearity of adding is independently gone g_0^{*}, g_1^{*}... .., $g^{*}_{\gamma(q-1)-r-1}$, to constitute complete generator matrix G_{Qc}Expression formula:

G * qc = G - Q = IOLO G 0,0 G 0,1 L G 0 , γ - 1 OILO G 1,0 G 1,1 L G 1 , γ - 1 MMOMMM G i , j M OOLI G ρ - γ - 1,0 G ρ - γ - 1,1 L G ρ - γ - 1 , γ - 1 - g * 0 g * 1 M g * γ (q - 1) - r - 1

And g_i^{*}Obtain by finding the solution following matrix equality:

B 0 , ρ - γ B 0 , ρ - γ + 1 L B 0 , ρ - 1 B 1 , ρ - γ B 1 , ρ - γ + 1 L B 1 , ρ - 1 M M L M B γ - 1 , ρ - γ B γ - 1 , ρ - γ + 1 L B γ - 1 , ρ - 1 g * i , (ρ - γ) (q - 1) g * i , (ρ - γ) (q - 1) + 1 M g i i , ρ (q - 1) - 1 = 0 .

Cited By (14)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations

CN101087180B *2006-06-082012-05-23华为技术有限公司Decoding method, device and application of wireless channel

CN101350625B *2007-07-182011-08-31北京泰美世纪科技有限公司High-efficiency all-purpose decoder for QC-LDPC code and decoding method thereof

KR101503059B1 *2008-02-262015-03-19삼성전자주식회사Apparatus and method for channel encoding and decoding in communication system using low-density parity-check codes

CN101414868B *2008-12-032012-08-22中国航天科技集团公司第五研究院第五〇四研究所CDAS station DCPR communication protocol and receiving method

WO2010087546A1 *2009-01-292010-08-05Lg Electronics Inc.Apparatus for transmitting and receiving a signal and method of transmitting and receiving a signal

CN102301704B *2009-02-062014-04-23Lg电子株式会社Apparatus for transmitting and receiving a signal and method of transmitting and receiving a signal

JP2011135456A *2009-12-252011-07-07Sony CorpReceiver, receiving method, program and receiving system

CN102821071B *2012-08-242014-12-03电子科技大学Signal channel and noise variance joint estimation method of OFDM (orthogonal frequency division multiplexing) system

CN104518847B *2013-09-292018-02-02中国科学院上海高等研究院Signalling coding method and system based on BCH code and the cascade of short LDPC code

CN104734807B *2015-02-132018-03-02中国科学院上海高等研究院A kind of wireless broadcast communication system and method

WO2017045142A1 *2015-09-162017-03-23华为技术有限公司Decoding method and decoding device for ldpc truncated code

WO2017061892A1 *2015-10-092017-04-13Huawei Technologies Co., Ltd.Encoding and decoding of generalized concatenated codes with inner piggybacked codes for distributed storage systems

WO2017078562A1 *2015-11-022017-05-11Huawei Technologies Co., Ltd.Method and device for encoding/decoding data by using m-th order gel codes
CN107026654B *2016-02-022019-06-18中国科学院声学研究所A kind of fast frequency-domain coding method of quasi- circulation q-ary low-density parity-check
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

Publication	Publication Date	Title
Kim et al.	2016	Low-density parity-check codes for ATSC 3.0
EP2660983B1	2018-01-24	Encoding and decoding for rate 4/15, 7/15 and 8/15 16k ldpc codes
JP5664919B2	2015-02-04	Data processing apparatus and data processing method
KR20160061328A	2016-05-31	Data processing device and data processing method
RU2595585C2	2016-08-27	Data processing apparatus and data processing method
US9071276B2	2015-06-30	Data-processing device and data-processing method
Liva et al.	2006	Design of LDPC codes: A survey and new results
EP2056464B1	2012-12-12	Data processing apparatus and method
US7383487B2	2008-06-03	IPHD (iterative parallel hybrid decoding) of various MLC (multi-level code) signals
JP5601182B2	2014-10-08	Data processing apparatus and data processing method
Kang et al.	2010	Quasi-cyclic LDPC codes: An algebraic construction
JP2015170911A	2015-09-28	Data processor and data processing method
EP2675068B1	2017-05-03	Column-twist interleaving for ldpc codes
US20160164540A1	2016-06-09	Data processing device and data processing method
KR101197073B1	2012-11-06	Communication signal decoding with iterative cooperation between inner and outer codes

EP1553706B12008-06-25Bandwidth efficient LDPC (low density parity check) coded modulation scheme based on MLC (multi-level code) signals
US9184768B22015-11-10Data-processing device and data-processing method
US20160211868A12016-07-21Data processing device and data processing method
CN104221292B2019-07-02Data processing equipment and data processing method
US9203435B22015-12-01Multiple size and rate FEC code combination with minimum shortening and maximum combined code rate
Zeng et al.2008Construction of nonbinary cyclic, quasi-cyclic and regular LDPC codes: a finite geometry approach
US9621191B22017-04-11Data processing device and data processing method
KR102023558B12019-09-23Data processing device, and data processing method
Park et al.2014Raptor-like rate compatible LDPC codes and their puncturing performance for the cloud transmission system
US9432143B22016-08-30Combining CRC and FEC on a variable number of NCPs

Priority And Related Applications

Priority Applications (1)

ApplicationPriority dateFiling dateTitle
CN 200510086336 2005-09-02 2005-09-02 Correction coding method for ground digital television broadcast

Applications Claiming Priority (1)

ApplicationFiling dateTitle
CN 200510086336 2005-09-02 Correction coding method for ground digital television broadcast

Legal Events

DateCodeTitleDescription
2007-03-07PB01Publication

2007-03-07C06Publication
2007-05-02SE01Entry into force of request for substantive examination
2007-05-02C10Entry into substantive examination
2010-01-20GR01Patent grant
2010-01-20C14Grant of patent or utility model

[19] 中华人民共和国国家知识产权局

[51] Int. Cl.
H04N 7/26 (2006.01)

[12] 发明专利说明书

专利号 ZL 200510086336.5

[45] 授权公告日 2010 年 1 月 20 日

[11] 授权公告号 CN 100584011C

[22] 申请日 2005.9.2

[21] 申请号 200510086336.5

[73] 专利权人 清华大学

地址 100084 北京市北京 100084 - 82 信箱

[72] 发明人 杨 林 杨知行 张晓林 门爱东

[56] 参考文献

CN1481130 A 2004.3.10

CN1547806 A 2004.11.17

CN1625058 A 2005.6.8

审查员 魏 玮

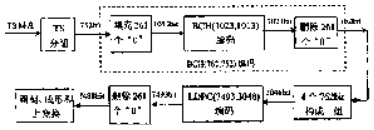
权利要求书 11 页 说明书 40 页 附图 5 页

[54] 发明名称

用于地面数字电视广播的纠错编码方法

[57] 摘要

本发明涉及地面数字电视广播错编码方法，属于数字信息传输技术领域。本发明方法包括：将输入传递流分为 752 比特子组，在每子组前加 261 个“0”，得 1013 比特，进行 BCH(1023，1013)编码，去掉编码后的前 261 个比特得 BCH(762，752)；将一定数目的 BCH(762，752)构成一组，按给定的生成矩阵进行 LDPC(7493，3048)、LDPC(7493，4572)或 LDPC(7493，6096)编码，删除输出的 7493 个比特的前 5 个校验比特，得到最后的 7488 个输出比特。本发明采用了较短的码长，实现简单，具有较小的时延和良好的纠错性能，更好地适应各种应用。



1、一种采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传递流包每 2 个分为一组，长度为 3008 比特，即 188×8×2=3008 比特，然后再进一步分为 4 个子组，每一子组的长度为 752 比特；

2）对 4 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的总的输出长度为 4×762=3048 比特；

3） 对 4 个 BCH(762,752) 码 C_{BCH} 输出的 3048 个比特进行 LDPC(7493,3048) 编码，即与 LDPC(7493,3048) 预定的生成矩阵 G_{qc1} 相乘，得到 QC-LDPC (7493,3048) 码 C_{qc1}，所述生成矩阵 G_{qc1} 的构造基于预定的校验矩阵 H_{qc1}；

4）删除 LDPC(7493,3048) 码 C_{qc1} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出纠错编码比特。

2、如权利要求1所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^3+1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH 码是系统码，去掉的 261 个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

3、如权利要求 1 所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 的校验矩阵 H_{qc1} 由循环置换矩阵构成：

$$H_{qc1}=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，$\gamma=35$，$\rho=59$，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$ 阶循环矩阵，q=128，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$。

4、如权利要求 1 所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 校验矩阵 H_{qc1} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽逻华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移一位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，对应于 $GF(p^s)$ 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，用 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0\le i\le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix}A_0\\ \mathrm{M}\\ A_{\gamma-1}\end{bmatrix}=\begin{bmatrix}A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1}\\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M}\\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1}\end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 γ × ρ 阶掩蔽矩阵 W(γ, ρ)，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc1}：

$$H_{qc1}=\begin{bmatrix}w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho}\\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho}\\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M}\\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho}\end{bmatrix}$$

其中

$$w_{i,j}A_{i,j}=\begin{cases}A_{i,j}, & \text{对于 } w_{i,j}=1\\ 0, & \text{对于 } w_{i,j}=0\end{cases}。$$

5、如权利要求 1 所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 生成矩阵 G_{qc1} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=35，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。

6、如权利要求 1 所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 生成矩阵 G_{qc1} 的构造基于所述的校验矩阵 H_{qc1}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma\times\rho$ 阶矩阵 D：

$$D=\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T}=|O|$，|O| 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_i u^T+Dz_i^T=0$，则

$$z_i^T=D^{-1}M_i u^T$$

其中

$z_i=(g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\rho-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_i=[B^T{}_{0,i},\ldots\ldots B^T{}_{\gamma-1,i}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u=(1\ \ 0\ \mathrm{K}\ \ 0)$ 表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b) 由 $z_i^T=D^{-1}M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

c) 在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*,g_1^*\ldots\ldots,g^*_{\gamma(q-1)-r-1}$，以构成完整的生成矩阵 G_{qc} 表达式：

$$G^*_{qc}=\begin{bmatrix}G\\ \cdots\\ Q\end{bmatrix}=\begin{bmatrix}I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1}\\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M}\\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1}\\ \hline & & & & g^*_0 & & & \\ & & & & g^*_1 & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-r-1} & & & \end{bmatrix}$$

而 g_i^* 由求解下列矩阵等式获得：

$$\begin{bmatrix}B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1}\\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M}\\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1}\end{bmatrix}\begin{bmatrix}g^*_{i,(\rho-\gamma)(q-1)}\\ g^*_{i,(\rho-\gamma)(q-1)+1}\\ \mathrm{M}\\ g^*_{i,\rho(q-1)-1}\end{bmatrix}=0\ 。$$

7、一种采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传递流包每 3 个分为一组，长度为 4512 比特，即 188×8×3=4512 比特，然后再进一步分为 6 个子组，每一子组的长度为 752 比特；

2）对 6 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的总的输出长度为 6×762=4572 比特；

3）对 6 个 BCH(762,752) 码 C_{BCH} 输出的 4572 个比特进行 LDPC(7493,4572) 编码，即与 LDPC(7493, 4572) 的生成矩阵 G_{qc2} 相乘，得到 QC-LDPC (7493, 4572) 码 C_{qc2}，所述生成矩阵 G_{qc2} 的构造基于预定的校验矩阵 H_{qc2}；

4）删除 LDPC(7493,4572)码 C_{qc2} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出纠错编码比特。

8、如权利要求 7 所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^3-1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH码是系统码，去掉的261个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

9、如权利要求 7 所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,4572) 的校验矩阵 H_{qc2} 由循环置换矩阵构成：

$$H_{qc2}=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，γ=23，ρ=59，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$ 阶循环矩阵，q=128，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$，所述的校验矩阵 H_{qc2} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽逻华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移 一位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，对应于 $GF(p^s)$ 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，用 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0\le i\le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix} A_0 \\ \mathrm{M} \\ A_{\gamma-1}^{\rho} \end{bmatrix}=\begin{bmatrix} A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1} \\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M} \\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1} \end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 γ × ρ 阶掩蔽矩阵 W(γ, ρ)，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc2}：

$$H_{qc2}=\begin{bmatrix} w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho} \\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho} \\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M} \\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho} \end{bmatrix}$$

其中

$$w_{i,j}A_{i,j}=\begin{cases} A_{i,j}, & \text{对于 } w_{i,j}=1 \\ 0, & \text{对于 } w_{i,j}=0 \end{cases}。$$

10、如权利要求 7 所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,4572) 生成矩阵 G_{qc2} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=23，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。所述的生成矩阵 G_{qc2} 的构造基于所述的校验矩阵 H_{qc2}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma\times\rho$ 阶矩阵 D：

$$D=\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T}=|O|$，|O| 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_1u^{T}+Dz_1^{T}=0$，则

$z_i^T = D^{-1} M_i u^T$

其中

$z_i = (g_{i,0}\ g_{i,1}\ \mathrm{K}\ g_{i,\gamma-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_i = [B^T_{0,i} \ldots\ldots B^T_{\gamma-1,i}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u = (1\ \ 0\ \mathrm{K}\ \ 0)$ 表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b）由 $z_i^T = D^{-1} M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

c）在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_c^*, g_1^* \ldots\ldots, g^*_{\gamma(q-1)-r-1}$，以构成完整的生成矩阵 G_{qc} 表达式：

$$G^*_{qc} = \begin{bmatrix} G \\ - \\ Q \end{bmatrix} = \begin{bmatrix} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \\ \hline & & & & g^*_0 & & & \\ & & & & g^*_1 & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-r-1} & & & \end{bmatrix}$$

而 g_i^* 由求解下列矩阵等式获得：

$$\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix} \begin{bmatrix} g^*_{i,(\rho-\gamma)(q-1)} \\ g^*_{i,(\rho-\gamma)(q-1)+1} \\ \mathrm{M} \\ g^*_{i,\rho(q-1)-1} \end{bmatrix} = 0 。$$

11、一种采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传递流包每 4 个分为一组，长度为 6016 比特，即 188×8×4=6016 比特，然后再进一步分为 8 个子组，每一子组的长度为 752 比特；

2）对 8 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错

编码后的总的输出长度为 8×762=6096 比特；

3）对 8 个 BCH(762,752) 码 C_{BCH} 输出的 6096 个比特进行 LDPC(7493,6096) 编码，即与 LDPC(7493, 6096) 的生成矩阵 G_{qc3} 相乘，得到 QC-LDPC (7493, 6096) 码 C_{qc3}，所述生成矩阵 G_{qc3} 的构造基于预定的校验矩阵 H_{qc3}；

4）删除 LDPC(7493,6096) 码 C_{qc3} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出纠错编码比特。

12、如权利要求11所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^3+1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH码是系统码，去掉的261个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

13、如权利要求 11 所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,6096) 的校验矩阵 H_{qc3} 由循环置换矩阵构成：

$$H_{qc3}=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，$\gamma=11$，$\rho=59$，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$ 阶循环矩阵，q=128，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$，所述的校验矩阵 H_{qc3} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽罗华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移 i 位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，

对应于 GF(p^s) 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L},b_{\rho-1})$ 为 C_b 中的码字，用 z(b_j) 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0\le i\le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix}A_0\\ \mathrm{M}\\ A_{\gamma-1}\end{bmatrix}=\begin{bmatrix}A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1}\\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M}\\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1}\end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 $\gamma\times\rho$ 阶掩蔽矩阵 W(γ，ρ)，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc3}：

$$H_{qc3}=\begin{bmatrix}w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho}\\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho}\\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M}\\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho}\end{bmatrix}$$

其中

$$w_{i,j}A_{i,j}=\begin{cases}A_{i,j}, & \text{对于 } w_{i,j}=1\\ 0, & \text{对于 } w_{i,j}=0\end{cases}。$$

14、如权利要求 11 所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,6096) 生成矩阵 G_{qc3} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix}I & C & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1}\\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M}\\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1}\end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=11，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$，所述的

生成矩阵 G_{qc3} 的构造基于所述的校验矩阵 H_{qc3}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma \times \rho$ 阶矩阵 D：

$$D = \begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T} = |O|$，$|O|$ 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_t u^T + Dz_t^T = 0$，则

$$z_t^T = D^{-1}M_t u^T$$

其中

$z_t = (g_{t,0}\ g_{t,1}\mathrm{K}\ g_{t,\gamma-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_t = [B^T_{0,t}.....B^T_{\gamma-1,t}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u = (1\ \ 0\ \mathrm{K}\ \ 0)$ 表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_t 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b) 由 $z_t^T = D^{-1}M_t u^T$ 求得 z_t 中 r 个线性独立的元素；

c) 在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*, g_1^*.....g^*_{\gamma(q-1)-r-1}$，以构成完整的生成矩阵 G_{qc} 表达式：

$$G^*_{qc} = \begin{bmatrix} G \\ - \\ Q \end{bmatrix} = \left[\begin{array}{cccccccc} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\rho-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \\ \hline & & & & g_0^* & & & \\ & & & & g_1^* & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-r-1} & & & \end{array}\right]$$

而 g_i^* 由求解下列矩阵等式获得：

$$\begin{bmatrix} B_{0,p-\gamma} & B_{0,p-\gamma+1} & \mathrm{L} & B_{0,p-1} \\ B_{1,p-\gamma} & B_{1,p-\gamma+1} & \mathrm{L} & B_{1,p-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,p-\gamma} & B_{\gamma-1,p-\gamma+1} & \mathrm{L} & B_{\gamma-1,p-1} \end{bmatrix} \begin{bmatrix} g^{+}_{1,(p-\gamma)(q-1)} \\ g^{+}_{1,(p-\gamma)(q-1)+1} \\ \mathrm{M} \\ g^{+}_{1,p(q-1)-1} \end{bmatrix} = 0 \circ$$

用于地面数字电视广播的纠错编码方法

技术领域

本发明属于数字信息传输技术领域，特别涉及一种用于地面数字多媒体电视广播的纠错编码方法。

背景技术

在经历了机械电视时代、电子黑白电视和彩色电视时代以后，日本广播公司 NHK 从 70 年代初期开始研究高清晰度电视，后来有许多国家和国际标准组织也积极开展了这方面的研发。但直到 1990 年之前，所采用的技术方案还都是全模拟的或者模拟数字混合的方案。在 1990 年 6 月 1 日，美国 GI 公司向美国高级电视顾问委员会（ACATS）提交了全数字地面 HDTV 制式，电视从此进入了一个新时代——数字电视时代。

经过多年坚持不懈的研究和发展，许多国家和地区已经制定了地面数字电视广播（Digital Television Terrestrial Broadcasting, DTTB）标准，主要有三个国际地面数字电视标准（美国 ATSC、欧洲 DVB-T 和日本 ISDB-T）。在国内，清华大学也提出了自己的地面数字多媒体电视广播（DMB-T）传输方案。这些标准和方案在系统构成、设计参数和目标等许多方面，或多或少地存在着差异，存在着各种各样的不同要求。

不同的地面数字电视广播系统的业务要求。在地面电视广播数字化的过程中，需要为用户提供不同的业务：

✓例如提供包括 HDTV 在内的多媒体广播和综合数据服务业务；

✓支持固定、便携、手持等接收设备；

✓高度灵活的频率规划和广播覆盖区域。

不同的设计参数和环境要求。根据我国地面广播电视的具体情况，地面数字多媒体电视广播传输系统面临的的设计要求有：

✓射频范围： 小于 900 兆赫兹

✓基带频率带宽： 8 兆赫兹

✓移动接收机的时速： 小于 每小时 130 公里

✓单发射机覆盖范围： 半径小于 75 公里

✓适用地区： 都市、郊区、城镇

✓适用地貌： 平原、丘陵、沿海、湖边

✓使用环境： 室内、室外、车载、便携

能够处理不同的信道模型要求。这些地面数字多媒体电视广播传输系统的信道特征有：

✓频率选择性：时延分布(Delay spread)约为 50 微秒

✓时间选择性：多普勒分布(Doppler spread) 约为 100 赫兹

✓信道模型：广义平稳并非相关散射(WSSUS)

✓噪声模型：加性高斯白噪声(AWGN)、脉冲式干扰、单频干扰等

采用不同的调制和纠错方式，适应多种码率要求：

✓码率：5Mb/s～30Mb/s，适应从移动到固定的不同接收情况；

✓调制方式：QPSK、16QAM 和 64QAM

✓纠错方式：RS 码、BCH 码、卷积码、TCM 码、TPC 码、Turbo 码、SCCC 码、LDPC 码等。

要求与 TS 流结构保持一致。MPEG TS 流的每个包有 188byte 组成，其中 1byte 的同步字和 187byte 的净荷信息，这就要求地面数字电视系统的帧结构最好传输整数个 TS 码块，即有效的帧长度是 188byte 的整数倍，这样便于与信源编码接口。

要求与 OFDM 长度保持一致。在多载波调制方式的地面数字电视系统中，采用了不同长度的正交频分复用（OFDM）调制。欧洲 DVB-T 使用了 1705（2k）或 6817（8k）个子载波的 OFDM 调制，而清华 DMB-T OFDM 使用了 3744 个有效子载波。

满足移动性能的要求。在移动接收过程中，由于多径反射和其它传播路径的影响，接收信号出现快慢衰落，是一个时变信道。为了便于处理，需要把处理限制在一个 OFDM 帧内，使之在一个 OFDM 帧近似为时不变信道。这样对 OFDM 帧长提出了一定的限制。

由于要满足上述这些要求，并且地面数字电视接收机正常工作的误码率要求在 10^{-11} 以下，因此，不同的地面数字电视广播系统采用了不同的纠错编码方式、不同的编码长度和不同的级联方式等，以便满足不同的要求。

1948 年，香农提出著名的信道理论：对于给定的有扰信道，若信道容量为 C，只要发送端以低于信道容量 C 的速率 R 发送信息，则一定存在一种编码方法，使得误码率随着码长的增加变得任意小。这一理论为实现高可靠的数据传输给出了研究的方向。向香农限无限靠近，成为无数科技人员辛勤工作的强劲动力。

现有的三个国际地面数字电视标准（美国 ATSC、欧洲 DVB-T 和日本 ISDB-T）都采用了由内码、交织、外码级联的基本结构。它们的外码都采用了 RS 缩短码，而内码有较大区别。美国 ATSC 纠错编码的内码采用了固定码率的格型码(TCM)，虽然性能很好，但只适应用于高码率。欧洲 DVB-T 和日本 ISDB 采用了多种码率的卷积码作为纠错编码的内码，虽然高码率时（64QAM）性能不如格型码，但同时适用于所有的码率（64QAM、16QAM、QPSK）。

随着纠错编码技术的发展，低密度纠错码（Low Density Parity Check ，LDPC）正在称为研究和应用的热点。

欧洲在 1994年制定了卫星数字视频广播标准 DVB-S (EN 300 421)，DVB-S 标准规定 QPSK 调制、卷积和Reed-Solomon 级联信道编码，现在它被世界范围内大部分的卫星电视和数据广播服务商所采用。在2004年底，欧洲又制定了第二代卫星数字视频广播标准 DVB-S2。DVB-S2 系统定义了实现基带数字信号的功能模块，其中前向纠错 (FEC) 编码模块

由 BCH 外码和内码(LDPC (Low Density Parity Check) 级联完成，码率包括1/4、1/3、1/2、3/5、2/3、3/4、4/5、5/6、8/9 和 9/10。依赖于应用领域，FEC 编码分组的长度是 *nldpc*=64800bits 或 16200 bits，如表1所示，从表中可以看到，DVB-S2 的BCH 码和 LDPC 码的码长都非常长，其实现复杂。当使用VCM 和 ACM 时，FEC 和调制模式可以在不同帧间进行改变，但在一个帧内保持不变。在 8PSK、16APSK和 32APSK 的 FEC 编码比特中采用了比特交织。调制的符号映射依赖于应用领域，可以采用QPSK、8PSK、16APSK 和 32APSK 星座映射。DVB-S2 采用的 BCH 和 LDPC 级联码的 BER~C/N 性能曲线如图 1 所示。

表 1 DVB-S2 的 BCH 和 LDPC 级联码的编码参数

LDPC code	BCH Uncoded Block K_{bch}	BCH coded block N_{bch} LDPC Uncoded Block k_{ldpc}	BCH t-error correction	LDPC Coded Block n_{ldpc}
1/4	16 008	16 200	12	64 800
1/3	21 408	21 600	12	64 800
2/5	25 728	25 920	12	64 800
1/2	32 208	32 400	12	64 800
3/5	38 688	38 880	12	64 800
2/3	43 040	43 200	10	64 800
3/4	48 408	48 600	12	64 800
4/5	51 648	51 840	12	64 800
5/6	53 840	54 000	10	64 800
8/9	57 472	57 600	8	64 800
9/10	58 192	58 320	8	64 800

上面阐述了地面数字电视有着多种多样的要求，为了适应这些不同的要求，采用的纠错编码也是各不相同，其中 LDPC 是目前纠错编码研究和应用的热点。在介绍 LDPC 码之前，我们首先描述几个与 LDPC 码相关的纠错编码概念或定义，包括线性分组码（linear block code）、系统线性分组码（systematic linear block code）、伴随式（或校正子 syndrome）、最小汉明距离、循环码（cyclic code）和准循环码（Quasi-cyclic code）等。

元素个数有限的域称为有限域，或者伽逻华域（Galois Field，GF），域中元素的个数称为域的阶，GF(q) 表示 q 阶有限域。

对于（n，k）二进制分组码，共有 2^k 个码字。若没有结构性特性，则编码器将需要记住 2^k 个消息序列和码字之间映射，当 k 值很大时，这将是不现实的。但线性特性使得编码器更为简单。当且仅当两个码字的模 2 和仍然是一个码字时，则二进制分组码是线性的。更正式的定义如下：

（n，k）分组码是线性分组码的条件是当且仅当它的 2^k 个码字构成 GF（q）上的 n 维线性空间 V_n 中的一个 k 维子空间 $V_{n,k}$。

因为（n，k）线性分组码 C 可以表示为 k 个线性独立的码字 g_0，g_1，....，g_{k-1} 的线性组合，因此，这 k 个线性独立的向量可以构成一个 k×n 矩阵，如下所示：

$$G=\begin{bmatrix} g_0 \\ g_1 \\ \mathrm{M} \\ g_{k-1} \end{bmatrix}=\begin{bmatrix} g_{00} & g_{01} & \mathrm{L} & g_{0,n-1} \\ g_{10} & g_{11} & \mathrm{L} & g_{1,n-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ g_{k-1,0} & g_{k-1,1} & \mathrm{L} & g_{k-1,n-1} \end{bmatrix}$$

其中，$g_i=(g_{i0}, g_{i1}, \ldots, g_{i,n-1})$，$0\leqslant i<k$。令 $u=(u_0, u_1, \ldots, u_{k-1})$ 表示要编码的信息，则相应的码字为：

$$v=u\cdot G=u_0g_0+u_1g_1+\ldots+u_{k-1}g_{k-1}$$

其中*矩阵 G 称为码字 C 的生成矩阵*。

线性分组码一个值得注意的特性是它可以构成一个系统结构（Systematic Structure），所谓的线性系统分组码的系统性就是每个码字可以分成清清楚楚的两部分，一部分是 k 个没有任何改变的原始信息数据，另一部分是（n－k）个冗余校验数据，它们是信息数据的线性组合。

线性系统分组码（n，k）完全可以用 k×n 阶的矩阵 G 表示：

$$G=\begin{bmatrix} g_0 \\ g_1 \\ \mathrm{M} \\ g_{k-1} \end{bmatrix}=\begin{bmatrix} 1 & 0 & \mathrm{L} & 0 & p_{0,0} & p_{0,1} & \mathrm{L} & p_{0,n-k-1} \\ 0 & 1 & \mathrm{L} & 0 & p_{1,0} & p_{1,1} & \mathrm{L} & p_{1,n-k-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & p_{i,j} & \mathrm{M} \\ 0 & 0 & \mathrm{L} & 1 & p_{k-1,0} & p_{k-1,1} & \mathrm{L} & p_{k-1,n-k-1} \end{bmatrix}$$

其中，$p_{i,j}=0$ 或 1。若令 I_k 表示 k×k 阶单位方阵，则上式表示为：

$$G=[I_k \ P]$$

其中 P 是 G 的子矩阵，它由 G 矩阵的后 n－k 列组成。

对于（n，k）线性分组码 C 的 k×n 阶生成矩阵 G，存在一个（n－k）×n 阶矩阵 H，H 由 n－k 个线性独立的行组成，以便 G 和 H 是正交的，即 H 的任意行与 G 的行空间是正交的。这表明 n 阶 v 是 C 的一个码字，当且仅当 $v\cdot H^T=0$ 时。*矩阵 H 称为码字的校验矩阵（parity-check matrix）*。V_n 子空间中与 C 中的码向量保持正交的每个向量构成了一个（n，n－k）线性分组码 C_d，*C_d 称为 C 的对偶码*。C_d 的生成矩阵是 C 的校验矩阵 H。

如果（n，k）线性分组码的生成矩阵是系统形式的，即 $G=[I_k \ P]$，则其校验矩阵可以表示为如下形式：

$$H=\begin{bmatrix} P^T & I_{n-k} \end{bmatrix}=\begin{bmatrix} p_{0,0} & p_{1,0} & \mathrm{L} & p_{k-1,0} & 1 & 0 & \mathrm{L} & 0 \\ p_{0,1} & p_{1,1} & \mathrm{L} & p_{k-1,1} & 0 & 1 & \mathrm{L} & 0 \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ p_{0,n-k-1} & p_{1,n-k-1} & \mathrm{L} & p_{k-1,n-k-1} & 0 & 0 & \mathrm{L} & 1 \end{bmatrix}$$

它清楚地表明了 G 和 H 之间的正交性，$G\cdot H^T=0$。

在（n,k）线性分组码的码字集合中，挑选前 i 个信息位数字均为 0 的所有码字，组成一个新的子集，则分组码（n-i，k-i）为 （n，k）码的*缩短码或截短码*。由于该子集的前 i 为均取 0，故传输时可以不传送它们，仅传后面的 k-i 位即可。该子集是 n-i 维线性空间中一个 k-i 维子群，构成一个（n-i,k-i）分组缩短码。换句话讲，对于（n-i，k-i）线性分组缩短码，可以在 k-i 位信息前填充 i 为零信息，使得信息为长度变为 k，然后进行（n,k）线性分组编码，最后去掉前 i 个信息位，即前面填充的 i 位零信息，就得到了（n-i，k-i）线性分组缩短码。该缩短码的最小距离至少与原码相同。缩短码的 G 矩阵只是在原 [n，k] 码的 G 阵中去掉左边 i 列和上面 i 行即可。缩短码或截短码的 H 矩阵，只要在原码 H 矩阵中去掉若干列即可得到。缩短码的编码码率 R 比原码要小。

如果一个（n，k）线性分组码的生成矩阵为G，校验矩阵为H，$v=(v_0,v_1,...,v_{n-1})$表示传输的码字，$r=(r_0,r_1,...,r_{n-1})$ 表示解调器输出端接收到的硬判决向量。由于信道噪声，r 可能不同于 v。向量和

$$e=r+v=(e_0,e_1,...,e_{n-1})$$

称为n阶错误图样（error pattern）。e 中的 1 是信道噪声引起的传输错误。由上式可知：

$$r=v+e$$

为了实现错误检测，解码器需要计算下面的 （n- k）重向量：

$$s=r\cdot H^T=(s_0,s_1,...,s_{n-k-1})$$

s 称为接收向量 r 的伴随式（或校正子）。那么 s=0，当仅当 r 是正确的码字时。因此，如果 $s\neq 0$，r 就不是码字，表明检测到有错误存在。当错误图样 e 刚好是一个非零码字时，虽然 r 中存在错误，但 r 有可能仍然是一个码字。这种错误图样称为不可检测错误图样。

一个码字检测随机错误和校正随机错误的能力取决于码字的最小距离（minimum distance）。首先我们定义二进制 n 重 v 中非零码元的个数为它的*汉明重量（Hamming Weight）*，表示为 w(v)。两个 n 重 v、u 之间对应位取值不同的个数，称为它们之间的*汉明距离*，用 d(v,u) 表示。汉明距离满足三角不等性，令 v、u、x 是三个 n 重向量，则

$$d(v,u)+d(u,x)\geq d(v,x)$$

两个 n 重 v、u 之间的汉明距离等于它们和的汉明距离，即 $d(v,u)=w(v+u)$。

对于给定的分组码 C，可以计算任意两个码字之间的汉明距离，C 的最小汉明距离表示为 d_{min}，其定义如下：

$$d_{\min}=\min\{d(v,u):v,u\in C,v\neq u\}$$

如果 C 是一个线性分组码，两个码字之和仍然是一个 C 中的码字，因此，C 中两个码字 v、

u之间的汉明距离就等于C中另外一个码字（v+u）的汉明重量，即

$$\begin{aligned} d_{\min} &= \min\{d(v,u): v,u \in C, v \neq u\} \\ &= \min\{w(v+u): v,u \in C, v \neq u\} \\ &= \min\{w(x): v,x \in C, x \neq 0\} \\ &\triangleq w_{\min} \end{aligned}$$

其中 w_{min} 是线性码C的最小重量。因此，在纠错编码中有如下一个重要的定理：

线性分组码的最小距离等于非零码字的最小重量，反之也成立。

下面的定理表述了线性分组码最小距离和它的校验矩阵之间的关系：

具有校验矩阵H的线性码C的最小距离等于H中其和为0的列的最小数目。

线性分组码的一个重要子类是循环码（Cyclic Code），其编码和伴随式计算可以通过带反馈的移位寄存器实现。循环码具有严谨的代数结构，其性能易于分析；循环码具有循环特性，编解码电路简单，易于实现。目前已发现的大部分线性码与循环码有密切关系，它们中的大部分都可归结于循环码。

令 $v=(v_0,v_1,...,v_{n-1})$ 是长度为n的向量，如果我们把v的码元循环右移i位，0<i<n，将得到另一个n重 $v^{(i)}=(v_{n-i},v_{n-i+1},...,v_{n-1},v_0,v_1,...,v_{n-i-1})$。很明显，v循环右移i位等同于v循环左移n—i位。

(n，k)线性码C是循环码，当且仅当C中码字的每一个循环移位仍然是C中码字时。

循环码的码字 $v=(v_0,v_1,...,v_{n-1})$ 可以用多项式表示，其系数是v的组合：

$$v(X)=v_0+v_1X+v_2X^2...+v_{n-1}X^{n-1}$$

循环码C中最小度（degreen）的非零码多项式是唯一的。如果令 $g(X)=g_0+g_1X+g_2X^2...+g_{r-1}X^{r-1}+X^r$ 是（n，k）循环码C中最小度的非零码多项式，那么常数项 g_0 将恒为1。同时，n-1或更小度的二进制多项式是一个码字的多项式，当且仅当它是g(x)的乘积。在（n，k）循环码中，有且只有一个n-k度的码多项式：

$$g(X)=1+g_1X+g_2X^2+...+g_{n-k-1}X^{n-k-1}+X^{n-k}$$

因为最小度g(X)的非零码多项式是唯一的，因此它称为码C的生成多项式。

g(X)和其循环右移k位的 $g^{(k)}(X)$ 之间存在着简单的关系：

$$X^k g(X)=(X^n+1)+g^{(k)}(X)$$

因为 $g^{(k)}(X)$ 是码多项式，那么g(X)的乘法意味着g(X)是 X^n+1 的因式。如果g(x)是n—k度

的多项式，并且是 X^n+1 的因子，则 g(x) 就可生成一个 （n，k）循环码。

另一类重要的纠错码是准循环码（Quasi-cyclic code，QC code）。循环码 C 的每一个码字左移或右移一位后仍然是 C 的一个码字，即循环码在循环移位下具有不变性。但是，对某些码而言，并不具有这种性质。虽然码字循环移位一位得到的不是该码的码字，但若循环移位 n_0 （≠1）位，得到的仍然是该码的一个码字。因此，

一个（mn_0，mk_0）线性分组码，若它的任意一个码字左或右循环移位 n_0 （≠1）位后，得到的仍是该码的一个码字，则称这类码为准循环码（Quasi-cyclic code，QC code）。

（mn_0，mk_0）准循环码 C 的生成多项式具有如下形式：

$$G=\begin{bmatrix} IP_0 & 0P_1 & \mathrm{L} & 0P_{m-1} \\ 0P_{m-1} & IP_0 & \mathrm{L} & 0P_{m-2} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ 0P_1 & 0P_2 & \mathrm{L} & IP_0 \end{bmatrix}$$

其中，I 和 0 分别表示 $k_0\times k_0$ 阶的单位方阵和零矩阵，P_i 是 $k_0\times(n_0-k_0)$ 阶矩阵。所以，准循环码的每一个码字由 m 段组成，每段前面位 k_0 个信息元，后面 n_0-k_0 个为校验元。相应的校验矩阵为：

$$H=\begin{bmatrix} P_0^T I & P^T{}_{m-1}0 & \mathrm{L} & P_1^T 0 \\ P_1^T 0 & P_0^T I & \mathrm{L} & P_2^T 0 \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ P^T{}_{m-1}0 & P^T{}_{m-2}0 & \mathrm{L} & P_0^T I \end{bmatrix}$$

其中，I 和 0 分别表示 （n_0-k_0）×（n_0-k_0）阶的单位方阵和零矩阵。我们可以通过对 G 矩阵进行列交换而得到准循环码系统码形式的 G 矩阵和 H 矩阵。

（mn_0，mk_0）准循环码生成多项式更一般的表达形式如下：

$$G=\begin{bmatrix} G_0 & G_1 & \mathrm{L} & G_{m-1} \\ G_{m-1} & G_0 & \mathrm{L} & G_{m-2} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ G_1 & G_2 & \mathrm{L} & G_0 \end{bmatrix}$$

其中每个 G_i 是一个 $k_0\times n_0$ 阶子矩阵，可以看到，这种形式的生成矩阵完全由它的前 k_0 行决定。

循环矩阵是一个方阵，每一行是它上面一行的循环移位（右移一位），第一行是最后

一行的循环移位。对于这种循环矩阵，每一列是它左面一列向下的循环移位，第一列是最后一列的循环移位。循环矩阵的行、列重量是相等的，即 w。简单些，我们就称循环矩阵的重量为 w。如果 w=1，那么循环矩阵是置换矩阵，*称为循环置换矩阵(circulant permutation matrix)*。对于循环矩阵，列集合（从上到下读）与行集合（从右到左读）是相同的。循环矩阵的特性完全可以由它的第一行所体现，因此，第一行*称为循环矩阵的生成多项式*。

我们把准循环码的生成矩阵 G 表示成循环矩阵形式，如下：

$$G=\left[M_0 M_1 \ldots\ldots M_{m-1}\right],\ \text{其中，}M_l=\begin{bmatrix} G_l \\ G_{l-1} \\ \mathrm{M} \\ G_0 \\ G_{m-1} \\ \mathrm{M} \\ G_{l+1} \end{bmatrix}\ for\ 0\le l<m$$

对于 $0\leqslant j<n_0$，令 Q_j 为抽取 $M_0,M_1,\ldots,M_{m-1}$ 中第 j 列而构成的一个 $mk_0\times m$ 阶矩阵。因此，G 可以表示如下形式：

$$G'=\left[Q_0 Q_1 \ldots\ldots Q_{n_0-1}\right]$$

对于 $0\leqslant i<k_0$，令 $Q_{i,j}$ 为抽取 Q_j 中第 (lk_0+i) 列而构成的一个 $m\times m$ 阶矩阵，$0\leqslant l<m$，。因此，G 可以表示成如下循环矩阵形式：

$$G''=\begin{bmatrix} Q_{0,0} & Q_{0,1} & \mathrm{L} & Q_{0,n_0-1} \\ Q_{1,0} & Q_{1,1} & \mathrm{L} & Q_{1,n_0-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ Q_{k_0-1,0} & Q_{k_0-1,1} & \mathrm{L} & Q_{k_0-1,n_0-1} \end{bmatrix}$$

类似地，准循环码的校验矩阵也可以表示成循环矩阵形式。

如果准循环码的生成矩阵具有如下形式：

$$G_{qc}=\begin{bmatrix} G_0 \\ G_1 \\ \mathrm{M} \\ G_{k_0-1} \end{bmatrix}=\begin{bmatrix} I & O & \mathrm{L} & O & Q_{0,0} & Q_{0,1} & \mathrm{L} & Q_{0,n_0-k_0-1} \\ O & I & \mathrm{L} & O & Q_{1,0} & Q_{1,1} & \mathrm{L} & Q_{1,n_0-k_0-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ O & O & \mathrm{L} & I & Q_{k_0-1,0} & Q_{k_0-1,1} & \mathrm{L} & Q_{k_0-1,n_0-k_0-1} \end{bmatrix}=\left[I_{k_0}\ P\right]$$

其中，I 是 b×b 单位矩阵，O 是 b×b 阶零阵，$Q_{i,j}$ 是 b×b 阶循环矩阵，$0\leqslant i\leqslant k_0-1$，$0\leqslant j\leqslant n_0-k_0-1$。这种形式的生成矩阵 G_{qc} *称为系统循环（Systematic Circulant，SC）形式*，它由两部分组成，左边是 P 矩阵，它对应于码字的冗余校验部分，右边是 I_{k0} 矩阵。因为 P 是一个循环矩阵的阵列，因此，我们说 G_{qc} 是 SC 形式。

LDPC（Low Density Parity Check，低密度校验码）码是一种基于稀疏矩阵的奇偶校验码。LDPC码是麻省理工学院罗伯特.哥拉格（Robert Gallagher）在1963年在他的博士论文中提出的一种好码，其性能接近Shannon限。他不但证明了LDPC码的优良性能：(1)这些码字的典型最小距离随码长的增加线性增加；(2)BSC信道下译码错误的典型概率随码长指数减小，而且提出了两种迭代算法。但由于其实现，特别是硬件实现复杂(需要大量的乘除运算)，当时并不被人所关注。这种码后来被称作古典码型。

随后在很长一段时间内，没有受到人们的重视，早已被人遗忘。1993年法国人Berrou等提出了Turbo迭代译码后，1995年Mackay和Neal发现LDPC码的性能足可以和Turbo码相比，而且在实现上更有优势，于是重新介绍了LDPC，这种沉寂了多年的码字很快成为一种引人瞩目的纠错编码方式，其设计、构造、解码、性能分析和应用等成为了人们研究的热点。1998年，MacKay和Spielman发明了不规则的LDPC。Richardson和Urbanke开创了用译码分析设计码型的方法。

LDPC码是很长的线性分组码，它的校验矩阵 $H_{(n-k)\times n}$ 是一个稀疏矩阵，每个码字满足一定数目的线性约束，而约束的数目通常是非常小的,易于译码。

LDPC码将要发送的信息 $u=\{u_1,u_2,\ldots,u_m\}$ 转换成被传输的码字 $v=\{v_1,v_2,\ldots,v_n\}=u\ G$，$n>m$，n表示分组的长度，n的取值范围通常从数千到几十万。与生成矩阵G相对应的是一个校验矩阵H，H满足 $Hv^T=0$，H是一个几乎全部由0组成的稀疏矩阵,每行和每列中1的数目都很少，例如3、4和5等。

Gallagher定义的（n,p,q）LDPC码是码长为n的码字，在它的校验矩阵H中，每一行和列中1的数目（码重）是固定的，其中每一列1的个数（重量）都是 γ，行的个数（重量）都是 ρ，$\gamma\geq 3$。任意两行(或两列)之间1的重叠数目小于等于1，*这称为行列约束(Row-Column Constraint，RC约束)*。如果校验矩阵H的每一行是线性独立的，那么码率为 $(\rho-\gamma)/\rho$，否则码率是 $(\rho-\gamma')/\rho$，其中 γ' 是校验矩阵H中行线性独立的数目。

图2是由Gallagher构造的一个（20,3,4）LDPC码的校验矩阵，它的 $d_{min}=6$，设计码率为1/4，实际码率为7/20。

这种校验矩阵每行和每列中1的数目（汉明重量）相同的LDPC码称为规则LDPC码(Regular LDPC code)，由规则LDPC码的校验矩阵H得到如图3所示的双向图（Bipartite Graph)。在图的下方每一个节点 x_n 代表的是变量节点（Variable Node)，上方每一个节点 z_m 代表的是校验节点（Check Node)。把某列 x_n 与该列中非零处的 z_m 相连，例如对于 x_2 列,这列中三个1分别对应于 z_1、z_7 和 z_{12} 行，这样就把 x_2 和 z_1、z_7 和 z_{12} 连接起来。从行的角度考虑，把某一行 z_m 中非零点处的 x_n 相连，得到同一个双向图。在规则LDPC码中，与每个信息节点相连边的数目是相同的，校验节点也具有相同的特点。在译码端，把与某一个校验节点 z_m

相连的 x_n 求和，结果若为 0，则无错误发生。

Gallagher 对规则 LDPC 码译码采用的迭代概率译码算法称为信度传播（Belief Propagation）。

与规则 LDPC 码相对应的是非规则 LDPC 码 (Irregular LDPC code)，其校验矩阵 H 中每行或列中 1 的个数（重量）不同，例如3、4和5，列中1的个数也不一样。其编码方法与规则 LDPC 码基本相同，非规则双向图中信息节点之间、校验节点之间的度由可能不同。因此，对于非规则图构造的 LDPC 码，它的校验矩阵 H 的列重量不相同，是一个变化的值，这是非规则码与规则码之间的重要区别。

非规则码的性能要好于规则码。最近几年的研究表明，对于在 GF(8) 构造的非规则码，它的性能要比 Turbo 码还好，能够显著提高码字性能，其性能非常接近于 Shannon 限。非规则的译码可以采用可信传播迭代译码算法，也可以采用序列译码和并行译码算法等。

准循环低密度校验码（Quasi-cyclic LDPC Code，QC-LDPC）的生成矩阵为稀疏循环置换矩阵，QC-LDPC 与其它类型的 LDPC 相比，编码实现简单，采用简单的反馈移位寄存器即可，其复杂度与校验比特的数目呈线性比例。关于QC-LDPC 码，请参考

(1) Z. W. Li, L. Chen, S. Lin, W. Fong, and P. S. Yeh, "Efficient Encoding of Quasi-Cyclic LDPC Codes," submitted to IEEE Trans. On Communications, 2003。

QC-LDPC 码生成矩阵 G_{qc} 和校验矩阵是一个稀疏循环置换矩阵，其中校验矩阵 H_{qc} 是一个 $\gamma \times \rho$ 阵列，其构成矩阵 $B_{i,j}$ 为（q-1）×(q-1) 阶循环置换矩阵，为：

$$H_{qc}(\gamma,\rho)=\begin{bmatrix} B_{1,1} & B_{1,2} & \mathrm{L} & B_{1,\rho} \\ B_{2,1} & B_{1,2} & \mathrm{L} & B_{2,\rho} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ B_{\gamma,1} & B_{\gamma,2} & \mathrm{L} & B_{\gamma,\rho} \end{bmatrix}$$

其满足 RC 约束。$H(\gamma,\rho)$ 的零空间给出了 QC-LDPC 码 C_{qc}，长度为 $n=(q-1)\rho$。

人们提出了许多构造循环置换矩阵的方法。在早期的方法中，主要基于本原域，产生的循环置换矩阵可能满足或不满足 RS 约束，这类方法有：

(2) R. J. G. Smith, "Easily Decodable Efficient Self-Orthogonal Block Codes," Electronics Letters, vol. 13, no. 7, pp. 173-174, March 1977。

(3) M. Blaum, P. Farrell, and H. van Tilborg, "Array Codes," Handbook of Coding Theory, V. S. Pless and W. C. Huffman Eds., Elsevier 1998。

(4) R. M. Tanner, "Minimum-Distance Bounds by Graph Analysis," IEEE Trans. on Inform Theory, vol 47, no. 2, pp. 808-821, February, 2001。

(5) J. -L. Kim, U. Peled, I. Perepelitsa, and V. Pless, "Explicit Construction of families of

LDPC Codes with Girth at Least Six," the 40th Annual Allerton Conference on Communications, Control and Computing, Oct. 2002。

(6) R. M. Tanner, D. Sridhara, A. Sridharan, T. E. Fuja, and D. J. Costello, jr, "LDPC Block and Convolutional Codes based on Circulant Matrices," submitted to IEEE Trans. on Inform. Theory, 2004。

(7) B. Vasic and O. Milenkovic, "Combinatorial Construction of Low Density Parity Check Codes for Iterative Decoding," IEEE Trans. On Inform. Theory, vol 50, no. 6, pp.1156-1176, June 2004。

上述方法基本相同，其构造的矩阵也类似。后来人们又提出了基于矩阵分解的循环置换矩阵构造方法，这类方法的参考文献有：

(8) L. Chen, J. Xu, I. Djurdjevic, and S. Lin, "Near Shannon Limit Quasi-Cyclic LDPC Codes," IEEE Trans. on Commun. Theory, vol. 52,no. 7, July 2004 (also in Proc. 2003 IEEE GlobeCom., pp. 2030-2035, San Francisco, December 1-5, 2003)。

(9) B. Ammar, E. Honary, Y. Kou, J. Xu, and S. Lin, "Construction of Low Density Parity Check Codes based on Balanced Incomplete Block Designs," IEEE Trans. Inform. Theory, vol. 50, no. 6, pp. 1257-1268, June, 2004。

上述方法中，把循环矩阵 H 按行和列进行分解，得到递减的 $\gamma \times \rho$ 阶的循环子矩阵：

$$H = \begin{bmatrix} B_1^{(1)} & B_2^{(1)} & \mathrm{L} & B_\rho^{(1)} \\ B_1^{(2)} & B_2^{(2)} & \mathrm{L} & B_\rho^{(2)} \\ \mathrm{M} & \mathrm{M} & \mathrm{M} & \mathrm{M} \\ B_1^{(\gamma)} & B_2^{(\gamma)} & \mathrm{L} & B_\rho^{(\gamma)} \end{bmatrix}$$

如果 $B_t^{(k)}$ 的重量为1，则 H 是循环置换矩阵。基于矩阵分解的构造方法有欧几里德几何法和投影几何法，以及具有两个信息符号的 RS 码法，其中RS 码法的参考文献为：

(10) Ivana Djurdjevic, Jun Xu, Khaled Abdel-Ghaffa and Shu Lin, "A Class of Low-Density Parity-Check Codes Constructed Based on Reed-Solomon Codes With Two Information Symbols", IEEE Comm. Letters, vol. 7, no. 7, pp317-318, July 2003。

RS 码是 MIT Lincoln Lab 的 Reed 和 Solomon 在 1960 年发表的论文 "Polynomial Codes over Certain Finite Fields" 中提出的。 RS 码是一种效率很高的分组码，既适用于纠随机误码，也特别适用于纠突发误码。RS 码是一类非二进制 BCH 码, 每个符号由 m 比特组成。去掉 RS 码的某些信息码元后, 分组长度缩短, 只要监督码元数不变, 码的最小距离就不会减少, 即任何一种缩短的 RS 码仍是一个最大码。RS 码的码字内的任何 k 个位置都可用作信息集

合。即任何一个 GF(q) 上的 (n,k) RS 码, 对任何 k 个符号位置, 将只有一个与这 k 个位置相对的码字。这便于得到任何 RS 码的确切权分布。RS 码的精确纠错性能是由码的最小距离和权分布决定的, RS 码的这两项指标都完全由 n 和 k 决定。这非常便于根据指标设计 RS 码, 得到广泛应用。

在（n，k）RS 码中，输入信息被分成 km 比特一组，每组包括 k 个符号，每个符号由 m 比特组成。纠正 t 个符号错误的 RS 码参数如下：

码长 n = 2m-1 符号， 或 m(2m-1) 比特

信息段 k 符号， 或 km 比特

监督段 n-k=2t 符号， 或 m(n-k) 比特

最小码距 d=2t+1 符号， 或 m(2t+1) 比特

LDPC 码的性能与码的结构、生成方式、码长度、最小汉明距离、校验矩阵行列的码重分布、解码算法、双向图的周长、双向图的循环分布、双向图的连接特性、双向图的节点度分布等有关。例如分别以列均匀方式和行列均匀方式生成码率为 1/2、帧长为 250 的规则 LDPC 码，其 E_b/N_0～BER 性能如图 4 所示，误码率为 10^{-4} 时，采用列方式（evencol）需要提供 3dB 的信噪比，而采用行列方式（evenboth）只需要 2.7dB 就可以了。在此例中，evenboth 方式的效果明显比 evencol 的效果好。在给定相同的信噪比的情况下，evenboth 方式的误码率比 evencol 的误码率低，并随着信噪比的增大，evenboth 的误码率下降的更快。

再如，以行列均匀方式生成的不同帧长的 1/2 码率规则 LDPC 码，帧长分别的性能比较 250、1000、5000，BER 误码率与 E_b/N_0 信噪比的关系如图 5 所示，从中可以看出，在每一种帧长下，误码率都随着信噪比的增大而快速降低。并且帧长越长，这种下降速度越快，达到相同的误码率所需要的信噪比就越低。

LDPC 码从 60 年代的初露头角，到今天的大放异彩，经历了将近半个世纪。这几十年里，发展并完善了 LDPC 码的一整套理论和相关技术。LDPC 码的广泛应用背景，成为当今世界各国学术界和 IT 业界高度重视的热点。

任何一项优良的技术，只有在现实中得到实际应用，才能体现出其真正的价值。LDPC 码无疑是一种优秀的纠错码型，它与香农限的靠近程度是目前已有的其它任何码型都无法比拟的，但是在有些方面还有待改进，例如实现复杂度、系统时延、编码效率等，或者换句话讲，需要根据实际应用系统的需求，例如前面所述的地面数字电视广播的特殊需求，需要对 LDPC 码的设计、构造、解码、性能分析和应用等方面的问题进行具体的分析和设计：

1）LDPC 码的设计、构造及优化

目前，人们提出了各种 LDPC 码设计和构造方法。基于构造的方法来分类，LDPC 码分

为两类：随机码（Random Code）和结构码（Structured Code），

随机 LDPC 码的构造采用计算机搜索方式，但需要满足一定的限制条件，或需要的码图结构，例如节点的周长和度分布。这种搜索生成方式需要的计算量呈$o(n^2)$，不规则 LDPC 的性能要在长码下才更好，但计算量会更大。因此，需要寻求减少计算的方法，探求校验矩阵 H 的结构。虽然目前已有了一些代数方式可以减少计算量，使得对于某些具有特殊结构的 H 的码型，编码计算量可以随码长的增加而线性增长。但是，这些结构是否为最优结构，度的分布更复杂时这个结论是否依然成立还不得而知。

而结构化的 LDPC 码是基于代数和组合的方法来构造的。长的随机 LDPC 码通常比同等的结构化 LDPC 码更接近于仙农极限，但是它们没有充分的结构化，从而简化编码。相反，结构化 LDPC 码通常有更好的编码优势，特别是循环或准循环 LDPC 码（Quasi-Cyclic LDPC，QC-LDPC），它们可以采用简单的移位寄存器实现编码，对于串行编码，其复杂度线性正比于校验比特的数目；对于并行编码，其复杂度正比于码长。对于一个实际应用的码字长度，人们已经证明了一个好的设计的结构化的 LDPC 码性能（误比特率、误块率和误码平台）等同于随机 LDPC 码。

因此，寻找更好的或更适合于特定应用的 LDPC 的编码方式和 LDPC 码结构是必须解决的一个问题之一。

2）LDPC 在非标准信道下性能的研究

目前已有的文献对 LDPC 的研究都是在标准信道下，尤其是对称信道下进行。而在地面数字电视广播中，发射塔到移动接收是典型的非标准信道，并且存在着非常严重的多径效应和多普勒效应的影响。LDPC 码要在移动接收环境中得以应用，必须要研究它在这种信道下的性能是否依然卓越，并且寻求专用于这种信道下的优化方法。

3）更有效的解码算法

目前有各种不同的 LDPC 译码算法。最简单的是采用硬判决的大数逻辑（Majority-logic，MLG）算法。常用的 LDPC 解码算法是比特翻转（Bit-flipping，BF）算法、加重比特翻转（Weighted BF）算法和基于信度传递的迭代译码（Iterative Decoding Based on Belief Progagation，DBP）算法等三种，其中 IDBP 也就是和积算法（Sum-Product Algorithm，SPA），在上述解码算法和实现中，SPA 算法给出了最好的性能。

因此，寻找更有效、性能更好的解码算法也是当前 LDPC 研究的热点。

4）LDPC 的硬件实现

由于 LDPC 的计算量大，因此，采用专用芯片来实现 LDPC 的编码和解码将大大提高系统效率，研究利用专用的 DSP 芯片实现 LDPC 码的编译码功能也是一大热点。

当然还存在一些其它问题，在地面数字电视广播实际应用中，我们都要面对这些问题，

只有对这些问题给出一个综合考虑和解决方案，才能得到一个适用的 LDPC 码

发明内容

本发明的目的是针对我国地面数字电视广播传输的各种需求，例如不同的调制方式；便于与 TS 码流接口，携带整数个 TS 码块；与 OFDM 调制长度保持一致；为了提供更好的移动性能，时变特性限制在一个 OFDM 帧中，提出了一种基于 BCH 码和 LDPC 码串行级联的多码率纠错编码方法，通过不同的 LDPC 码构造，得到多种编码码率，以便适用不同的信道情况。本发明采用了较短的码长，实现结构简单，具有较小的时延和良好的纠错性能，更好地适应各种应用。

本发明提出了一种采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传递流包每 2 个分为一组，长度为 3008 比特，即 188×8×2=3008 比特，然后再进一步分为 4 个子组，每一子组的长度为 752 比特；

2）对 4 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的总的输出长度为 4×762=3048 比特；

3） 对 4 个 BCH(762,752) 码 C_{BCH} 输出的 3048 个比特进行 LDPC(7493,3048) 编码，即与 LDPC(7493,3048) 预定的生成矩阵 G_{qc1} 相乘，得到 QC-LDPC (7493,3048) 码 C_{qc1}，所述生成矩阵 G_{qc1} 的构造基于预定的校验矩阵 H_{qc1}；

4）删除 LDPC(7493,3048) 码 C_{qc1} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出纠错编码比特。

所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^3+1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH 码是系统码，去掉的 261 个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 的校验矩阵 H_{qc1} 由循环置换矩阵构成：

$$H_{qc1}=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，γ=35，ρ=59，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$阶循环矩阵，q=128，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$。

所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 校验矩阵 H_{qc1} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽逻华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移一位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，对应于 $GF(p^s)$ 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，用 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0\le i\le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix}A_0\\ \mathrm{M}\\ A_{\gamma-1}\end{bmatrix}=\begin{bmatrix}A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1}\\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M}\\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1}\end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 $\gamma\times\rho$ 阶掩蔽矩阵 $W(\gamma,\rho)$，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc1}：

$$H_{qc1}=\begin{bmatrix} w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho} \\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho} \\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M} \\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho} \end{bmatrix}$$

其中

$$w_{i,j}A_{i,j}=\begin{cases} A_{i,j}, & \text{对于 } w_{i,j}=1 \\ 0, & \text{对于 } w_{i,j}=0 \end{cases}。$$

所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 生成矩阵 G_{qc1} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=35，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。

所述的采用 BCH(762,752) 和 LDPC(7488, 3048) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,3048) 生成矩阵 G_{qc1} 的构造基于所述的校验矩阵 H_{qc1}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma\times\rho$ 阶矩阵 D：

$$D=\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T}=|O|$，|O| 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_i u^T+Dz_i^T=0$，则

$$z_i^T=D^{-1}M_i u^T$$

其中

$z_i=(g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\gamma-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_i=[B^T_{0,j}\ldots\ldots B^T_{\gamma-1,j}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u=(1\ \ 0\ \mathrm{K}\ \ 0)$ 表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的

QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b）由 $z_i^T = D^{-1}M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

c）在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*, g_1^* \cdots\cdots, g_{\gamma(q-1)-r-1}^*$，以构成完整的生成矩阵 G_{qc} 表达式：

$$G_{qc}^* = \begin{bmatrix} G \\ - \\ Q \end{bmatrix} = \left[\begin{array}{cccccccc} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \\ \hline & & & & g_0^* & & & \\ & & & & g_1^* & & & \\ & & & & \mathrm{M} & & & \\ & & & & g_{\gamma(q-1)-r-1}^* & & & \end{array}\right]$$

而 g_i^* 由求解下列矩阵等式获得：

$$\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix} \begin{bmatrix} g_{i,(\rho-\gamma)(q-1)}^* \\ g_{i,(\rho-\gamma)(q-1)+1}^* \\ \mathrm{M} \\ g_{i,\rho(q-1)-1}^* \end{bmatrix} = 0 \text{。}$$

所述的一种采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传送流包每 3 个分为一组，长度为 4512 比特，即 188×8×3=4512 比特，然后再进一步分为 6 个子组，每一子组的长度为 752 比特；

2）对 6 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的总的输出长度为 6×762=4572 比特；

3）对 6 个 BCH(762,752) 码 C_{BCH} 输出的 4572 个比特进行 LDPC(7493,4572) 编码，即与 LDPC(7493, 4572) 的生成矩阵 G_{qc2} 相乘，得到 QC-LDPC (7493, 4572) 码 C_{qc2}，所述生成矩阵 G_{qc2} 的构造基于预定的校验矩阵 H_{qc2}；

4）删除 LDPC(7493,4572)码 C_{qc2} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个

输出纠错编码比特。

所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^3+1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH码是系统码，去掉的261个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7488,4572) 的校验矩阵 H_{qc2} 由循环置换矩阵构成：

$$H_{qc2}=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，$\gamma=23$，$\rho=59$，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$ 阶循环矩阵，$q=128$，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$。所述的校验矩阵 H_{qc2} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽逻华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移一位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，对应于 $GF(p^s)$ 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，用 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或

相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0 \le i \le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix}A_0\\ \mathrm{M}\\ A_{\gamma-1}\end{bmatrix}=\begin{bmatrix}A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1}\\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M}\\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1}\end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 $\gamma\times\rho$ 阶掩蔽矩阵 $W(\gamma,\rho)$，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc2}：

$$H_{qc2}=\begin{bmatrix}w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho}\\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho}\\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M}\\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho}\end{bmatrix}$$

其中

$$w_{i,j}A_{i,j}=\begin{cases}A_{i,j}, & \text{对于 } w_{i,j}=1\\ 0, & \text{对于 } w_{i,j}=0\end{cases}。$$

所述的采用 BCH(762,752) 和 LDPC(7488, 4572) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,4572) 生成矩阵 G_{qc2} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix}I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1}\\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M}\\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1}\end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=23，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。所述的生成矩阵 G_{qc2} 的构造基于所述的校验矩阵 H_{qc2}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma\times\rho$ 阶矩阵 D：

$$D=\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T}=|O|$，|O| 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_i u^T + Dz_i^T = 0$，则

$$z_i^T = D^{-1}M_i u^T$$

其中

$z_i=(g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\rho-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_i=[B^T_{0,i}.....B^T_{\gamma-1,i}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u=(1\ \ 0\ \mathrm{K}\ \ 0)$表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b) 由 $z_i^T = D^{-1}M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

c) 在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*, g_1^*.....g^*_{\gamma(q-1)-r-1}$，以构成完整的生成矩阵 G_{qc} 表达式：

$$G^*_{qc}=\begin{bmatrix} G \\ - \\ Q \end{bmatrix}=\left[\begin{array}{ccccccccc} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \\ \hline & & & & g^*_0 & & & \\ & & & & g^*_1 & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-r-1} & & & \end{array}\right]$$

而 g_i^* 由求解下列矩阵等式获得：

$$\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix} \begin{bmatrix} g^{*}_{i,(\rho-\gamma)(q-1)} \\ g^{*}_{i,(\rho-\gamma)(q-1)+1} \\ \mathrm{M} \\ g^{*}_{i,\rho(q-1)-1} \end{bmatrix} = 0 。$$

所述的一种采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，包括以下步骤：

1）将输入的 TS 传递流包每 4 个分为一组，长度为 6016 比特，即 188×8×4=6016 比特，然后再进一步分为 8 个子组，每一个子组的长度为 752 比特；

2）对 8 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的总的输出长度为 8×762=6096 比特；

3）对 8 个 BCH(762,752) 码 C_{BCH} 输出的 6096 个比特进行 LDPC(7493,6096) 编码，即与 LDPC(7493, 6095) 的生成矩阵 G_{qc3} 相乘，得到 QC-LDPC (7493, 6096) 码 C_{qc3}，所述生成矩阵 G_{qc3} 的构造基于预定的校验矩阵 H_{qc3}；

4）删除 LDPC(7493,6096) 码 C_{qc3} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出纠错编码比特。

所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的BCH(762, 752) 是BCH(1023,1013) 的缩短码，按以下步骤生成：

1) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

2) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，其生成多项式为 $x^{10}+x^{3}+1$，输出 1023 个纠错编码比特；

3) 去掉纠错编码比特的前 261 个比特，因为 BCH码是系统码，去掉的261个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}。

所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,6096) 的校验矩阵 H_{qc3} 由循环置换矩阵构成：

$$H_{qc3} = \begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中，$\gamma=11$，$\rho=59$，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$阶循环矩阵，q=128，$0 \le i \le \gamma-1$ 和 $0 \le j \le \rho-1$。所述的校验矩阵 H_{qc3} 的构造基于两信息符号的里德-所罗门码，即 RS 码，步骤如下：

1) 在伽逻华域 GF(q) 中生成 $RS(q-1,q-\rho,\rho-1)$ 码 C_{RS}，其中码长为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$；

2) 从 C_{RS} 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b；

3) 构造码字集合 $C_b^{(i)}$：c 是 C_b 中一个码重为 ρ 的码字，C_b 中 $q-1$ 个非零码字构成一维码字的集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$，基于子码 $C_b^{(1)}$，通过循环右移一位，把 C_b 分割为 $q-1$ 个陪集，即 $C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的；

4) 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$：$z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重，对应于 GF(p^s) 的 $q-1$ 个非零的域元素 α^i，而 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，用 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$，那么 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合为 $Z(C_b^{(i)})=\{z(b):b\in C_b^{(i)}\}$，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的；

5) 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i：对于 $0\le i\le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量；

6) 由符号位置矩阵 A_i 构造一个规则的基矩阵 $H_{GA}(\gamma,\rho)$：

$$H_{GA}(\gamma,\rho)=\begin{bmatrix}A_0\\ \mathrm{M}\\ A_{\gamma-1}\end{bmatrix}=\begin{bmatrix}A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1}\\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M}\\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1}\end{bmatrix}$$

其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点；

7) 通过计算机随机构造一个 GF(2) 上非规则的 $\gamma\times\rho$ 阶掩蔽矩阵 $W(\gamma,\rho)$，使其列、行重量分布分别等于预先计算确定的变量节点度分布 v(X) 和校验节点度分布 c(X)；

8) 通过下列矩阵乘法获得 QC-LDPC 的校验矩阵 H_{qc3}：

$$H_{qc3}=\begin{bmatrix}w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho}\\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho}\\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M}\\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho}\end{bmatrix}$$

其中

$$w_{i,j}A_{i,t}=\begin{cases}A_{i,j}, & \text{对于 } w_{i,j}=1\\ 0, & \text{对于 } w_{i,j}=0\end{cases}。$$

所述的采用 BCH(762,752) 和 LDPC(7488, 6096) 级联码的纠错编码方法，其特征在于，所述的 LDPC(7493,6096) 生成矩阵 G_{qc3} 由循环置换矩阵构成：

$$G_{qc}=\begin{bmatrix} I & O & \cdots & O & G_{0,0} & G_{0,1} & \cdots & G_{0,\gamma-1}\\ O & I & \cdots & O & G_{1,0} & G_{1,1} & \cdots & G_{1,\gamma-1}\\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & G_{i,j} & \vdots\\ O & O & \cdots & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \cdots & G_{\rho-\gamma-1,\gamma-1}\end{bmatrix}$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，q=128，γ=11，ρ=59，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。所述的生成矩阵 G_{qc3} 的构造基于所述的校验矩阵 H_{qc3}，步骤如下：

1) 把 H_{qc} 的循环矩阵的列重新排列，构造一个 $\gamma\times\rho$ 阶矩阵 D：

$$D=\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma-1} & \cdots & B_{0,\rho-1}\\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \cdots & B_{1,\rho-1}\\ \vdots & \vdots & \ddots & \vdots\\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \cdots & B_{\gamma-1,\rho-1}\end{bmatrix}$$

D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

2) 根据编码理论 $H_{qc}G_{qc}^{T}=|O|$，|O| 是 $\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵，得等式 $M_i u^T+Dz_i^T=0$，则

$z_i^T=D^{-1}M_i u^T$

其中

$z_i=(g_{i,0}\ g_{i,1}\cdots\ g_{i,\gamma-1})$ 表示 G_{qc} 生成矩阵中 P 矩阵的第 i 行

$M_i=[B^T_{0,i}.....B^T_{\gamma-1,i}]^T$ 表示 H_{qc} 循环矩阵的第 i 列；

$u=(1\ 0\cdots\ 0)$ 表示“1”位于开始的单位 $(q-1)$-重；

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}；

3) 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

a) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，其数量为 $\gamma(q-1)-r$；

b) 由 $z_i^T=D^{-1}M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

c) 在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*,g_1^*.....g^*_{\gamma(q-1)-r-1}$，以构成完整的

生成矩阵 G_{qc} 表达式：

$$G^*_{qc}=\begin{bmatrix}G\\-\\Q\end{bmatrix}=\left[\begin{array}{ccccccc}I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1}\\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M}\\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1}\\ \hline & & & & g^*_0 & & & \\ & & & & g^*_1 & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-\rho-1} & & & \end{array}\right]$$

而 g_i* 由求解下列矩阵等式获得：

$$\begin{bmatrix}B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1}\\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M}\\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1}\end{bmatrix}\begin{bmatrix}g^*_{i,(\rho-\gamma)(q-1)}\\ g^*_{i,(\rho-\gamma)(q-1)+1}\\ \mathrm{M}\\ g^*_{i,\rho(q-1)-1}\end{bmatrix}=0\text{。}$$

本发明的特点：

本发明是针对地面数字电视广播中各种特性对信道纠错编码的要求而提出了一种多码率纠错编码方法，此方法采用 BCH 码和 LDPC 码串行级联的方式，得到多种编码码率，以便适用不同的信道情况。相比于 DVB-S2 等系统中采用的很长（64800bit）的 LDPC 码，本发明提出的 LDPC 码具较短的码长，编解码实现结构简单，时延小，误码平台（error floor）低，性能优于 DVB-S2 LDPC 码。

本发明只需要有限次的迭代就能得到很好的纠错性能，提高了系统的抗误码性能，并且适用于多种传输调制方式（QPSK、16QAM 和 64QAM 等），可应用于地面、有线、卫星等数字电视广播，以及数字通信系统中。

附图说明

图 1 为 DVB-S2 纠错编码性能曲线

图 2 为 Gallagher 构造的一个（20,3,4）LDPC 码的校验矩阵

图 3 为 LDPC 码的双向图表示

图 4 为相同码长，但不同编码方式时 LDPC 码性能的比较

图 5 为相同编码方式，但不同码长时 LDPC 码性能的比较

图 6 为 TDS-OFDM 系统的帧结构

图 7 为本发明提出的纠错编码方法编码框图

图 8 为 LDPC 码和积译码算法信号流程图

图 9 为采用本发明的地面数字多媒体电视广播系统的发射端框图

图 10 为采用本发明的地面数字多媒体电视广播系统的接收端框图。

图 11 为采用本发明的地面数字多媒体电视广播系统的测试性能

具体实施方式

下面将结合附图对本发明的具体实施例进行详细描述。

清华 DMB-T 中采用的时域同步正交频分复用（TDS-OFDM）调制属多载波技术，但与欧洲 DVB-T 采用的编码的正交频分复用（COFDM）不同，在 TDS-OFDM 中没有插入频域导频信号，而是在 OFDM 的保护间隔中以时域的方式插入了伪随机（PN）序列，用于帧同步、频率同步、定时同步、信道传输特性估计和跟踪相位噪声等。

为了实现快速和稳定的同步，清华 TDS-OFDM 传输系统采用了分级帧结构。帧结构的基本单元称为信号帧，如图 6 所示。200/225 个信号帧定义为一个帧群，512 个帧群定义为一个超帧。帧结构的顶层称为日帧，由超帧组成。帧群中的每一个信号帧有唯一的帧号，它被编码在帧头的 PN 序列中。

TDS-OFDM 传输系统的信号帧使用时域同步的正交频分复用调制，或者称为以 PN 序列为保护间隔的正交频分复用调制。一个信号帧由帧同步和帧体两部分组成，它们具有相同的基带符号率 7.56 MS/s(1/T)。一个信号帧可以作为一个正交频分复用(OFDM)块。一个 OFDM 块进一步分成一个保护间隔和一个离散傅里叶逆变换(IDFT)块。对于 TDS-OFDM 来说，帧同步 PN 序列作为 OFDM 的保护间隔，而帧体作为 IDFT 块，长度为 3780 点，其中有 36 点用于 PN 长度、纠错方式、调制方式等系统参数，真正用于传输用户数据的净荷长度为 3744 点。因此，为了处理方便，纠错编码方法的长度与 3744 有密切关系，最好是它的整数倍，但同时还要满足纠错性能要求，在本发明中，我们所提出的纠错编码方法的长度为 3744 的 2 倍，为 7488。

关于 DMB-T、TDS-OFDM 的相关情况详见授权号为 00123597.4 名为“地面数字多媒体电视广播系统”、授权号为 01115520.5 名为“时域同步正交频分复用调制方法”、授权号为 ZL01130659.9 名为“地面数字多媒体电视广播系统中的帧同步产生方法”，以及授权号为 01124144.6 名为“正交频分复用调制系统中保护间隔的填充方法”等清华大学申请的中国发明专利。

下面我们就解释本发明提出的 LDPC 码的编码和解码的原理和步骤，它包括三种码率的 FEC 前向纠错编码：码率为 0.4 的 FEC(7488, 3008)、码率为 0.6 的 FEC(7488, 4512) 和码率为 0.8 的 FEC(7488, 6016)。

在本实施例中，FEC 前向纠错编码采用了 BCH(762,752) 和 LDPC 码的串行级联码，其中

LDPC 码采用了准循环低密度校验码（Quasi-cyclic LDPC Code，QC-LDPC），准循环纠错编码的概念如同前面技术背景中所述。本发明提出的 QC-LDPC 有三种：LDPC(7493, 3048)、LDPC(7493, 4752) 和 LDPC(7493, 6096)。删除这三种 QC-LDPC 级联码的前 5 个校验比特后，使得所提 LDPC 的码长由 7493 变为 7488。本发明所提出的完整的 BCH 和 LDPC 级联码的构造过程如下所示：

1) 编码码率为 0.4 的 FEC

数据码流→ BCH(762, 752) → LDPC(7493, 3048) → 删除前 5 个校验比特→LDPC(7488，3048)；

2) 编码码率为 0.6 的 FEC

数据码流→ BCH(762, 752) → LDPC(7493, 4572) →删除前 5 个校验比特→LDPC(7488，4512)；

3）编码码率为 0.8 的 FEC

数据码流→ BCH(762, 752) → LDPC(7493, 6096) →删除前 5 个校验比特→LDPC(7488，6096)。

从前面关于纠错编码的描述中可知，生成矩阵 G 或校验矩阵 H 是决定 LDPC 性能的关键因素之一，不同的生成矩阵 G 得到的纠错性能就不同，因此，寻找或设计 LDPC 码的生成矩阵 G 或校验矩阵 H 就成为 LDPC 码的重要问题。QC-LDPC 同样如此，一个好的设计的 QC-LDPC 误码性能可与随机的规则或非规则 LDPC 码相比。因此，下面我们首先描述本发明所提出的三个 QC-LDPC 码的校验矩阵 H 和生成矩阵 G，然后再描述本实施例中三种级联纠错码的实施步骤。

如前所述，循环矩阵是一个方阵，它的每一行和每一列都具有循环移位特性，并且它的行重量等于列重量，为 w。若 w＝1，那么就得到*循环置换矩阵（circulant permutation matrix）*。

QC-LDPC 码的校验矩阵 H_{qc} 是一个稀疏的循环置换矩阵，在本实施例中，H_{qc} 的构造采用了基于两信息符号的 RS 码方法，其过程如下所述：

第一步 生成 $RS(q-1, q-\rho, \rho-1)$ 码 C_{RS}

在一个伽逻华域 GF(p^s) 中，其中 p 为素数，s 是正整数，令 $q = p^s$，α 表示 GF(p^s) 一个本原元素，则 $\alpha^{\infty} = 0, \alpha^0 = 1, \alpha^1, \alpha^2, \alpha^3, ..., \alpha^{q-2}$ 构成了 GF(p^s) 的所有元素。令 ρ 为一个正整数，且 $2 \le \rho < q$，那么循环 $RS(q-1, q-\rho, \rho-1)$ 码 C_{RS} 的长度为 q-1，信息符号长度为 $q-\rho-1$，最小汉明距离为 $\rho-1$，其生成多项式为：

$$g(X)=(X-\alpha)(X-\alpha^2)\mathrm{L}\ (X-\alpha^{\rho-2})$$
$$=g_0+g_1X+g_2X^2+\mathrm{L}\ +X^{\rho-2}$$

其中 $g_i\in GF(p^s)$

第二步 从 C 中删除前 $q-\rho-1$ 信息符号得到缩短 $RS(\rho,2,\rho-1)$ 码，构造基码 C_b

把每个码字 C_{RS} 中的前 $q-\rho-1$ 个信息符号删除，把码字 C_{RS} 长度缩短，得到缩短 $RS(\rho,2,\rho-1)$ 码 C_b，它只有 2 个信息符号，其生成矩阵为：

$$G_b=\begin{bmatrix}g_0 & g_1 & g_2 & \mathrm{L} & 1 & 0\\ 0 & g_0 & g_1 & g_2 & \mathrm{L} & 1\end{bmatrix}$$

那么 C_b 非零码字具有两个不同的码重，$\rho-1$ 和 ρ。我们*称 C_b 为基码（Base Code）*。

第三步 构造码字集合 $C_b^{(i)}$

因为 C_b 的最小汉明距离为 $\rho-1$，所以 C_b 中任意两个码字至多在一个位置具有相同的码符号。我们需要利用 C_b 的一些结构特性，以便构造一类规则 LDPC 码，使得 LDPC 的双向图避免了长度为 4 的循环，这是规则 LDPC 码校验矩阵 H 设计的基本原则。

令 c 是 C_b 中的一个码字，其码重为 ρ，则 C_b 中 $q-1$ 个非零码字集合 $C_b^{(1)}=\{\beta c:\beta\in GF(p^s)\}$ 构成了 C_b 的一维子码，$C_b^{(1)}$ 中的每个非零码字具有码重 ρ。$C_b^{(1)}$ 中的两个码字在每个位置都是不同的。基于子码 $C_b^{(1)}$，通过循环右移一位，把 C_b 分割为 $q-1$ 个陪集（coset），$C_b^{(1)},C_b^{(2)},\mathrm{L}\ C_b^{(q-1)}$，那么任意陪集 $C_b^{(i)}$ 中两个码字在所有的位置都是不同的。

如果把陪集 $C_b^{(i)}$ 中的 $q-1$ 个非零码字表示为 $(q-1)\times\rho$ 阶矩阵，则矩阵中任意列的所有 $q-1$ 个元素都是不相同的。

第四步 由 $C_b^{(i)}$ 构造符号位置矢量集合 $Z(C_b^{(i)})$

令 $z=(z_0,z_1,\mathrm{L}\ ,z_{q-2})$ 是 GF(2) 上的 $(q-1)$-重（数组），它的分量成分对应于 $GF(p^s)$ 的 $q-1$ 个非零的元素，即 z_i 对应于非零的域元素 α^i。我们*称 α^i 为 z_i 的位置号码（Location Number）*。对于 $i=3,1,\mathrm{L}\ ,q-2$，我们*定义 α^i 的位置矢量（Location Vector）*为 GF(2) 上的 $(q-1)$-重，其中第 i 个分量 z_i 等于 1，其它所有的 $q-2$ 个分量等于 0。

令 $b=(b_0,b_1,\mathrm{L}\ ,b_{\rho-1})$ 为 C_b 中的码字，对于 $0\le j\le\rho-1$，用它的位置矢量 $z(b_j)$ 代替 b 的每个分量 b_j，则得到 GF(2) 上的 $\rho(q-1)$-重：

$$z(b)=(z(b_0),z(b_1),.....,z(b_{\rho-1}))$$

其码重为 ρ，*称为b的符号位置矢量*。因为 C_b 中的任意两个码字至多在一个位置具有相同的码符号，因而它们的符号位置矢量 z(b) 至多有 1 个“1-分量”是共有的。

令 $Z(C_b^{(i)}) = \{z(b) : b \in C_b^{(i)}\}$ 是 $C_b^{(1)}$ 的第 i 个陪集 $C_b^{(i)}$ 中的 $q-1$ 个码字的符号位置矢量的集合，根据前面 $C_b^{(1)}$ 的陪集 $C_b^{(i)}$ 的结构特性可知，$Z(C_b^{(i)})$ 中任意两个符号位置矢量没有任何“1-分量”是共有的或相同的。

第五步 构造 $Z(C_b^{(i)})$ 的符号位置矩阵 A_i

对于 $0 \le i \le q-2$，在 GF(2) 上构造一个 $(q-1)\times\rho(q-1)$ 阶矩阵 A_i，它的行是 $Z(C_b^{(i)})$ 中的 $q-1$ 个符号位置矢量。

因为 $Z(C_b^{(i)})$ 中每个矢量的重量为 ρ，所以 A_i 中“1-条目（entries）”的总数为 $\rho(q-1)$。因为 $Z(C_b^{(i)})$ 中两个符号位置矢量没有任何“1-分量”是相同的，所以 A_i 每列的重量为 1。因此，A_i 是一个 $(1,\rho)$- 规则矩阵，其行、列重量各为 ρ 和 1。根据 C_b 中码字的符号位置矢量的定义，以及每个陪集 $C_b^{(i)}$ 中码字的结构特点，A_i 是一个 $(q-1)\times\rho(q-1)$ 阶置换矩阵，*矩阵 A_i 称为符号位置矩阵（Symbol Loacation Matrix）*。

符号位置矩阵 A_i 构成的类（class），$\boldsymbol{A} = \{A_0, A_1, \ldots, A_{q-2}\}$，具有如下结构特性：(1) 相同的矩阵 A_i 中没有两行具有共有的“1-分量”；(2) 来自两个不同的矩阵 A_i 和 A_j 的两行共有的“1-分量”不会超过 1。

第六步 构造一个 LDPC 码校验矩阵的基矩阵 $H_{GA}(\gamma,\rho)$

令 γ 是正整数，且 $1 \le \gamma \le (q-1)$，构造下述 GF(2) 的 $\gamma(q-1)\times\rho(q-1)$ 阶的*基矩阵（Base Matrix）*：

$$H_{GA}(\gamma,\rho) = \begin{bmatrix} A_0 \\ \mathrm{M} \\ A_{\gamma-1} \end{bmatrix} = \begin{bmatrix} A_{0,0} & A_{0,1} & \mathrm{L} & A_{0,\rho-1} \\ A_{1,0} & A_{1,1} & \mathrm{L} & A_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & A_{i,j} & \mathrm{M} \\ A_{\gamma-1,0} & A_{\gamma-1,1} & \mathrm{L} & A_{\gamma-1,\rho-1} \end{bmatrix}$$

这个矩阵是一个 (γ,ρ)- 规则矩阵，其行、列重量各为 ρ 和 γ，它具有其成员矩阵 A_i 的结构特点，即 $H_{GA}(\gamma,\rho)$ 任意两行（或两列）至多有 1 个“1-分量”是共有的。这表示 $H_{GA}(\gamma,\rho)$ 中矩形四个角上不可能有 4 各“1”，这将避免 $H_{GA}(\gamma,\rho)$ 的双向图具有长度为 4 的循环，并且它的周长至少为 6。

这个 $H_{GA}(\gamma,\rho)$ 矩阵是规则 Gallager LDPC 生成矩阵的典型形式，其生成的 LDPC 码长为 $n = \rho(q-1)$，码率至少为 $(\rho-\gamma)/\rho$。

因为 $H_{GA}(\gamma,\rho)$ 两行中“1-分量”的共有数量不超过 1，矩阵每列的重量为 γ，因此 $H_{GA}(\gamma,\rho)$ 的最小码距 $d_{min}(\gamma)$ 至少为 γ-1。如果考虑校验矩阵 $H_{GA}(\gamma,\rho)$ 的结构，最小码距下

限 $d_{min}(\gamma)$ 可能那改变：

$$d_{min}(\gamma) \geq \begin{cases} \gamma+1, & \text{对于 odd } \gamma \\ \gamma+2, & \text{对于 even } \gamma \end{cases}$$

第七步 构造一个掩蔽矩阵 W(γ，ρ)

上述生成的基矩阵是一个(γ，ρ)- 规则 QC-LDPC 码的校验矩阵，通过掩蔽所选择的 H_{GA}(γ，ρ) 循环置换矩阵的集合，可以得到各种 QC-LDPC 码的校验矩阵 H_{qc}(γ，ρ)。掩蔽操作可以通过掩蔽矩阵 W(γ，ρ) 与 H_{GA}(γ，ρ) 矩阵相乘获得。

如果掩蔽矩阵 W(γ，ρ) 是规则矩阵，行和/或列具有相同的码重，则 H_{qc}(γ，ρ) 是规则矩阵，获得的码字是规则 QC-LDPC 码；如果 W(γ，ρ) 具有可变的列重量和/或可变的行重量，则 H_{qc}(γ，ρ) 是非规则矩阵，得到的码字是非规则 QC-LDPC 码。

掩蔽矩阵 W(γ，ρ) 是 GF(2) 上的 γ × ρ 阶矩阵，γ，ρ 通常较小，一般小于 500。

具有很大周长的掩蔽矩阵可以通过计算机搜索来构造，参加下列文献：

(11) X. -Y Hu, E. Eleftheriou, and D. -M. Arnold, "Progressive edge-growth Tanner graphs," Proc. 2001 GlobeCom, pp. 995-1001, Nov. 25-29, 2002.

(12) L. Lan, Y. Y. Tai, L. Chen, S. Lin and K. Abdel-Ghaffar, "A trellis-based method for removing cycles from bipartite graphs and construction of LDPC codes," IEEE Communicat ion Letters, vol.8,no.7, July 2004.

规则的掩蔽矩阵 W(γ，ρ) 可以通过代数法构造，例如循环矩阵分解、几何分解、平衡非完备块设计、叠加、乘积图、汉明码的最小码重码字等。

而本实施例采用了非规则的掩蔽矩阵 W(γ，ρ)，非规则掩蔽矩阵 W(γ，ρ)的行、列分别具有不等的重量。因而，它的双向图具有变化的变量节点度和/或变化的校验节点度。这两类节点的度分布表示为两个多项式形式：

$$v(X) = \sum_{i=1}^{d_v} v_i X^{i-1}$$

$$c(X) = \sum_{i=1}^{d_c} c_i X^{i-1}$$

其中 v_i 和 c_i 分别表示度为 i 的双向图中变量节点和校验节点的小数，d_v 和 d_c 分别表示变量节点和校验节点的最大度。

因为双向图的变量节点和校验节点对应于校验矩阵 H 的列和行，因此，v(X) 和 c(X) 也就给出了校验矩阵 H 的列和行重量分布。人们已经证明非规则 LDPC 码的误码性能依赖于它的双向图中变量节点和校验节点的度分布，通过仿真 BP 算法中这两类节点之间消息传递的概率密度分布，优化这两个度分布 v(X) 和 c(X)，可以设计逼近仙农限的 LDPC 码。

根据所要求的码字长度 n 和编码码率 R，选择合适的 γ、ρ 和 q。

在码构造中，一旦设计了度分布 v(X) 和 c(X)，就可以根据度分布通过计算机以随机方式确定 W(γ，ρ)，使得 W(γ，ρ) 的列和行的重量分布等于（或接近）度分布 v(X) 和 c(X)。

第八步 生成 QC-LDPC 的校验矩阵 $H_{qc}=W\times H_{GA}$

上述构造的掩蔽矩阵 $W(\gamma,\rho)=[w_{i,j}]$ 是一个 GF(2) 上的 γ × ρ 阶矩阵，QC-LDPC 码的校验矩阵 $H_{qc}(\gamma,\rho)$ 通过下列矩阵乘法获得：

$$H_{qc}(\gamma,\rho)=W(\gamma,\rho)\times H_{GA}(\gamma,\rho)$$

$$=\begin{bmatrix} w_{1,1}A_{1,1} & w_{1,2}A_{1,2} & \mathrm{L} & w_{1,\rho}A_{1,\rho} \\ w_{2,1}A_{2,1} & w_{2,2}A_{2,2} & \mathrm{L} & w_{2,\rho}A_{2,\rho} \\ \mathrm{M} & \mathrm{M} & w_{i,j}A_{i,j} & \mathrm{M} \\ w_{\gamma,1}A_{\gamma,1} & w_{\gamma,2}A_{\gamma,2} & \mathrm{L} & w_{\gamma,\rho}A_{\gamma,\rho} \end{bmatrix}$$

$$=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix}$$

其中

$$B_{i,j}=w_{i,j}A_{i,j}=\begin{cases} A_{i,j}, & \text{对于 } w_{i,j}=1 \\ 0, & \text{对于 } w_{i,j}=0 \end{cases}$$

即通过与 W(γ，ρ) “0”项的相乘，简单地掩盖掉 $H_{GA}(\gamma,\rho)$ 的某些置换矩阵集合 $A_{i,j}$。$H_{qc}(\gamma,\rho)$ 中置换矩阵 $A_{i,j}$ 的分布等同于 $W(\gamma,\rho)$ 中“1”项的分布。因为掩蔽减少了 $H_{GA}(\gamma,\rho)$ 中“1”的密度，即掩蔽操作保持了 RC 行列约束，因而 $H_{qc}(\gamma,\rho)$ 也满足行列约束，而且 $H_{qc}(\gamma,\rho)$ 的双向图短循环的数目要少于 $H_{GA}(\gamma,\rho)$ 的。如果 $W(\gamma,\rho)$ 双向图的周长大于等于 6，则 $H_{qc}(\gamma,\rho)$ 的周长至少为 6。

在本实施例中，基于 2 信息符号的 RS 码生成的 QC-LDPC 校验矩阵 $H_{GA}(\gamma,\rho)$ 是规则，但掩蔽矩阵 W(γ，ρ) 是非规则矩阵。所以，两者矩阵相乘后得到的 QC-LDPC 最终的校验矩阵是非规则的，即本实施例中采用的 QC-LDPC 码是非规则的。

综合上述步骤，QC-LDPC 码最终的校验矩阵 $H_{qc}(\gamma,\rho)$ 具有形式如下：

$$H_{qc}(\gamma,\rho)=\begin{bmatrix} B_{0,0} & B_{0,1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,0} & B_{1,1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & B_{i,j} & \mathrm{M} \\ B_{\gamma-1,0} & B_{\gamma-1,1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix} \quad (1)$$

其中，$B_{i,j}$ 是一个 $(q-1)\times(q-1)$ 阶矩阵，$0\le i\le\gamma-1$ 和 $0\le j\le\rho-1$。相比于它的大小 $q-1$，重量 w 是非常小的。对于 QC-LDPC 码，本发明中采用了 w=1。如同上述定义，$B_{i,j}$

是循环置换矩阵。

令对于矩阵（1）中的 $B_{i,j}=k$，$0\le k\le q-2$，它意味着这个循环置换矩阵的第 0 行中，从左到右各列标识为 0 到 q-1，其中只有第 k 列为“1”，其它 q-2 列都为“0”。其余行（第 1 行到第 q-2 行）中，每一行是其上一行的循环移位（向右移一位），第 0 行实际上是第 q-2 行的循环移位。因此，这种循环置换矩阵 $B_{i,j}$ 的构造完全可由“k”值来描述。例如，如果 k=3，q=8，则 $B_{i,j}=3$ 循环置换矩阵如下所示：

$$B_y=\begin{bmatrix}0&0&0&1&0&0&0\\0&0&0&0&1&0&0\\0&0&0&0&0&1&0\\0&0&0&0&0&0&1\\1&0&0&0&0&0&0\\0&1&0&0&0&0&0\\0&0&1&0&0&0&0\end{bmatrix}$$

本实施例中提出的三种 QC-LDPC 码的校验矩阵的参数分别为：

1）LDPC (7493, 3048) 码的校验矩阵 H_{qc1} 的格式如上式（1）所示，其中 γ=35，ρ=59 和 $q=p^s=2^7=128$，码长 $n=\rho\times(q-1)=59\times127=7493$，信息比特长度 $k=(\rho-\gamma)\times(q-1)=24\times127=3048$。

2）LDPC (7493, 4572) 码的校验矩阵 H_{qc2} 的格式如上式（1）所示，其中 γ=23，ρ=59 和 q=128，码长 $n=\rho\times(q-1)=59\times127=7493$，信息比特长度 $k=(\rho-\gamma)\times(q-1)=4572$。

3）LDPC (7493, 6096) 码的校验矩阵 H_{qc3} 的格式如上式（1）所示，其中 γ=11，ρ=59 和 q=128，码长 $n=\rho\times(q-1)=59\times127=7493$，信息比特长度 $k=(\rho-\gamma)\times(q-1)=6096$。

上面描述了本实施例所提出的 QC-LDPC 码收端解码所需要的校验矩阵 H_{qc} 的生成过程，但在发送端，QC-LDPC 码编码需要一个生成矩阵 G_{qc}，下面就描述如何有循环形式的校验矩阵 H_{qc} 获得系统循环形式的生成矩阵 G_{qc}。

由校验矩阵 H_{qc} 给出的 QC-LDPC 码表示为 C_{qc}，假设 H_{qc} 的秩 r 等于 H_{qc} 矩阵的行数，即 $r=\gamma(q-1)$，并且 Hqc 的前 $(\rho-\gamma)\times(q-1)$ 比特对应着信息比特。我们把 H_{qc} 的循环矩阵的列重新排列，使得下列 H_{qc} 的 $\gamma\times\rho$ 阶子矩阵的秩为满秩 $\gamma(q-1)$，与 H_{qc} 的秩 r 相同：

$$D=\begin{bmatrix}B_{0,\rho-\gamma}&B_{0,\rho-\gamma+1}&\mathrm{L}&B_{0,\rho-1}\\B_{1,\rho-\gamma}&B_{1,\rho-\gamma+1}&\mathrm{L}&B_{1,\rho-1}\\\mathrm{M}&\mathrm{M}&\mathrm{L}&\mathrm{M}\\B_{\gamma-1,\rho-\gamma}&B_{\gamma-1,\rho-\gamma+1}&\mathrm{L}&B_{\gamma-1,\rho-1}\end{bmatrix}\quad(2)$$

要求的 QC-LDPC 码 C_{qc} 的生成矩阵 G_{qc} 具有如下的系统循环置换矩阵形式：

$$G_{qc}=\begin{bmatrix} G_0 \\ G_1 \\ \mathrm{M} \\ G_{\rho-\gamma-1} \end{bmatrix}$$

$$=\begin{bmatrix} I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1} \\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M} \\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1} \end{bmatrix}$$

$$=\begin{bmatrix} I_{(\rho-\gamma)(q-1)} & P \end{bmatrix}$$

其中，I是$(q-1)\times(q-1)$阶单位矩阵，O是$(q-1)\times(q-1)$阶零阵，而$G_{i,j}$是$(q-1)\times(q-1)$阶循环矩阵，$0\le i\le\rho-\gamma-1$，$0\le j\le\gamma-1$。

上式形式的生成矩阵G_{qc}是所谓的系统循环（Systematic Circulant, SC）形式，它由两部分组成，左边部分 I 是主对角线为 $(\rho-\gamma)(q-1)\times(q-1)$阶单位矩阵，右边部分 P 是$(q-1)\times(q-1)$阶循环矩阵构成的 $(\rho-\gamma)\times\gamma$ 阶阵列。由编码理论可知，G_{qc}是系统码形式，其左边部分P称为P矩阵，对应于系统码的校验位。P矩阵是循环矩阵的阵列，具有SC形式。

G_{qc}是C_{qc}生成矩阵的充要条件是 $H_{qc}G_{qc}^{T}=|O|$，这里|O|是$\gamma(q-1)\times(\rho-\gamma)(q-1)$ 阶零阵。对于$0\le i\le\rho-\gamma-1$和 $0\le j\le\gamma-1$，令 $g_{i,j}$ 表示循环矩阵 $G_{i,j}$ 的生成多项式。我们一旦知道了这些$g_{i,j}$，就可以构造G_{qc}的所有循环矩阵$G_{i,j}$。因此，Gqc的特性完全可以由 $\gamma\times(\rho-\gamma)$阶循环矩阵生成式的集合所表征，称为QC-LDPC码C_{qc}的生成矩阵。

令 $u=(1\ 0\ \mathrm{K}\ 0)$ 表示“1”位于开始的单位 $(q-1)$-重，而$0=(0\ 0\mathrm{K}\ 0)$表示全零$(q-1)$-重。对于 $0\le i\le\rho-\gamma-1$，G_i的子矩阵的第i行是：

$$g_i=(0\ \mathrm{K}\ \mathbf{0}\ \boldsymbol{u}\ \mathbf{0}\ \mathrm{K}\ 0\ g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\gamma-1})$$

其中，单位 $(q-1)$-重u位于g_i的第i位。

若G_{qc}是C_{qc}的生成矩阵，则必然有 $H_{qc}g_i^{T}=0$。令 $z_i=(g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\gamma-1})$ 表示G_{qc}生成矩阵中P矩阵的第i行，$M_i=[B^{T}_{0,i}.....B^{T}_{\gamma-1,i}]^{T}$ 表示H_{qc}循环矩阵的第i列。那么 $H_{qc}g_i^{T}=0$变为下列等式：

$$M_i u^{T}+Dz_i^{T}=0$$

因为上式中D是前面定义的方阵，它是满秩的、非奇异的，并且逆矩阵D^{-1}存在。

因此，由上式得：

$$z_i^T = D^{-1}M_i u^T \qquad (3)$$

此式中 D、M 和 u 都是已知量，求解此方程，得到 $z_0, z_1, \ldots\ldots, z_{\gamma-1}$。因为 $z_i = (g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\gamma-1})$，因此，从 $z_0, z_1, \ldots\ldots, z_{\gamma-1}$ 可以获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有的生成多项式 $g_{i,j}$，即得到了 QC-LDPC 码的生成矩阵 G_{qc}。

在上面描述的由 H_{qc} 求解 G_{qc} 的过程中，我们假设 H_{qc} 的秩 r 等于 H_{qc} 矩阵的行数，即 $r = \gamma(q-1)$，但也可能存在 $r < \gamma(q-1)$ 的情况，我们也需要处理这种情况下如何求解 G_{qc}。

在这种情况下，仍然使 $\gamma \times \rho$ 阶 D 矩阵的秩等于 H_{qc} 的秩 r，但 $r < \gamma(q-1)$，这表明 D 不是满秩的，某些列之间存在线性相关性。为了使上述的式（3）

$$z_i^T = D^{-1}M_i u^T$$

有唯一的解，我们把 z_i 中 $\gamma(q-1)-r$ 个线性相关的元素设为 0，这些元素对应于 D 矩阵中线性相关的列。那么我们就可以由上式获得 z_i 中 r 个线性独立的元素。

由此，我们得到了 $\gamma(\rho-\gamma)$ 个循环矩阵 $G_{i,j}$ 的 $\gamma(\rho-\gamma)$ 个生成多项式 $g_{i,j}$。由这些生成式出发，可以构造一个系统循环结构的生成矩阵 G，此矩阵生成 C_{qc} 的子码，维数为 $(\rho-\gamma)(q-1)$。为了生成全部码字 C_{qc}，必须在 G 添加 $\gamma(q-1)-r$ 个线性独立的行，以构成 C_{qc} 完整的生成矩阵 G_{qc}。

令 $g_0^*, g_1^*, \ldots\ldots, g^*_{\gamma(q-1)-r-1}$ 表示添加的那些行。对于 $0 \le i < \gamma(q-1)$，第 i 个添加行的构成形式为：

$$g_i^* = (0\mathrm{K}\ 0\mathrm{K}\ 0\ g^*_{i,(\rho-\gamma)(q-1)}\ g^*_{i,(\rho-\gamma)(q-1)+1}\mathrm{K}\ g^*_{i,\rho(q-1)-1})$$

它具有下列结构特性：

(1) 前 $(\rho-\gamma)(q-1)$ 比特为 0；

(2) $z_i = (g^*_{i,(\rho-\gamma)(q-1)}\ g^*_{i,(\rho-\gamma)(q-1)+1}\mathrm{K}\ g^*_{i,\rho(q-1)-1})$ 的 $\gamma(q-1)-r$ 个比特对应于 D 中线性相关的列，*称为相关比特（Dependent bit）*，相关比特具有（0, ..., 0, 1, 0, ..., 0）的形式，“1”表示第 i 个相关比特的位置。

同上面一样，G 中添加的行 g_i^* 也必须满足等式 $H_{qc}g_i^{*T} = 0$，$0 \le i < \gamma(q-1)-r-1$，从而得到下列矩阵：

$$\begin{bmatrix} B_{0,\rho-\gamma} & B_{0,\rho-\gamma+1} & \mathrm{L} & B_{0,\rho-1} \\ B_{1,\rho-\gamma} & B_{1,\rho-\gamma+1} & \mathrm{L} & B_{1,\rho-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{L} & \mathrm{M} \\ B_{\gamma-1,\rho-\gamma} & B_{\gamma-1,\rho-\gamma+1} & \mathrm{L} & B_{\gamma-1,\rho-1} \end{bmatrix} \begin{bmatrix} g^*_{i,(\rho-\gamma)(q-1)} \\ g^*_{i,(\rho-\gamma)(q-1)+1} \\ \mathrm{M} \\ g^*_{i,\rho(q-1)-1} \end{bmatrix} = 0 \qquad (4)$$

解此矩阵，得到所有的 g_i^*。

在 G 中添加 g^*_0，g^*_1K $g^*_{\gamma(q-1)-r-1}$ 行，得到此种情况下 QC-LDPC 码 C_{qc}的生成矩阵：

$$G^*_{qc}=\begin{bmatrix}G\\-\\Q\end{bmatrix}=\left[\begin{array}{cccccccc}I & O & \mathrm{L} & O & G_{0,0} & G_{0,1} & \mathrm{L} & G_{0,\gamma-1}\\ O & I & \mathrm{L} & O & G_{1,0} & G_{1,1} & \mathrm{L} & G_{1,\gamma-1}\\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & \mathrm{M} & \mathrm{M} & G_{i,j} & \mathrm{M}\\ O & O & \mathrm{L} & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \mathrm{L} & G_{\rho-\gamma-1,\gamma-1}\\ \hline & & & & g^*_0 & & & \\ & & & & g^*_1 & & & \\ & & & & \mathrm{M} & & & \\ & & & & g^*_{\gamma(q-1)-r-1} & & & \end{array}\right] \quad (5)$$

G^*_{qc} 所有的行都是线性独立的，因而其秩为 $\rho(q-1)-r$。

令 v 表示要编码的信息比特，则相应的 QC-LDPC 码字 C_{qc}为：

$$C_{qc}=v\cdot G_{qc}$$

或

$$C_{qc}=v\cdot G^*_{qc}$$

综上所述，在本实施例中所提出的 QC-LDPC 码 C_{qc} 的生成矩阵 G_{qc} 的构造步骤为：

第一步 由 H_{qc} 构造矩阵 D，D 是方阵、满秩、非奇异，即逆矩阵 D^{-1} 存在，其秩与 H_{qc} 的秩相同；

第二步 根据编码理论，由等式 $M_i u^T+Dz_i^T=0$ 得式（3）：

$$z_i^T=D^{-1}M_i u^T$$

其中

$$z_i=(g_{i,0}\ g_{i,1}\mathrm{K}\ g_{i,\gamma-1})$$

$$M_i=[B^T_{0,i}.....B^T_{\gamma-1,i}]^T$$

$$u=(1\ \ 0\ \mathrm{K}\ \ 0)$$

从而获得 G_{qc} 中循环矩阵 $G_{i,j}$ 的所有的生成多项式 $g_{i,j}$，即得到了矩阵 D 满秩情况下的 QC-LDPC 码的生成矩阵 G_{qc}。

第三步 在矩阵 D 非满秩情况时，对 G_{qc} 构造的修正：

1) 把 z_i 中对应于 D 矩阵中线性相关列的元素设为 0，线性相关列的数量为 $\gamma(q-1)-r$；

2) 由 $z_i^T=D^{-1}M_i u^T$ 求得 z_i 中 r 个线性独立的元素；

3）在 G_{qc} 中添加 $\gamma(q-1)-r$ 个线性独立的行，$g_0^*, g_1^*, \ldots\ldots, g_{\gamma(q-1)-r-1}^*$，以构成完整的生成矩阵 G_{qc} 表达式（5），而 g_i^* 由矩阵等式（4）获得。

经过上述步骤，最后得到 QC-LDPC 码 $C_{q,c}$ 的生成矩阵 G_{qc} 如下矩阵所示：

$$G_{qc}=\begin{bmatrix} I & O & \cdots & O & G_{0,0} & G_{0,1} & \cdots & G_{0,\gamma-1} \\ O & I & \cdots & O & G_{1,0} & G_{1,1} & \cdots & G_{1,\gamma-1} \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & G_{i,j} & \vdots \\ O & O & \cdots & I & G_{\rho-\gamma-1,0} & G_{\rho-\gamma-1,1} & \cdots & G_{\rho-\gamma-1,\gamma-1} \end{bmatrix} \quad (6)$$

其中，I 是 $(q-1)\times(q-1)$ 阶单位矩阵，O 是 $(q-1)\times(q-1)$ 阶零阵，而 $G_{i,j}$ 是 $(q-1)\times(q-1)$ 循环矩阵，令 $m=\rho-\gamma$，$0\le i\le m-1$，$0\le j\le \gamma-1$。

本实施例中提出的三种 QC-LDPC 码的生成矩阵 G_{qc} 的参数分别为：

1）LDPC(7493,3048) 码的生成矩阵 G_{qc1} 具有上述式（6） 所示的矩阵形式，其中参数 m=24，γ=35 和 q=128。

2）LDPC(7493,4572) 码的生成矩阵 G_{qc2} 具有上述式（6） 所示的矩阵形式，其中参数 m=36，γ=23 和 q=128。

3）LDPC(7493,6096) 码的生成矩阵 G_{qc3} 具有上述式（6） 所示的矩阵形式，其中参数 m=48，γ=11 和 q=128。

在上面描述的校验矩阵 H_{qc} 和生成矩阵 G_{qc} 构造方法的基础上，下面描述本实施例中提出的三种前向级联纠错码（BCH + QC-LDPC）的完整实施步骤。

本发明的纠错编码方法用于地面数字电视广播发端的实施例一中，FEC 前向纠错编码采用了 BCH(762,752) 码和 LDPC(7488, 3048) 码的串行级联，如图 7 所示，其实现步骤如下：

1）将输入的信息流每 2 个 TS 传递码流包分为一组，长度为 3008 比特，即 188byte×2=3008bit，然后再进一步分为 4 个子组，每一子组的长度为 752 比特；

2）对 4 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的输出长度为 4×762=3048 比特，其中 BCH(762，752) 是 BCH（1023，1013）的缩短码，其生成多项式为 $x^{10}+x^3+1$，缩短码按以下步骤生成：

a) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

b) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，得到 1023 个纠错编码比特；

c) 去掉纠错编码比特的前 261 个比特，因为 BCH 码是系统码，去掉的 261 个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}；

3） 对 4 个 BCH(762,752) 码 C_{BCH} 输出的 3048 个比特进行 LDPC(7493,3048) 编码，即与 LDPC(7493,3048) 的生成矩阵 G_{qc1} 相乘，得到 QC-LDPC (7493,3048) 码 C_{qc1}；

4）删除 LDPC(7493,3048) 码 C_{qc1} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出比特；

5）对得到的 7488 个纠错编码比特进行相应的 QPSK 或 mQAM 符号映射；

6）进行时域同步正交频分复用（TDS-OFDM）调制，然后再经过成形滤波、基带信号帧上变换等处理后送给信道传输。

本发明的纠错编码方法用于地面数字电视广播发端的实施例二中，FEC 前向纠错编码采用了 BCH(762,752) 码和 LDPC(7488, 4572) 码的串行级联，其框图类似于图 7，具体参数有差异，其实现步骤如下：

1）将输入的信息流每 3 个 TS 传递码流包分为一组，长度为 4512 比特，即 188byte×3=4512bit，然后再进一步分为 6 个子组，每一子组的长度为 752 比特；

2）对 6 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的输出长度为 6×762=4572 比特，其中 BCH(762，752) 是 BCH（1023，1013）的缩短码，其生成多项式为 $x^{10}+x^3+1$，缩短码按以下步骤生成：

a) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

b) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，得到 1023 个纠错编码比特；

c) 去掉纠错编码比特的前 261 个比特，因为 BCH 码是系统码，去掉的 261 个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}；

3） 对 6 个 BCH(762,752) 码 C_{BCH} 输出的 4572 个比特进行 LDPC(7493,4572) 编码，即与 LDPC(7493, 4572) 的生成矩阵 G_{qc2} 相乘，得到 QC-LDPC (7493, 4572) 码 C_{qc2}；

4）删除 LDPC(7493,4572)码 C_{qc2} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出比特；

5）对得到的 7488 个纠错编码比特进行相应的 QPSK 或 mQAM 符号映射；

6）进行时域同步正交频分复用（TDS-OFDM）调制，然后再经过成形滤波、基带信号帧上变换等处理后送给信道传输。

除了上述两个实施例外，本发明的纠错编码方法用于地面数字电视广播发端的实施例三中，FEC 前向纠错编码采用了 BCH(762,752) 码和 LDPC(7488, 6096)码的串行级联，其框图类似于图 7，具体参数有差异，其实现步骤如下：

1）将输入的信息流每 4 个 TS 传递码流包分为一组，长度为 6016 比特，即 188byte×4=6016bit，然后再进一步分为 8 个子组，每一子组的长度为 752 比特；

2）对 6 个子组的 752 个信息比特分别进行二进制 BCH(762，752) 纠错编码，得到纠错编码后的输出长度为 8×762=6096 比特，其中 BCH(762，752) 是 BCH（1023，1013）的缩短码，其生成多项式为 $x^{10}+x^3+1$，缩短码按以下步骤生成：

a) 在 752 个数据比特前增加 261 个“0”比特，组成 1013 比特的信息；

b) 对这 1013 个信息比特进行 BCH(1023,1013) 纠错编码，得到 1023 个纠错编码比特；

c) 去掉纠错编码比特的前 261 个比特，因为 BCH 码是系统码，去掉的 261 个比特是“0”比特，最后得到 BCH (762,752) 码 C_{BCH}；

3） 对 8 个 BCH(762,752) 码 C_{BCH} 输出的 6096 个比特进行 LDPC(7493,6096) 编码，即与 LDPC(7493, 6096) 的生成矩阵 G_{qc3} 相乘，得到 QC-LDPC (7493, 6096) 码 C_{qc3}；

4）删除 LDPC(7493,6096) 码 C_{qc3} 输出的 7493 个比特的前 5 个校验比特，得到 7488 个输出比特；

5）对得到的 7488 个纠错编码比特进行相应的 QPSK 或 mQAM 符号映射；

6）进行时域同步正交频分复用（TDS-OFDM）调制，然后再经过成形滤波、基带信号帧上变换等处理后送给信道传输。

在接收端，接收到的信号经过解调后，首先进行 LDPC 码译码，然后送给 BCH(762,752) 译码，得到纠错后的发送码流。其中 BCH 译码非常成熟，本实施例中没有采用特殊的算法，因此，在此就不再赘述了。

目前有各种不同的 LDPC 译码算法。常用的 LDPC 解码算法是比特翻转（Bit-flipping，BF）算法、加重比特翻转（Weighted BF）算法和基于信度传递的迭代译码（Iterative Decoding Based on Belief Progagation，IDBP）算法等三种，其中 IDBP 也就是和积算法（Sum-Product Algorithm，SPA）。

对于本发明提出的 LDPC 码，根据不同的要求，可以采用不同的 LDPC 解码算法。在上述三个实施例中，我们采用了基于信度传递的迭代译码算法，即和积算法。

BP (belief-propagation，信度传播) 算法，是与双向图紧密结合的，如图 3 所示，各边上传递的消息分为 check to variable 和 variable to check 两种。假设比特 '0' 映射为+1，比特 '1' 映射为-1。这两种消息的定义如下：

✓check to variable： R_{mn}^{a} (a=+1或a=-1) 是校验节点 z_m 向变量节点 x_n 传递的校验信息，表示 z_m 根据除 x_n 外其他相邻变量节点 $\{x_k : k\neq n, k\in N(m)\}$ 当前状态向 x_n 宣称的“x_n=a 使 z_m 满足”的可信度；

✓variable to check： Q_{mn}^{a} (a=+1或a=-1) 是变量节点 x_n 向校验节点 z_m 传递的变量信息，表示 x_n 根据除 z_m 外其他相邻校验节点 $\{z_l : l\neq m, l\in M(n)\}$ 和初始概率消息 P_n^a 向 z_m 宣称的“x_n=a”的可信度。

基于“信度传播”的迭代算法的过程如下：

①初始化：计算初始消息 $\{P_n^a\}$ 和初始化 $\{R_{mn}^a\}$ 值；

② 消息传递迭代和更新：

阶段1：各个 P_n 和 z_m 节点向其所有相邻变量节点 x_n 分别传递消息 P_n^a 和 R_{mn}^a，而各个 x_n 节点根

据接收消息更新Q_{mn}^{a}；

阶段2：各个x_n节点向其所有相邻校验节点z_m传递已更新的消息Q_{mn}^{a}，而各个z_m节点根据接收消息更新R_{mn}^{a}。

③ 迭代终止：x_n节点根据所有接收消息计算最终消息Q_n^a。采取的迭代终止策略为：

✓每轮迭代结束时,计算最终消息并根据不同的消息量度判决$\hat{x}$，计算伴随式$\hat{x}\cdot H^T$；

✓如果伴随式为零向量，则宣告成功，终止迭代；否则取前一次计算得出的值，转到上述第二个步骤继续迭代。若迭代次数达到预先设定的最大次数$I_{\max}$仍未成功，则根据最终消息判决得x,终止迭代，译码不成功。

本实施例中的LDPC译码使用了基于上述“信度传播”迭代原理的和积算法。LDPC与和积算法有一种天然的对应，译码的每次迭代都是一次“求和”与“乘积”运算的实现。

根据上述信度传播算法中队消息传递和更新过程的描述，在概率测度下，变量消息定义为：

$$Q_{mn}^{a}=P(x_n=a\mid z_m,\{z_l:l\subset M(n)\setminus m\})$$

定义校验消息为：

$$R_{mn}^{a}=P(z_m\mid x_n=a),$$

设码字先验等概，则初始概率消息P_n^a表示$x_n=a$的先验消息，R_{mn}^{a}初始化为1，$a\in\{+1,-1\}$。和积算法的具体步骤如图8所示：

① 初始化

将x_n的似然值赋给P_n^a。例如，在加性白色高斯噪声信道AWGN下，似然值为：

$$P_n^a=\frac{1}{1+\exp(\frac{-2ay_n}{\sigma^2})}$$

② check to variable（从校验节点到变量节点）信息：

$$\begin{aligned}R_{mn}^{a}&=P(z_m\mid x_n=a)\\&=\sum_{x:x_n=a}P(z_m,x\mid x_n=a)\\&=\sum_{x:x_n=a}P(z_m\mid x_n=a,x)P(x\mid x_n=a)\\&=\sum_{x:x_n=a}P(z_m\mid x)P(x\mid x_n=a)\\&=\sum_{x:x_n=a}P(z_m\mid x)\prod_{k\in N(m)\setminus n}P(x_k)\\&=\sum_{x:x_n=a}P(z_m\mid x)\prod_{k\in N(m)\setminus n}Q_{mk}^{x_k}\end{aligned}$$

例如假设第m个校验方程为$x_2\times x_3\times x_5=+1$。计算$x_2$=+1时满足这个校验方程的概率$R_{m2}^{0}$：

$$R_{m2}^{+1}=P\{x_2\times x_3\times x_5=+1\mid x_2=+1,x_3=+1,x_5=+1\}\cdot Q^{+1}{}_{m3}\cdot Q^{+1}{}_{m5}$$
$$+P\{x_2\times x_3\times x_5=+1\mid x_2=+1,x_3=+1,x_5=-1\}\cdot Q^{+1}{}_{m3}\cdot Q^{-1}{}_{m5}$$
$$+P\{x_2\times x_3\times x_5=+1\mid x_2=+1,x_3=-1,x_5=+1\}\cdot Q^{-1}{}_{m3}\cdot Q^{+1}{}_{m5}$$
$$+P\{x_2\times x_3\times x_5=+1\mid x_2=+1,x_3=-1,x_5=-1\}\cdot Q^{-1}{}_{m3}\cdot Q^{-1}{}_{m5}$$

其中的条件概率 P{}，如果满足校验方程则为 1，不满足为 0。同理，也可以计算 R_{m2}^{-1}。

⑤ variable to check（从变量节点到校验节点）信息：

$$Q_{mn}^{a}=P(x_n=a\mid y_n,\{z_l:l\in M(n)\setminus m\})$$
$$=\frac{P(x_n=a,y_n,\{z_l:l\in M(n)\setminus m\})}{P(y_n,\{z_l:l\in M(n)\setminus m\})}$$
$$=\frac{P(x_n=a\mid y_n)P(\{z_l:l\in M(n)\setminus m\}\mid y_n,x_n=a)}{P(\{z_l:l\in M(n)\setminus m\}\mid y_n)}$$
$$=\frac{P(x_n=a\mid y_n)\prod_{l\in M(n)\setminus m}P(z_l\mid x_n=a)}{P(\{z_l:l\in M(n)\setminus m\})}$$
$$=\alpha_{mn}P_n^a\prod_{l\in M(n)\setminus m}R_{ln}^a$$

其中，α_{mn} 为归一化参数，使 $Q_{mn}^{+1}+Q_{mn}^{-1}=1$。

最后得到后验概率：

$$Q_n^a=\alpha_{mn}P_n^a\prod_{l\in M(n)\setminus m}R_{ln}^a$$

⑥ 判决

如果 $Q_n^{+1}<Q_n^{-1}$，判定 $x_n=+1$，则 $X_n=0$，判断发送端传送了‘0’；

如果 $Q_n^{+1}>Q_n^{-1}$，判定 $x_n=-1$，则 $X_n=1$，判断发送端传送了‘1’。

直到 $X\cdot H^T=0$，否则转到上述第二个步骤。

根据需要，可以设定迭代次数，若经过了设定的迭代次数仍未译码成功，则终止迭代，译码失败。

需要指出的是不同的译码算法和实现，将得到不同的纠错编码性能，本实施例中采用的 LDPC 和积译码算法只是目前普遍采用的算法之一。

本发明所提出的用于地面数字电视广播的纠错编码方法经过计算机仿真后，已经用现场可编程器件（FPGA）得到了硬件实现。一个采用本发明所述方法的地面数字多媒体电视广播发射系统组成如图 9 所示。输入的 MPEG TS 码流可以是视频、音频、图形、数据等多媒体信息，为了抵抗传输过程中产生的误码，TS 流经过本发明所述方法的 BCH 和三种 LDPC 之一的级联纠错编码后，送给调制器成为数字 QPSK/mQAM 调制信号，再送给 OFDM 正交频分复用多载波调制，最后与时域同步 PN 序列复接后，构成一个时域同步正交频分复用

（TDS-OFDM）信号，经数模转换模块，转换为合适的模拟信号，射频 RF 模块接收此模拟信号，处理后的结果送给发射天线或其它信号发射机。

一个采用本发明所述方法的地面数字多媒体电视广播接收系统组成如图 10 所示。天线或其它信号接收机接收调制信号，送给下变频模块进行频率变换后，送给模数转换变为数字信号，同时恢复载波、时钟等同步，然后经过 OFDM 多载波解调。OFDM 解调后的数字信号经过本发明所提出的 LDPC 码 BP 迭代译码后，送给后面 BCH 解码器，最后恢复 MPEG TS 码流。

对于一个传输系统，传输信息的可靠性指标是 BER～SNR，采用本发明所述方法的地面数字多媒体电视广播系统在64QAM 调制、1/9保护间隔、0.6编码码率和迭代4次情况小，实际测试所获得的 BER～SNR 性能如图11所示。从图中可以看到，在视频广播逐常要求的 BER=10^{-11} 性能时，系统所需要的 SNR 为14.5dB，性能良好，并且由于采用的 LDPC 码长度只有 7488 个比特，相比于欧洲 DVB-S2 所采用的 64800比特的 LDPC码，本发明所提出的 LDPC 码时延小，实现简单，性能优异，满足地面数字电视广播对高码率的性能需求。

上面结合附图对本发明的具体实施例进行了详细说明，但本发明并不限制于上述实施例，在不脱离本申请的权利要求的精神和范围情况下，本领域的技术人员可作出各种修改或改型。

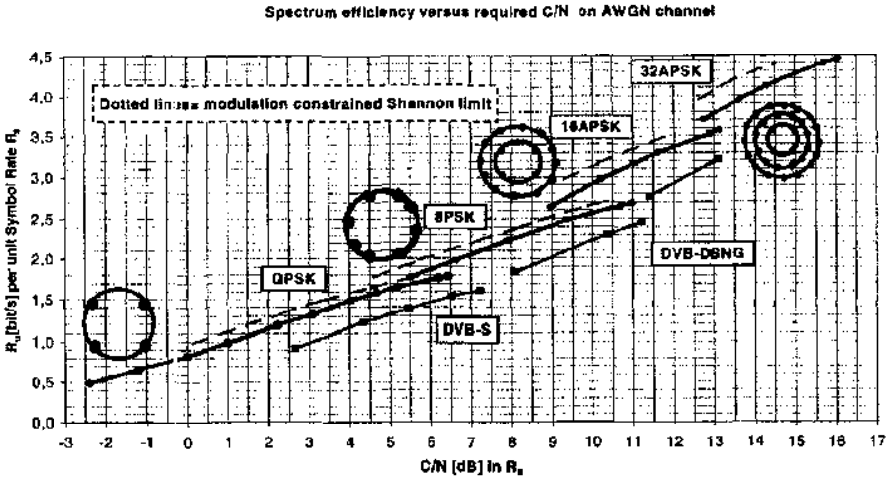

Spectrum efficiency versus required C/N on AWGN channel
Dotted lines: modulation constrained Shannon limit
32APSK
16APSK
8PSK
QPSK
DVB-DSNG
DVB-S
R_u[bit/s] per unit Symbol Rate R_s
C/N [dB] in R_s

图 1

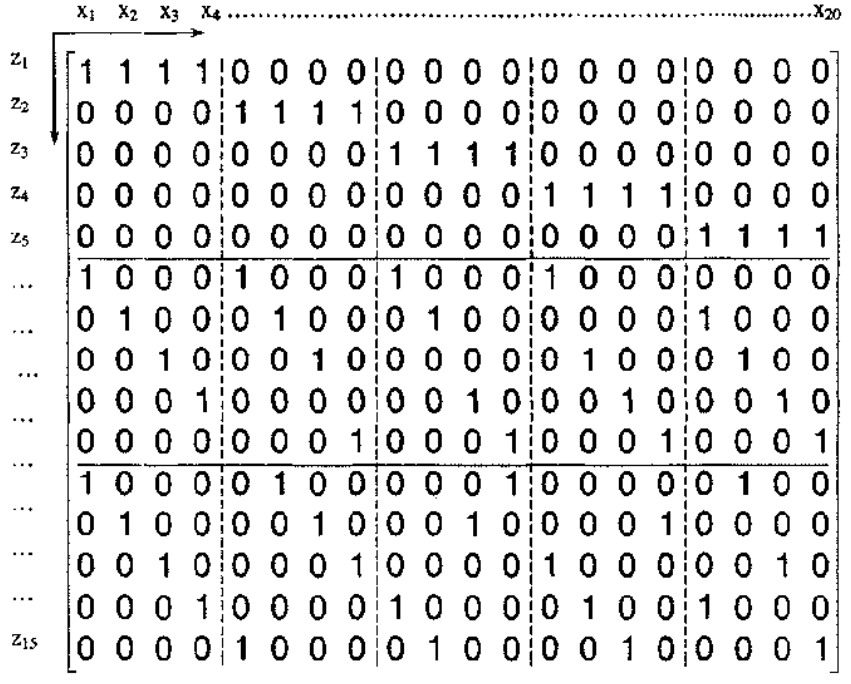

x_1 x_2 x_3 x_4 ..... x_{20}
z_1 z_2 z_3 z_4 z_5 ... ... z_{15}

图 2

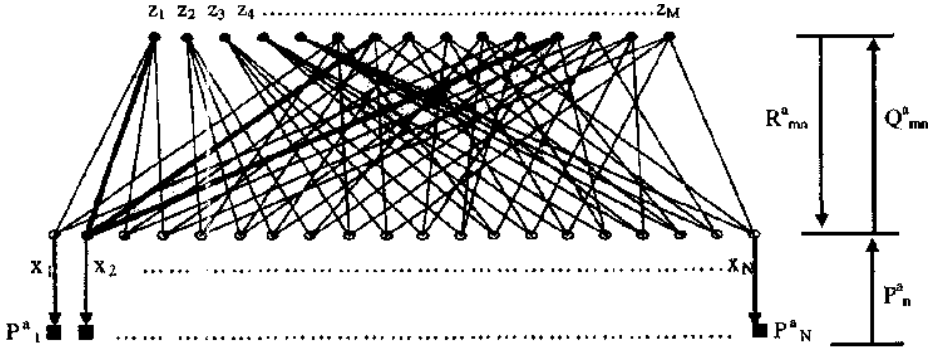

z_1 z_2 z_3 z_4 ..... z_M
R^a_{mn}
Q^a_{mn}
x_1
x_2
..... x_N
P^a_n
P^a_1
P^a_N

图 3

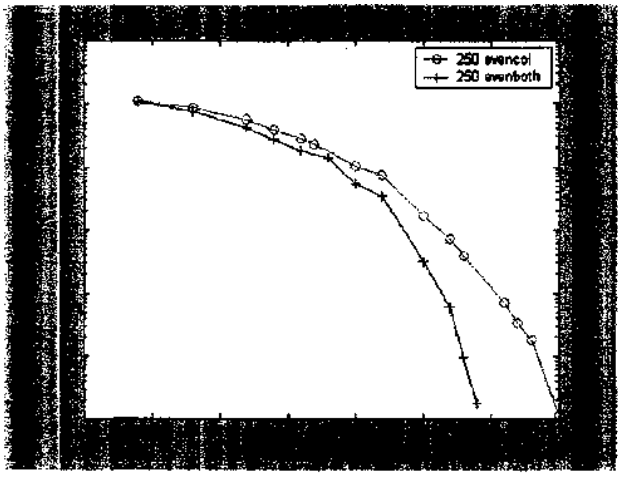

250 swencol
250 swenboth

图 4

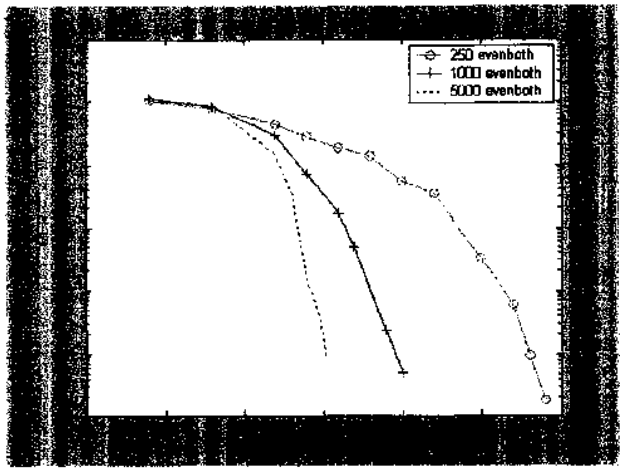

250 evenboth
1000 evenboth
5000 evenboth

图 5

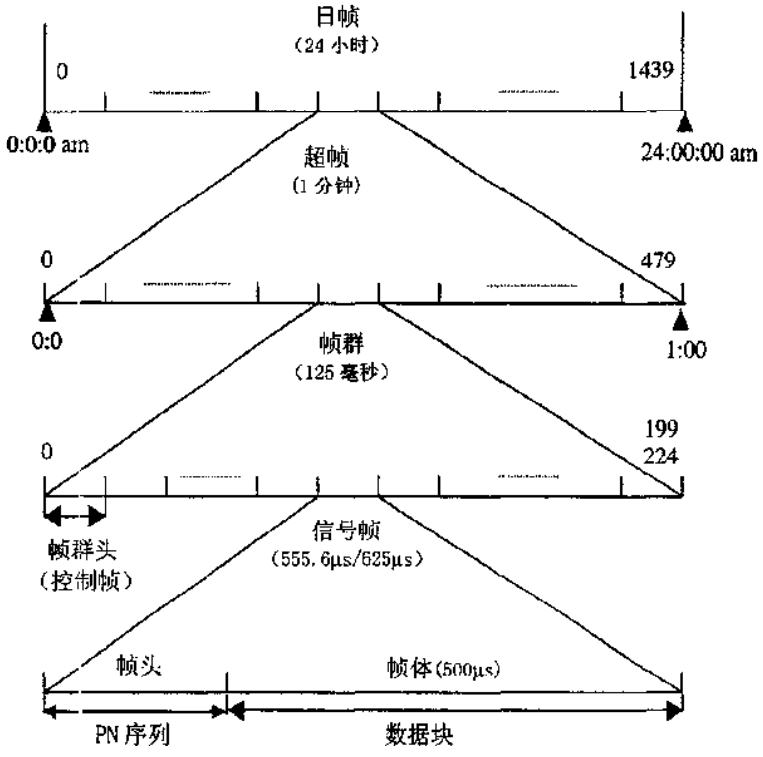

日帧
(24 小时)
0
1439
0:0:0 am
24:00:00 am
超帧
(1 分钟)
0
479
0:0
1:00
帧群
(125 毫秒)
0
199
224
帧群头
(控制帧)
信号帧
(555.6μs/625μs)
帧头
帧体(500μs)
PN 序列
数据块

图 6

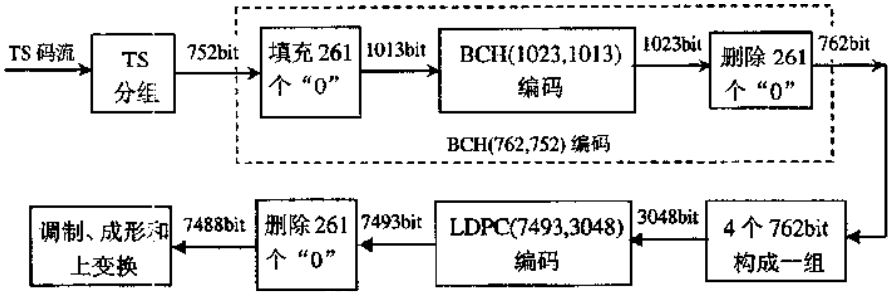

TS 码流
TS 分组
752bit
填充 261 个 "0"
1013bit
BCH(1023,1013) 编码
1023bit
删除 261 个 "0"
762bit
BCH(762,752) 编码
4 个 762bit 构成一组
3048bit
LDPC(7493,3048) 编码
7493bit
删除 261 个 "0"
7488bit
调制、成形和上变换

图 7

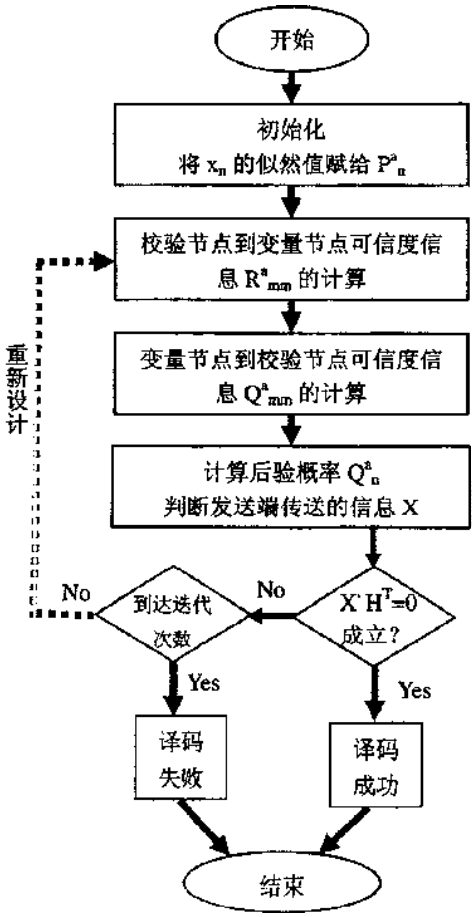

开始
初始化
将 x_n 的似然值赋给 P^a_n
校验节点到变量节点可信度信息 R^a_{mn} 的计算
变量节点到校验节点可信度信息 Q^a_{mn} 的计算
计算后验概率 Q^a_n
判断发送端传送的信息 X
$X \cdot H^T=0$ 成立?
No
到达迭代次数
No
重新设计
Yes
Yes
译码失败
译码成功
结束

图 8

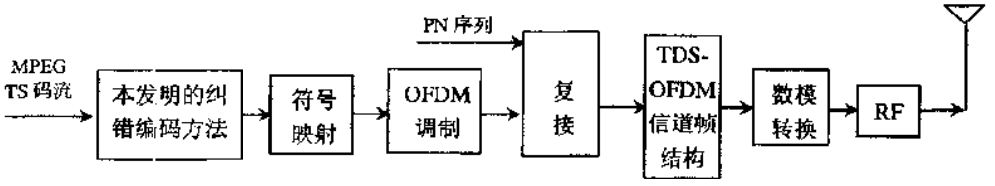

MPEG
TS 码流
本发明的纠
错编码方法
符号
映射
PN 序列
OFDM
调制
复
接
TDS-
OFDM
信道帧
结构
数模
转换
RF

图 9

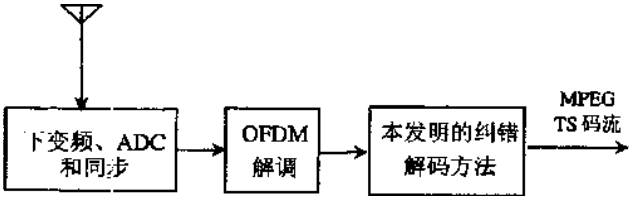

下变频、ADC
和同步
OFDM
解调
本发明的纠错
解码方法
MPEG
TS 码流

图 10

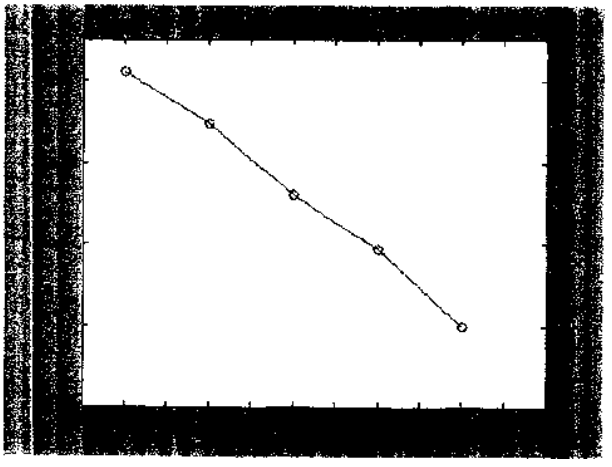


图 11

Iterative decoding algorithm for space hour bit interlaced modulating system and receiving system

Abstract

An iteration coding algorithm of idle time bit laced modulation system includes confirming symbol estimation value and currently detected send-signal then confirming more than one most possible estimation value of said send-signal according to confirmed symbol estimation value, calculating external information of each information bit by utilizing estimation value of obtained all send-signal afterall send-signals are detected. The receiving system of idle time bit interlacing modulation system is also disclosed.

CN100589469C

China

Download PDF Find Prior Art Similar

Other languages

Chinese

Inventor

李颖

李◆杰

Worldwide applications

2006 CN

Application CN200610103860A events

2006-08-04

Application filed by 华为技术有限公司, 西安电子科技大学

2006-08-04

Priority to CN200610103860A

2007-01-03

Publication of CN1889555A

2010-02-10

Application granted

2010-02-10

Publication of CN100589469C

Info

Non-patent citations (2)

Cited by (2)

Legal events

Similar documents

Priority and Related Applications

External links

Espacenet
Global Dossier
Discuss

Description

The interative encode method of space hour bit interlaced modulating system and SISO demodulator

Technical field

The present invention relates to the Space Time Coding technology, particularly a kind of iterative decoding algorithm of space hour bit interlaced modulating system and receiving system.

Background technology

Multiple-input and multiple-output (MIMO:Multiple-Input Multiple-Output) technology based on many antenna arrays can spatially provide separate a plurality of parallel sub-channels, improve information rate or communication quality effectively, opened up a brand-new research field and design

concept.Space Time Coding is a kind of suitable multiaerial system, can approaches the novel channel coding schemes of multi-antenna channel capacity.

With traditional binary channel coding technology and Space Time Coding technology cascade, and by inside and outside sign indicating number design, the performance that can obtain to approach the multi-antenna channel capacity.According to the difference of the space-time coding method that is adopted, existing cascade scheme can be divided three classes, i.e. divide into groups empty time-code and cascade hierarchical space-time code of the empty time-code of cascade grid (Trellis), cascade.Wherein, the cascade hierarchical space-time code is also referred to as space hour bit interlaced modulation (ST-BICM) system.Empty time-code of cascade Trellis and the cascade relatively difficulty of raising that a common problem that empty time-code exists is the system spectrum utilance of dividing into groups.

Compare with other two kinds of cascade schemes, the cascade hierarchical space-time code utilizes the conventional channel coding techniques that diversity gain and coding gain are provided, (BLAST) mode provides spatial multiplexing gain when utilizing layering empty simultaneously, can obtain the higher availability of frequency spectrum under the prerequisite of guaranteed performance, be one of comparatively promising solution.

ST-BICM system soft inputting and soft commonly used is exported in (SISO, Soft-Input Soft-Output) iterative decoding algorithm, and the SISO demodulator is according to M received signal and bit prior information $L_{A,demo}^{b}$, calculate the external information L of each coded-bit $_{E,demo}^{b}$, i.e. log-likelihood ratio, again with the external information of corresponding all coded-bits after deinterleaving is handled, as prior information $L_{A,out}^{b}$Send into the outer code decoder device; The external information L of each bit among the prior information sequence of calculation u that the utilization of outer code decoder device receives $_{E,out}^{b}$, after interweaving as prior information $L_{A,demo}^{b}$Send into the SISO demodulator, thereby finish iterative process one time.For the first time during iteration, bit prior information $L_{A,demo}^{b}$Be initialized as 0.

Adopt the N transmit antennas, the emission system structural representation of the ST-BICM system of M root reception antenna is shown in Fig. 1 (a).The information sequence a that information source produces at

first sends into outer code coder and obtains coded sequence u, sequence u carries out being input to modulator after the interleaving treatment through interleaver, modulator is modulated the sequence that receives, send to cascade hierarchical space-time code module then, obtain N parallel sub data flow after spatial mappings, corresponding respectively N root antenna transmission is gone out.

If the i transmit antennas is c at t transmission signal constantly $_{t}^{i}$, the channel fading coefficient between i transmit antennas and the j root reception antenna is $\alpha_{i,j}$, corresponding received signal r on the j root reception antenna then $_{t}^{j}$Can be expressed as:

r t j = Σ i = 1 N a i , j c t j + η t j - - - (1)

Wherein, η_{t}^{j}Be the channel additive noise on the j root reception antenna, it is obeyed, and average is zero, variance is N_0Multiple Gaussian Profile.Simple for discussing, the hypothesis channel is a flat fading among the application, and does not consider the correlation between the antenna, promptly for different i and j, $\alpha_{i,j}$Be to add up independently.

The ST-BICM system adopts the cascade hierarchical space-time code.Channel coding technology that the ST-BICM system is adopted and SISO iterative decoding algorithm are two key factors that influence its systematic function and complexity.Referring to Fig. 1 (b), Fig. 1 (b) is the receiving system structural representation of ST-BICM system.Shown in Fig. 1 (b), the SISO iterative decoder of ST-BICM system mainly is to be made of outer code decoder device and SISO demodulator two parts, intercourses external information by interleaver and deinterleaver between code decoder and the SISO demodulator outside simultaneously and realizes deciphering iteration.The basic principle of SISO demodulating algorithm commonly used is at present: utilize all possible symbol sebolic addressing to calculate the log-likelihood ratio and the external information of each bit, index increases the complexity of calculating with the increase of number of transmit antennas and order of modulation.

Introduce maximum likelihood (ML) SISO iterative decoding algorithm commonly used below, and the computation complexity of ML SISO iterative decoding algorithm is analyzed.Suppose in the signal constellation (in digital modulation) totally 2^{b}Individual, the constellation symbol collection is The

signal that i root antenna is carved transmission at a time is c_i, the signal that then corresponding reception antenna j receives is:

$r_l = \alpha_{1,j} c_1 + \alpha_{2,j} c_2 + \ldots + \alpha_{N,j} c_N + \eta_j$ (2)

When number of transmit antennas is N, the signal r that reception antenna j received in per moment $_j$In include the information of bN bit, will constitute $c_1, c_2 \ldots, c_N$All bN bit informations be expressed as:

$b = (b_1, \ldots, b_b, b_{b+1}, \ldots, b_{bN})$ (3)

B wherein $_{B(i-1)+1} \ldots, b_{Bi}$Represent the constellation point c that the i transmit antennas sends $_{l,i}$=1,2 ..., N.

L bit information b among the sequence b_lLog-likelihood ratio can calculate by following formula

Λ (b l) = log Pr [b l = 1 | r l , · · · , r M |] Pr [b l = 0 | r l , · · · , r M |] = log Pr [b l = 1 , r 1 , · · · , r M] Pr [b l = 0 , r l , · · · , r M]

= log Σ b : b 1 = 1 Pr [r 1 , · · · , r M , b] Σ b : b 1 = 0 Pr [r 1 , · · · , r M , b] = log Σ c : c = f (b) , b 1 = 1 Pr [r 1 , · · · r M , c] Σ c : c = f (b) , b 1 = 0 Pr [r 1 , · · · · r M , c] - - - (4)

C=(c wherein $_1, c_2 \ldots, c_N$), f (b) is the mapping from b to c, following formula can further be written as:

Λ (b 1) = log Σ c : c = f (b) , b 1 = 1 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Pr [c 1 , · · · , c N] Σ c : c = f (b) , b 1 = 0 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Pr [c 1 , · · · , c N] - - - (5)

It is independent to suppose respectively to send signal, can get

Λ (b 1) = log Σ c : c = f (b) , b 1 = 1 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Π l = 1 N Pr [c 1] Σ c : c = f (b) , b 1 = 0 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Π l = 1 N Pr [c 1] - - - (6)

Owing to used bit interleaver, can think that constituting each bit that sends signal adds up independent, promptly

Pr [c] = Π k = 1 b Pr [b k] - - - (7)

By the notion of formula (6) (7) and bit external information, the decoding external information of l bit can be expressed as:

Λ (b 1) = log Σ c : c = f (b) , b 1 = 1 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Π k = 1 , k ≠ 1 bN Pr [b k] Σ c : c = f (b) , b 1 = 0 Pr [r 1 , · · · , r M | c 1 , · · · , c N] Π k = 1 , k ≠ 1 bN Pr [b k]

= log Σ c : c = f (b) , b 1 = 1 Π j = 1 M Pr [r j | c 1 , · · · , c N] Π k = 1 , k ≠ 1 bN Pr [b k] Σ c : c = f (b) , b 1 = 0 Π j = 1 M Pr [r j | c 1 , · · · , c N] Π k = 1 , k ≠ 1 bN Pr [b k] - - - (8)

By

Pr (r 1 , · · · , r M | c 1 , · · · , c N) = 1 π N 0 exp (- Σ j = 1 M | r j - Σ i = 1 N a ij c 1 N 0) - - - (9)

Formula (8) can further be written as:

Λ (b 1) = log Σ c : c = f (b) , b 1 = 1 Π j = 1 M exp (| r j - Σ i = 1 N a 1 , j c 1 | 2 N 0) Π k = 1 , k ≠ 1 bN Pr [b k] Σ c : c = f (b) , b 1 = 0 Π j = 1 M exp (| r j - Σ i = 1 N a 1 , j c 1 | 2 N 0) Π k = 1 , k ≠ 1 bN Pr [b k] - - - (9)

Formula (10) be SISO demodulator output about l bit b $_l$The output external information, in iterative process as the input prior information of outer code decoder device.

By formula (10) as can be known, the SISO demodulator is when calculating the external information of each bit, need to consider the possible sequence of all 2bN kinds, be that computation complexity increases with number of transmit antennas and order of modulation index, in some cases, very big or order of modulation is too high as number of transmit antennas, can produce the high complexity that can not realize.

During the vertical layered space of prior art (VBLAST) iterative decoding algorithm be a kind of complexity low, realize being easy to algorithm.

The BLAST receiver is based on and disturbs the United Technologies that suppress with Interference Cancellation.Because the detection quality of the signal of signal to noise ratio (snr) maximum is best relatively, simultaneously to having the greatest impact that the detection of other signal brings, so the VBAST basic idea is: the estimated value of calculating the symbol of SNR maximum earlier, from received signal, eliminate of the influence of this symbol then to all the other symbol detection, from the received signal of revising, carry out the detection of SNR maximum symbol again, and then eliminate of the influence of this symbol all the other symbol detection.So analogize, the symbol of SNR minimum detects at last.It is maximum that this detection makes the symbol detection of SNR minimum be benefited in proper order, thereby make the best performance of whole interference cancellation method.

(ZF-BLAST) algorithm is an example during below with the ZF vertical layered space, and brief description VBLAST algorithm detects the flow process of handling:

At first, signal model is reduced to: R=HS+N,

Initialization G matrix:

$G_1 = H^+, \quad i = 1$ (a)

In order to determine the transmission signal k of the i time detection $_i$, need try to achieve not detected yet all and send the SNR of signal, select a signal of SNR maximum as the i time detection, wherein:

k 1 = arg min | | (G i) j | | 2 j ∉ { k , , k 2 , · &Center Dot; · , k i - 1 } - - - (b)

Select the signal k of the i time detection $_i$Pairing ZF vector

$$w_{ki} = (G_i)_{ki}^T \quad \text{---(c)}$$

Wherein, Representing matrix G_i K_iOK

Utilize formula (d) to calculate the transmission signal k that detects for the i time $_i$The corresponding symbol estimated value

$$y_{ki} = w_{ki}^T R_i \quad \text{---(d)}$$

Utilize formula (e) to k_iThe corresponding symbol estimated value Quantize:

$$\hat{S}_{ki} = Q(y_{ki}) \quad \text{---(e)}$$

The transmission signal k that utilizes formula (f) from received signal, will detect $_i$Offset; Utilize formula (g) to upgrade channel matrix, with after testing transmission signal k in the channel matrix $_i$Corresponding row are changed to zero, and other column vector remains unchanged:

$$R_{i+1} = R_i - \hat{S}_{ki} \cdot (H)_{ki} \quad \text{---(f)}$$

$$G_{i+1} = H_{\bar{k}i+} \quad \text{---(g)}$$

Then, will detect number of times and be changed to i+1, return-formula (b) ～(h) continues to calculate, until $i=N_T$Stopped in+1 o'clock calculating, obtain the estimated value that all send signal:

Though the VBLAST algorithm is realized simple, performance is good inadequately.Because such algorithm utilizes formula (e) to k_iThe corresponding symbol estimated value The judging process that quantizes is to declare firmly, and when SNR was higher, the effect of bringing that offsets of utilizing formula (f) to carry out was a forward; When SNR is low, mistaken verdict may appear, and based on mistaken verdict, offseting of utilizing that formula (f) carries out is exactly negative sense, so mistaken verdict is very big to performance impact.

Summary of the invention

Main purpose of the present invention is to provide a kind of low complex degree interative encode method of space hour bit interlaced modulating

system, when reducing the iterative decoding algorithm complexity, obtains preferable performance.

Another main purpose of the present invention is to provide a kind of soft inputting and soft output SISO demodulator, when reducing the iterative decoding algorithm complexity, obtains preferable performance.

In order to realize the first aspect of above-mentioned main purpose, the invention provides a kind of interative encode method of space hour bit interlaced modulating system, this method may further comprise the steps:

A, the transmission signal of determining current detection and sign estimation value thereof;

B, the sign estimation value of utilizing steps A to determine, obtain current detection the transmission signal more than a most probable estimated value;

Wherein, comprising of the described transmission signal that obtains current detection more than a most probable estimated value:

Signal constellation which is divided into zone more than, and according to the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value;

C, after having detected all and having sent signals, all that utilize step B to obtain send the estimated value of signals, calculate the external information of each information bit.

The transmission signal and the sign estimation value thereof of the described definite current detection of steps A comprise:

A1, utilize predetermined algorithm to calculate the weighing vector of all transmission signals that do not detect of receiving terminal; Described predetermined algorithm is zero forcing algorithm or least-mean-square error algorithm;

A2, according to the weighing vector that steps A 1 obtains, determine the signal to noise ratio of the transmission signal that all are not detected, select

the transmission signal of the transmission signal of signal to noise ratio maximum as current detection;

The weighing vector of the transmission signal of the current detection that A3, utilization are selected is weighted processing to the signal on the every reception antenna, obtains the sign estimation value of the transmission signal of current detection.

In the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point more than a most probable estimated value be as this transmission signal: according to the phase place of the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value.

In the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point more than a most probable estimated value be as this transmission signal: according to the real part of the sign estimation value of the transmission signal of current detection and/or the size of imaginary part, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value.

In the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than after the most probable estimated value, further comprise: according to the residing zone of transmission signal of current detection, signal constellation which is divided into one with the upper strata, the constellation point number of electing additional member as required again, the sign estimation of selecting to send signal with this from each layer is worth nearest constellation point.

The described selection from each layer with the nearest constellation point of the sign estimation value of this transmissions signal is: the selection constellation point nearest with the sign estimation value of transmission signal current detection equal number from each layer; Perhaps from the

near more layer of the constellation point selected of distance select many more constellation point.

In order to realize the foregoing invention purpose on the other hand, the invention provides a kind of soft inputting and soft output SISO demodulator, comprising:

Outer code decoder device, soft inputting and soft output SISO demodulator, deinterleaver and interleaver is characterized in that described SISO demodulator comprises: sign estimation value module, constellation point estimation module and external information computing module,

Sign estimation value module is used for determining the transmission signal and the sign estimation value thereof of current detection, and the sign estimation value of determining is sent to the constellation point estimation module;

The constellation point estimation module, be used for according to the sign estimation value that receives, obtain and store current detection the transmission signal more than a most probable estimated value, comprise: signal constellation which is divided into zone more than, and according to the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value; And the estimated value of all the transmission signals that will store sends to the external information computing module;

The estimated value of all transmission signals that external information computing module, utilization receive is calculated the external information of each information bit.

Described sign estimation value module comprises: weighing vector module and weighted module,

The weighing vector module, be used to utilize predetermined algorithm to determine the weighing vector of all transmission signals that do not detect of receiving terminal, and determine the signal to noise ratio of the transmission signal that all are not detected according to the weighing vector of all transmission signals that do not detect, select the transmission signal of the transmission signal of signal to noise ratio maximum, and the result

that will select sends to the weighted module as current detection; Described predetermined algorithm is zero forcing algorithm or least-mean-square error algorithm;

The weighted module utilizes the weighing vector of the transmission signal of the current detection of selecting that the signal on the every reception antenna is weighted processing, obtains the sign estimation value of the transmission signal of current detection, sends it to the constellation point estimation module.

Described constellation point estimation module comprises: the constellation point determination module,

The constellation point determination module is used for signal constellation which is divided into more than one the zone, and selects most possible constellation point more than according to the sign estimation value of the transmission signal of the current detection that receives.

Described constellation point estimation module further comprises: hierarchical block, be used for the residing zone of transmission signal according to current detection, signal constellation which is divided into one with the upper strata, the constellation point number of electing additional member as required again, selection is worth nearest constellation point with the sign estimation of the transmission signal of current detection from each layer.

The iterative decoding algorithm of space hour bit interlaced modulating system provided by the invention and receiving system in conjunction with ML-SISO and VBLAST receiver, reduce the complexity of iterative decoding algorithm by the size that changes the set of alternative signal vector.Such as, when adopting QPSK to modulate 4 transmit antennas, if adopt traditional iterative decoding algorithm, need to calculate all $4^4=256$ kinds of four-dimensional signal vectors; If adopt decoding algorithm provided by the invention, when every antenna is only chosen three possibility signaling points, only need to calculate $3^4=81$ kinds of four-dimensional signal vectors, complexity only are about 1/3 of traditional algorithm.In fact, along with the increase of number of transmit antennas and order of modulation, the advantage of algorithm that the present invention carries on complexity can be more obvious.

Referring to Fig. 2, Fig. 2 is two two and receives relatively schematic diagram of the performance of ST-BICM system under different decoding algorithms.Here, two two receipts ST-BICM system emulation parameters are:

Chnnel coding adopts 64 state convolution codes of 1/2 code check; Channel is assumed to be independent decline, and receiving terminal can accurately be estimated the channel fading coefficient; Interleaver sizes is 1024, and modulation system is the QPSK modulation; The iterative decoding algorithm based on ZF of decoding algorithm for proposing among the present invention, based on iterative decoding algorithm and the existing iterative decoding algorithm based on ML of MMSE, iterations is respectively 1 time and 2 times.

Curve 21,22,23 is respectively the iterative decoding algorithm based on ZF that proposes among the present invention among Fig. 2, the performance curve during based on the iterative decoding algorithm of MMSE and existing iterative decoding algorithm iteration 1 time based on ML; Curve 24,25,26 is respectively the iterative decoding algorithm based on ZF that proposes among the present invention, the performance curve during based on the iterative decoding algorithm of MMSE and existing iterative decoding algorithm iteration 2 times based on ML.

Iterative decoding algorithm for the present invention's proposition based on ZF and MMSE, select 3 signaling points to constitute the alternative signal vector for every layer in this emulation, amount to 9 2D signal vectors, and to consider all 16 2D signal vectors based on the ML iterative decoding algorithm, therefore the complexity of the iterative decoding algorithm of the present invention's proposition is compared with the ML iterative decoding algorithm, has significantly and reduces.And analysis chart 2 as can be known, and it is very little that the performance of the iterative decoding algorithm that the present invention proposes and diversity gain and ML iterative decoding algorithm are compared difference, in allowed band.

In addition, decoding algorithm performance based on the MMSE detection criteria of the present invention is better than the decoding algorithm based on the ZF detection criteria, this mainly is because ZF criterion itself has only been considered the influence of channel fading coefficient matrix, does not consider the channel additive noise, has bigger error propagation.

In a word, the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention adopts MMSE to detect or ZF detects, and compares with existing ML Iterative detection algorithm, and its algorithm complex can reduce near half, and the loss of performance is less, in allowed limits.And the VBLAST decoding algorithm of the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention and prior art relatively, and decoding performance improves a lot.Therefore, utilize the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention and receiving system when reducing algorithm complex, to obtain preferable performance.

Description of drawings

Fig. 1 (a) is the emission system structural representation of ST-BICM system;

Fig. 1 (b) is the receiving system structural representation of ST-BICM system;

Fig. 2 is two two and receives relatively schematic diagram of the performance of ST-BICM system under different decoding algorithms;

Fig. 3 is the schematic flow sheet of the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention;

Fig. 4 is the schematic flow sheet of a preferred embodiment of the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention;

Fig. 5 is the QPSK signal constellation which;

Fig. 6 is nine sub-square area schematic of 16QAM signal constellation which of the present invention;

Fig. 7 is four belt-like zone schematic diagrames of 16QAM signal constellation which of the present invention;

Fig. 8 is the layering schematic diagram that example provides for the present invention with sub-square zone 1;

Fig. 9 is the layering schematic diagram that example provides for the present invention with sub-square zone 2;

Figure 10 is the layering schematic diagram that example provides for the present invention with sub-square zone 5;

Figure 11 is the layering schematic diagram that example provides for the present invention with belt-like zone 1;

Figure 12 is the structural representation of the receiving system of space hour bit interlaced modulating system of the present invention.

Embodiment

For making purpose of the present invention, technical scheme and beneficial effect clearer, below in conjunction with embodiment and accompanying drawing, the present invention is described in more detail.

The iterative decoding algorithm of space hour bit interlaced modulating system provided by the invention and receiving system, the transmission signal and the sign estimation value thereof of at first definite current detection; According to the sign estimation value of the transmission signal of current detection, determine this transmission signal more than a most probable estimated value; After having detected all transmission signals, utilize the estimated value of all transmission signals that obtain, calculate the external information of each information bit.

Referring to Fig. 3, Fig. 3 is the schematic flow sheet of the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention.As shown in Figure 3, this flow process may further comprise the steps:

Step 301 is determined the transmission signal and the sign estimation value thereof of current detection.

Here, determine that the transmission signal of current detection and sign estimation value thereof comprise:

A1, utilize predetermined algorithm to calculate the weighing vector of all transmission signals that do not detect of receiving terminal.

A2, according to the weighing vector that step a1 obtains, determine the signal to noise ratio of the transmission signal that all are not detected, select the transmission signal of the transmission signal of signal to noise ratio maximum as current detection.

The weighing vector of the transmission signal of the current detection that a3, utilization are selected is weighted processing to the signal on the every reception antenna, obtains the sign estimation value of the transmission signal of current detection.

Step 302, the sign estimation value of utilizing step 301 to determine, obtain current detection the transmission signal more than a most probable estimated value.

Step 303, after having detected all and sending signals, all that utilize that step 302 obtains send the estimated value of signals, calculate the external information of each information bit.

Referring to Fig. 4, Fig. 4 is the schematic flow sheet of a preferred embodiment of the iterative decoding algorithm of space hour bit interlaced modulating system of the present invention.As shown in Figure 4, this flow process may further comprise the steps:

Step 401 utilizes predetermined algorithm to determine the weighing vector of all not detected transmission signals of receiving terminal.Here, predetermined algorithm can be ZF detection algorithm or least-mean-square error algorithm.

If receiving terminal about all N the weighing vector formation matrix W that send signal formation is:

$$W=(w_1,\ w_2,\ \ldots,\ w_N)^T \quad (11)$$

According to ZF detection algorithm, matrix $W_{ZF}=H^+$, here, H is a channel matrix, H^+Generalized inverse matrix for matrix H.

According to least-mean-square error algorithm, the selection of matrix W should make mean square of error value $E\{ \| c-Wr \|^2$Minimum, the corresponding reception antenna weighing vector matrix based on the MMSE criterion is:

$W_{MMSE}=H^{H}(HH^{H}+\sigma^{2}I)^{+}$ (12)

Wherein, σ^{2}Be the noise that comprises in the received signal, H^{H}Be the conjugate transpose of matrix H, $(HH^{H}+\sigma^{2}I)$+be matrix $(HH^{H}+\sigma^{2}I)$ generalized inverse matrix.With W_{MMSE}The vectorial w of row $_{i}$As reception antenna for sending signal c_{i}Formed weighing vector.

Step 402 is according to weighing vector W_{ZF}Or W_{MMSE}Determine the signal to noise ratio that respectively sends signal (SNB) that all are not detected, the signal of selecting SNB maximum wherein is as detection signal that ought be last time.

Concrete method is: calculate weighing vector w_{i}(i=1,2 ..., the N) length of matrix, the shortest corresponding signal that sends of explanation of length is subjected to the noise jamming minimum, and the signal to noise ratio maximum then should be as the transmission signal that ought last time detect.

Step 403 utilizes the weighing vector of the last time detection signal of step 402 selection that the transmission signal on the every reception antenna is weighted processing, obtains the last time estimated value of detection signal.

Concrete processing method is: formula (1) is written as matrix form:

$r=Hc+\eta$ (13)

Suppose $r=(r_{1}, r_{2}..., r_{M})^{T}$Be M dimension received signal vector, $c=(c_{1}, c_{2}..., c_{N})^{T}$Be N dimension transmission signal vector, $\eta=(\eta_{1}, \eta_{2}..., \eta_{M})^{T}$Be M dimension additive noise vector, H is the channel matrix of M * N.In the present invention, suppose accurately estimating channel information of receiving terminal.

The weighing vector of detection signal last time With multiply by formula (13) two ends, the sign estimation value that obtains the transmission signal that ought last time detect is:

$y k i = w k i T r = c k i + v k i$ - - - (14)

Step 404 according to the sign estimation value of the transmission signal that ought last time detect, is determined Ni most probable estimated value of this transmission signal.

Here, N_iFor greater than 1 integer.

Concrete processing method is: utilize formula (14), select in the signal constellation which and the estimated value of detection signal last time Nearest N_iIndividual as sending signal N_iIndividual estimated value { c ^ k i , 1 , c ^ k i , 2 , · · · , c ^ k i , N i } .

Step 405 is upgraded received signal vector and channel matrix, the transmission signal that disappears and detected from received signal vector and channel matrix.

Concrete processing method is: utilize formula (f) and (g), detection signal is offset in received signal, and with in the channel matrix the row of detection signal correspondence be changed to zero.

Step 406 is judged current detection number of times whether less than the number N that sends signal, if less than, after then will detecting number of times and adding one, return step 401; Otherwise, execution in step 407.

Above-mentioned steps 405 also can be placed on step 406 back and carry out, judge earlier that promptly whether current detection number of times is less than the number that sends signal, when judging that current detection number of times when sending the number of signal, upgrades received signal vector and channel matrix earlier, and then returns step 401.

Step 407, the estimated value that all that obtain is sent signal constitutes the symbol sebolic addressing set, is designated as

The estimated value of supposing the signal correspondence of the i time detection is N_iIndividual, then when sending signal for total N, symbol sebolic addressing is gathered By N_1* N_2* ... * N_NIndividual N dimension symbol sebolic addressing constitutes.Wherein, N_1, N_2... N_NIt all is integer greater than 1.

Step 408, the symbol sebolic addressing set that utilizes step 407 to constitute And formula (10) calculates the external information of each information bit.

The characteristics that the present invention is directed to the planisphere of different modulating mode provide definite principle of several symbol sebolic addressings, be the sign estimation value of the above-mentioned steps 404 described bases transmission signal that ought last time detect, select in the letter x planisphere with estimated value nearest send definite principle of the most probable estimated value of signal more than a constellation point as this.

The symbol sebolic addressing that provides mpsk signal planisphere and MQAM signal constellation which is below respectively determined principle.

Therefore the mpsk signal planisphere is evenly distributed on the unit circle, can be according to having eliminated the property taken advantage of fading factor and i corrected signal y influencing of detection of transmitted signals $_{i}$Phase place, come to determine and corrected signal y $_{i}$Nearest several points.Here, i corrected signal y $_{i}$Be exactly the sign estimation value of the transmission signal of the i time detection, use $\angle$ y $_{i}$Expression corrected signal y $_{i}$Phase place.

With the QSPK signal constellation which is that the symbol sebolic addressing of example explanation mpsk signal planisphere is determined principle.Referring to Fig. 5, Fig. 5 is the QPSK signal constellation which.

For the QPSK signal, if select N $_{i}$=2, then:

When - π 4 ≤ ∠ y i ≤ π 4 The time, select signaling point 0 and 3 as two most probable alternative signal points;

When π 4 ≤ ∠ y i ≤ 3 π 4 The time, select signaling point 0 and 1 as two most probable alternative signal points;

When 3 π 4 ≤ ∠ y i ≤ 5 π 4 The time, select signaling point 1 and 2 as two most probable alternative signal points;

When 5 π 4 ≤ ∠ y i ≤ 7 π 4 The time, select signaling point 2 and 3 as two most probable alternative signal points.For the QPSK signal, if select N $_{i}$=3, then:

When 0 ≤ ∠ y i ≤ π 2 The time, select signaling point 0,1 and 3 as three most probable alternative signal points;

When π 2 ≤ ∠ y i ≤ π The time, select signaling point 0,1 and 2 as three most probable alternative signal points;

When ππ ≤ ∠ y i ≤ 3 π 2 The time, select signaling point 1,2 and 3 as three most probable alternative signal points;

When - 3 π 2 ≤ ∠ y i ≤ 2 π The time, select signaling point 0,2 and 3 as three most probable alternative signal points.

The MQAM signal constellation which is square substantially, can select most probable several signaling point according to the real part of corrected signal and the size of imaginary part.

The symbol sebolic addressing that with 16QAM is example explanation MQAM signal constellation which is determined principle.The concrete steps of this symbol sebolic addressing deterministic process comprise:

1) 16 signaling points is divided into 9 sub-square zones or four belt-like zones.Referring to Fig. 6 and Fig. 7, Fig. 6 is nine sub-square area schematic of 16QAM signal constellation which of the present invention; Fig. 7 is four belt-like zone schematic diagrames of 16QAM signal constellation which of the present invention.

2), select four most possible signaling points according to the size of corrected signal.In two kinds of situation:

First kind of situation: as | Re (y) |≤3d and | Im (y) | during≤3d, d is half of the Fig. 6 and the medium and small foursquare length of side of Fig. 7.Four signaling points most possible according to the choice of location in the residing sub-square of corrected signal zone.For example: if-d≤Re (y)≤d and-d≤Im (y)≤d, then select four signaling points in the zone 5; If-d≤Re (y)≤d and d≤Im (y)≤3d then selects four signaling points in the zone 6.

Second kind of situation: as | Re (y) |＞3d or | Im (y) | during＞3d, select four signaling points in the belt-like zone suitable among Fig. 7.For example: if Re (y)＞3d then selects four signaling points in the belt-like zone 1; If Re (y) ＜-3d, then select four signaling points in the belt-like zone 3; If Im (y)＞3d

then selects four signaling points in the belt-like zone 2; If Im (y)＜-3d, then select four signaling points in the belt-like zone 4.

For improving the performance of iterative decoding algorithm, can further elect signaling point additional member, concrete steps are as follows:

3) according to the residing zone of corrected signal, the signal constellation which of 16QAM is divided into several levels, determine then the signaling point of being elected additional member to divide three kinds of situation discussion below:

First kind of situation: when corrected signal is in sub-square zone 1,3,7 or 9, except that selecting four corresponding signaling points, remaining 12 signaling points are divided into two-layer, Fig. 8 is that example has provided the layering schematic diagram with sub-square regional 1.The signaling point number of electing additional member is as requested calculated the distance of signaling point in corrected signal and each sublayer, the signaling point that the nearest conduct of chosen distance is elected additional member then.For example, the distance between the signaling point gets final product on corrected signal and the ground floor if the signaling point number that requires to elect additional member smaller or equal to 5, can only be calculated.

Second kind of situation: in the time of among corrected signal is in sub-square zone 2,4,6 or 8, except that selecting four corresponding signaling points, simultaneously remaining 12 signaling points are divided into two-layerly, Fig. 9 is that example has provided the layering schematic diagram with sub-square regional 2.The signaling point number of electing additional member is as requested calculated the distance of signaling point in corrected signal and each sublayer, the signaling point that the nearest conduct of chosen distance is elected additional member then.For example,, can only calculate the distance between the signaling point on corrected signal and the ground floor, select nearest signaling point if the signaling point number that requires to elect additional member is counted less than the signal on the ground floor.

The third situation: in the time of among corrected signal is in sub-square zone 5, remaining 12 signaling points are divided into as shown in figure 10 two-layer, calculate the distance of signaling point in corrected signal and each sublayer then, the signaling point that the nearest conduct of chosen

distance is elected additional member.For example,, can only calculate the distance between the signaling point on corrected signal and the ground floor, select nearest signaling point if the signaling point number that requires to elect additional member is counted less than the signal on the ground floor.

The 4th kind of situation: as | Re (y) |＞3d or | Im (y) | during＞3d, except that selecting four corresponding signaling points, simultaneously remaining 12 signaling points are divided into 3 layers, Figure 11 is that example has provided the layering schematic diagram with belt-like zone 1, calculate the distance of signaling point in corrected signal and each sublayer then, the signaling point that the nearest conduct of chosen distance is elected additional member.For example,, can only calculate the distance between the signaling point on corrected signal and the ground floor, select nearest signaling point if the signaling point number that requires to elect additional member is counted less than the signal on the ground floor.

Here, the purpose of layering is in order to reduce amount of calculation, reduces the number of the Euclidean distance of asking as far as possible, and described Euclidean distance is promptly launched the distance between sign estimation value and the constellation point.At first, by judging the residing position of real part imaginary part of corrected signal, can obtain 4 nearest constellation point then with MQAM planisphere piecemeal; If needed constellation point number is greater than 4, again with remaining 12 constellation point layerings.The principle of layering is an Euclidean distance of asking minimum as far as possible.

The present invention detects in conjunction with MMSE and ZF detects, and has proposed a kind of iterative decoding algorithm of low complex degree, compares with existing ML iterative decoding algorithm, and complexity can reduce near half, and the loss of performance within the acceptable range.

In order further to reduce performance loss, the invention allows for the method for following raising performance:

Because detected which floor signaling point is bigger to Effect on Performance at first, therefore when each layer selected signaling point, adopt the method that successively reduces signaling point selection number to improve performance.Such as, for the situation of three transmitting antennas and QPSK modulation, can consider 4,3,3 detecting pattern,

promptly keep all 4 signaling points at ground floor, reduce the signal of selecting at second and the 3rd layer and count, consider 36 2D signal vectors altogether.

Referring to Figure 12, Figure 12 is the structural representation of the receiving system of space hour bit interlaced modulating system of the present invention, and this system comprises: outer code decoder device, soft inputting and soft output SISO demodulator, deinterleaver and interleaver.

Wherein, the SISO demodulator comprises: sign estimation value module 1201, constellation point estimation module 1202 and external information computing module 1203,

Sign estimation value module 1201 is used for determining the transmission signal and the sign estimation value thereof of current detection, and the sign estimation value of determining is sent to constellation point estimation module 1202;

Constellation point estimation module 1202 is used for according to the sign estimation value that receives, obtain and store current detection send signal more than a most probable estimated value, and the estimated value of all the transmission signals that will store sends to external information computing module 1203;

The estimated value of all transmission signals that external information computing module 1203, utilization receive is calculated the external information of each information bit.

Described sign estimation value module 1201 is made up of weighing vector module and weighted module.

The weighing vector module, be used to utilize predetermined algorithm to determine the weighing vector of all transmission signals that do not detect of receiving terminal, and determine the signal to noise ratio of the transmission signal that all are not detected according to the weighing vector of all transmission signals that do not detect, select the transmission signal of the transmission signal of signal to noise ratio maximum, and the result that will select sends to the weighted module as current detection.

The weighing vector of the transmission signal of the current detection that the utilization of weighted module is selected is weighted processing to the signal on the every reception antenna, obtains the sign estimation value of the transmission signal of current detection, sends it to constellation point estimation module 1202.

Described constellation point estimation module 1202 comprises the constellation point determination module, this constellation point determination module is used for signal constellation which is divided into more than one the zone, and selects most possible constellation point more than according to the sign estimation value of the transmission signal of the current detection that receives.

Described constellation point estimation module 1202 can further include hierarchical block, this hierarchical block is used for the residing zone of transmission signal according to current detection, signal constellation which is divided into one with the upper strata, the constellation point number of electing additional member as required again, selection is worth nearest constellation point with the sign estimation of the transmission signal of current detection from each layer.

By above embodiment as seen, the iterative decoding algorithm of space hour bit interlaced modulating system provided by the invention and receiving system, in conjunction with ML-SISO and VBLAST receiver, by changing the size of alternative signal vector set, realized when reducing the iterative decoding algorithm complexity, obtaining preferable performance.

In a word, the above is preferred embodiment of the present invention only, is not to be used to limit protection scope of the present invention.Within the spirit and principles in the present invention all, any modification of being done, be equal to replacement, improvement etc., all should be included within protection scope of the present invention.

Claims (10)

Hide Dependent

1, a kind of interative encode method of space hour bit interlaced modulating system is characterized in that, this method may further comprise the steps:
A, the transmission signal of determining current detection and sign estimation value thereof;
B, the sign estimation value of utilizing steps A to determine, obtain current detection the transmission signal more than a most probable estimated value;
Wherein, comprising of the described transmission signal that obtains current detection more than a most probable estimated value:
Signal constellation which is divided into zone more than, and according to the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value;
C, after having detected all and having sent signals, all that utilize step B to obtain send the estimated value of signals, calculate the external information of each information bit.
2, the method for claim 1 is characterized in that, the transmission signal and the sign estimation value thereof of the described definite current detection of steps A comprise:
A1, utilize predetermined algorithm to calculate the weighing vector of all transmission signals that do not detect of receiving terminal; Described predetermined algorithm is zero forcing algorithm or least-mean-square error algorithm;
A2, according to the weighing vector that steps A 1 obtains, determine the signal to noise ratio of the transmission signal that all are not detected, select the transmission signal of the transmission signal of signal to noise ratio maximum as current detection;
The weighing vector of the transmission signal of the current detection that A3, utilization are selected is weighted processing to the signal on the every reception antenna, obtains the sign estimation value of the transmission signal of current detection.
3, the method for claim 1 is characterized in that, in the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point more than a most probable estimated value be as this transmission signal:

According to the phase place of the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value.

4, the method for claim 1 is characterized in that, in the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point more than a most probable estimated value be as this transmission signal:

According to the real part of the sign estimation value of the transmission signal of current detection and/or the size of imaginary part, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value.

5, method as claimed in claim 4, it is characterized in that, in the described selection signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than after the most probable estimated value, further comprise:

The residing zone of transmission signal according to current detection is divided into one with the upper strata with signal constellation which, the constellation point number of electing additional member as required again, and the sign estimation of selecting to send signal with this from each layer is worth nearest constellation point.

6, method as claimed in claim 5, it is characterized in that described the selection with the nearest constellation point of the sign estimation value of this transmissions signal is: the selection constellation point nearest with the sign estimation value of transmission signal current detection equal number from each layer from each layer; Perhaps from the near more layer of the constellation point selected of distance select many more constellation point.

7, a kind of soft inputting and soft output SISO demodulator is characterized in that described SISO demodulator comprises: sign estimation value module, constellation point estimation module and external information computing module,

Sign estimation value module is used for determining the transmission signal and the sign estimation value thereof of current detection, and the sign estimation value of determining is sent to the constellation point estimation module;

The constellation point estimation module, be used for according to the sign estimation value that receives, obtain and store current detection the

transmission signal more than a most probable estimated value, comprise: signal constellation which is divided into zone more than, and according to the sign estimation value of the transmission signal of current detection, select in the signal constellation which with the sign estimation value of the transmission signal of current detection nearest more than a constellation point as this send signal more than a most probable estimated value; And the estimated value of all the transmission signals that will store sends to the external information computing module;
The estimated value of all transmission signals that external information computing module, utilization receive is calculated the external information of each information bit.
8, SISO demodulator as claimed in claim 7 is characterized in that, described sign estimation value module comprises: weighing vector module and weighted module,
The weighing vector module, be used to utilize predetermined algorithm to determine the weighing vector of all transmission signals that do not detect of receiving terminal, and determine the signal to noise ratio of the transmission signal that all are not detected according to the weighing vector of all transmission signals that do not detect, select the transmission signal of the transmission signal of signal to noise ratio maximum, and the result that will select sends to the weighted module as current detection; Described predetermined algorithm is zero forcing algorithm or least-mean-square error algorithm;
The weighted module utilizes the weighing vector of the transmission signal of the current detection of selecting that the signal on the every reception antenna is weighted processing, obtains the sign estimation value of the transmission signal of current detection, sends it to the constellation point estimation module.
9, as claim 7 or 8 described SISO demodulators, it is characterized in that described constellation point estimation module comprises: the constellation point determination module,
The constellation point determination module is used for signal constellation which is divided into more than one the zone, and selects most possible constellation point more than according to the sign estimation value of the transmission signal of the current detection that receives.
10, SISO demodulator as claimed in claim 9, it is characterized in that, described constellation point estimation module further comprises: hierarchical block, be used for the residing zone of transmission signal according to current detection, signal constellation which is divided into one with the upper strata, the constellation point number of electing additional

member as required again, selection is worth nearest constellation point with the sign estimation of the transmission signal of current detection from each layer.

Non-Patent Citations (2)

Title

串联级联空时码的研究与设计. 李颖.西安电子科技大学博士学位论文. 2005

串联级联空时码的研究与设计. 李颖.西安电子科技大学博士学位论文. 2005 *

* Cited by examiner, † Cited by third party

Cited By (2)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations

CN101626249B *2009-08-192013-04-10北京海尔集成电路设计有限公司 Soft-input soft-output coding system and soft-input soft-output coding method

CN102238132A *2010-04-302011-11-09联咏科技股份有限公司Modulation recognition method and device for communication system

* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle

US8755458B22014-06-17Spatio-temporal processing for communication

JP5397428B22014-01-22MIMO-OFDM communication method and MIMO-OFDM communication apparatus

CA2620067C2012-08-21Method and apparatus for selection of virtual antennas

Mesleh et al.2008Spatial modulation

RU2323525C22008-04-27Device and method for controlling transmission circuit according to condition of channel in communication system

KR100778647B12007-11-22Combined channel coding and space-block coding in a multi-antenna arrangement

JP4490922B22010-06-30Multi-antenna transmission apparatus and multi-antenna transmission method

RU2375823C22009-12-10Time-varying separation of periodic delay in ofdm system
CN100438392C2008-11-26Method and communication device using adaptive space-time encoding, modulation and error coding
US8462859B22013-06-11Sphere decoding apparatus
AU2003216479C12009-06-25Data transmission with non-uniform distribution of data rates for a multiple-input multiple-output (MIMO) system
US7194041B22007-03-20Ordered successive interference cancellation receiver processing for multipath channels
CN100364256C2008-01-23Time-domain transmit and receive processing with channel eigen-mode decomposition for systems
CN101068124B2011-11-23Method and system for processing signal in communication systems
KR100784744B12007-12-14Successive interference cancellation receiver processing with selection deversity
US7751506B22010-07-06Method for the soft bit metric calculation with linear MIMO detection for LDPC codes
US6898248B12005-05-24System employing threaded space-time architecture for transporting symbols and receivers for multi-user detection and decoding of symbols
EP2272194B12018-09-19Method and system for predicting channel quality index (cqi) values for maximum likelihood (ml) detection in a 2x2 multiple input multiple output (mimo) wireless system
CN101322328B2016-03-16The detection of complexity and the method for decoding and device is reduced for the receiver in communication system
AU2006223126C12010-09-16Systems and methods for beamforming in multi-input multi-output communication systems
US7508748B22009-03-24Rate selection for a multi-carrier MIMO system
TWI327427B2010-07-11Signal processing with channel eigenmode decomposition and channel inversion for mimo systems
US7379492B22008-05-27Adaptive modulation coding system and method in a mobile communication network
AU2006223127B22010-04-22Systems and methods for beamforming and rate control in a multi-input multi-output communication systems
ES2715398T32019-06-04Coding and modulation of multiple data flows in a communication system

Priority And Related Applications

Priority Applications (1)

ApplicationPriority dateFiling dateTitle
CN200610103860A2006-08-042006-08-04Iterative decoding algorithm for space hour bit interlaced modulating system and receiving system

Applications Claiming Priority (1)

ApplicationFiling dateTitle
CN200610103860A2006-08-04Iterative decoding algorithm for space hour bit interlaced modulating system and receiving system

Legal Events

DateCodeTitleDescription
2007-01-03PB01Publication
2007-01-03C06Publication
2007-02-28SE01Entry into force of request for substantive examination
2007-02-28C10Entry into substantive examination
2010-02-10GR01Patent grant
2010-02-10C14Grant of patent or utility model
2014-10-01CF01Termination of patent right due to non-payment of annual fee
Granted publication date: 20100210

Termination date: 20130804

2014-10-01C17Cessation of patent right

[19] 中华人民共和国国家知识产权局

[51] Int. Cl.
H04L 27/26 (2006.01)

[12] 发明专利说明书

专利号 ZL 200610103860.3

[45] 授权公告日 2010 年 2 月 10 日

[11] 授权公告号 CN 100589469C

[22] 申请日 2006.8.4
[21] 申请号 200610103860.3
[73] 专利权人 华为技术有限公司
地址 518129 广东省深圳市龙岗区坂田华为总部办公楼
共同专利权人 西安电子科技大学
[72] 发明人 李 颖 李 杰
[56] 参考文献
CN1579077A 2005.2.9
EP1453262A1 2004.9.1
WO03092206A1 2003.11.6
串联级联空时码的研究与设计. 李颖. 西安电子科技大学博士学位论文. 2005
审查员 李美丽

[74] 专利代理机构 北京德琦知识产权代理有限公司
代理人 张颖玲 王 琦

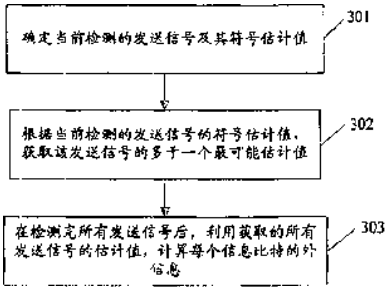
权利要求书 3 页 说明书 17 页 附图 5 页

[54] 发明名称

空时比特交织调制系统的迭代译码方法及 SISO 解调器

[57] 摘要

本发明公开了一种空时比特交织调制系统的迭代译码算法，首先确定当前检测的发送信号及其符号估计值；根据当前检测的发送信号的符号估计值，确定该发送信号的多于一个最可能的估计值；在检测完所有发送信号后，利用获得的所有发送信号的估计值，计算每个信息比特的外信息。本发明同时还公开了一种空时比特交织调制系统的接收系统，该系统的软输入软输出 SISO 解调器包括：符号估计值模块、星座点估计模块以及外信息计算模块。本发明的迭代译码算法与现有的 ML 迭代译码算法相比，复杂度可降低接近一半，而性能的损失在允许的范围内。利用本发明的迭代译码算法及接收系统可以在降低算法复杂度的同时，获得较好的性能。



1、一种空时比特交织调制系统的迭代译码方法，其特征在于，该方法包括以下步骤：

A、确定当前检测的发送信号及其符号估计值；

B、利用步骤 A 确定的符号估计值，获取当前检测的发送信号的多于一个最可能的估计值；

其中，所述获取当前检测的发送信号的多于一个最可能的估计值包括：

将信号星座图划分成一个以上区域，并根据当前检测的发送信号的符号估计值，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值；

C、在检测完所有发送信号后，利用步骤 B 获取的所有发送信号的估计值，计算每个信息比特的外信息。

2、如权利要求 1 所述的方法，其特征在于，步骤 A 所述确定当前检测的发送信号及其符号估计值包括：

A1、利用预定的算法计算接收端所有未检测的发送信号的加权向量；所述预定的算法为迫零算法、或最小均方误差算法；

A2、根据步骤 A1 获得的加权向量，确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号；

A3、利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值。

3、如权利要求 1 所述的方法，其特征在于，所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值为：

根据当前检测的发送信号的符号估计值的相位，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值。

4、如权利要求 1 所述的方法，其特征在于，所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值为：

根据当前检测的发送信号的符号估计值的实部和/或虚部的大小，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值。

5、如权利要求 4 所述的方法，其特征在于，所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值后，进一步包括：

根据当前检测的发送信号所处的区域，将信号星座图划分为一个以上层，再根据需要增选的星座点个数，从各层中选择与该发送信号的符号估计值距离最近的星座点。

6、如权利要求 5 所述的方法，其特征在于，所述从各层中选择与该发送信号的符号估计值距离最近的星座点为：从各层中选择相同数量的与当前检测的发送信号的符号估计值距离最近的星座点；或者从与距离已选择的星座点越近的层中选择越多的星座点。

7、一种软输入软输出 SISO 解调器，其特征在于，所述 SISO 解调器包括：符号估计值模块、星座点估计模块以及外信息计算模块，

符号估计值模块，用于确定当前检测的发送信号及其符号估计值，并将确定的符号估计值发送给星座点估计模块；

星座点估计模块，用于根据接收到的符号估计值，获取并存储当前检测的发送信号的多于一个最可能的估计值，包括：将信号星座图划分成一个以上区域，并根据当前检测的发送信号的符号估计值，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值；并将存储的所有发送信号的估计值发送给外信息计算模块；

外信息计算模块，利用接收到的所有发送信号的估计值计算每个信息比特的外信息。

8、如权利要求 7 所述的 SISO 解调器，其特征在于，所述符号估计值模块包括：加权向量模块和加权处理模块，

加权向量模块，用于利用预定的算法确定接收端所有未检测的发送信号的加权向量，并根据所有未检测的发送信号的加权向量确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号，并将选择的结果发送给加权处理模块；所述预定的算法为迫零算法、或最小均方误差算法；

加权处理模块，利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值，将其发送给星座点估计模块。

9、如权利要求 7 或 8 所述的 SISO 解调器，其特征在于，所述星座点估计模块包括：星座点确定模块，

星座点确定模块，用于将信号星座图划分成一个以上区域，并根据接收到的当前检测的发送信号的符号估计值选择最有可能的多于一个的星座点。

10、如权利要求 9 所述的 SISO 解调器，其特征在于，所述星座点估计模块进一步包括：分层模块，用于根据当前检测的发送信号所处的区域，将信号星座图划分为一个以上层，再根据需要增选的星座点个数，从各层中选择与当前检测的发送信号的符号估计值距离最近的星座点。

空时比特交织调制系统的迭代译码方法及SISO解调器

技术领域

本发明涉及空时编码技术，特别涉及一种空时比特交织调制系统的迭代译码算法及接收系统。

背景技术

基于多天线阵的多输入多输出(MIMO: Multiple-Input Multiple-Output)技术能够在空间上提供相互独立的多个并行子信道，有效地提高信息速率或通信质量，开拓了一个崭新的研究领域和设计理念。空时编码是一种适合多天线系统的、可逼近多天线信道容量的新型信道编码方案。

将传统二进制信道编码技术与空时编码技术级连，并通过内外码设计，可获得逼近多天线信道容量的性能。根据所采用的空时编码方法的不同，现有级连方案可分为三类，即级连网格（Trellis）空时码、级连分组空时码和级连分层空时码。其中，级连分层空时码也称为空时比特交织调制（ST-BICM）系统。级连Trellis空时码和级连分组空时码存在的一个共同的问题是系统频谱利用率的提高相对比较困难。

与其它两种级连方案相比，级连分层空时码利用传统信道编码技术提供分集增益和编码增益，同时利用分层空时（BLAST）方式提供复用增益，可以在保证性能的前提下获取更高的频谱利用率，是较为有前途的解决方案之一。

ST-BICM 系统常用的软输入软输出(SISO, Soft-Input Soft-Output)迭代译码算法中，SISO解调器根据 M 个接收信号和比特先验信息 $L^b_{A,demo}$，计算每个编码比特的外信息 $L^b_{E,demo}$，即对数似然比，再将对应所有编码比特的外信息经解交织处理后，作为先验信息 $L^b_{A,out}$ 送入外码译码器；外码译码器利用接

收到的先验信息计算序列 u 中每个比特的外信息 $L^b_{E,out}$，经交织后作为先验信息 $L^b_{A,demo}$ 送入 SISO 解调器，从而完成一次迭代过程。第一次迭代时，比特先验信息 $L^b_{A,demo}$ 初始化为 0。

采用 N 根发射天线，M 根接收天线的 ST-BICM 系统的发射系统结构示意图，如图 1（a）所示。信源产生的信息序列 a 首先送入外码编码器得到编码序列 u，序列 u 经交织器进行交织处理后输入到调制器，调制器对接收到的序列进行调制，然后发送到级连分层空时码模块，经空间映射后得到 N 个并行的子数据流，分别对应 N 根天线发送出去。

设第 i 根发射天线在 t 时刻的发送信号为 c_i^t，第 i 根发射天线与第 j 根接收天线之间的信道衰落系数为 $\alpha_{i,j}$，则第 j 根接收天线上对应的接收信号 r_j^t 可表示为：

$$r_j^t = \sum_{i=1}^{N} \alpha_{i,j} c_i^t + \eta_j^t \tag{1}$$

其中，η_j^t 为第 j 根接收天线上的信道加性噪声，其服从均值为零、方差为 N_0 的复高斯分布。为论述简单，本申请中假设信道是平坦衰落，且不考虑天线之间的相关性，即对于不同的 i 和 j，$\alpha_{i,j}$ 是统计独立的。

ST-BICM 系统采用级连分层空时码。ST-BICM 系统所采用的信道编码技术以及 SISO 迭代译码算法是影响其系统性能和复杂度的两个关键因素。参见图 1（b），图 1（b）为 ST-BICM 系统的接收系统结构示意图。如图 1（b）所示，ST-BICM 系统的 SISO 迭代译码器主要是由外码译码器和 SISO 解调器两部分构成，同时在外码译码器和 SISO 解调器之间通过交织器和解交织器互相交换外信息来实现译码迭代。目前常用的 SISO 解调算法的基本原理是：利用所有可能的符号序列计算每个比特的对数似然比以及外信息，计算的复杂度随发射天线数和调制阶数的增加而指数增加。

下面介绍常用的最大似然（ML）SISO 迭代译码算法，并对 ML SISO 迭代译码算法的计算复杂度进行分析。假设信号星座中共 2^b 个点，星座符号

集为$\{x_i\}_{i=1}^{2^b}$，第 i 根天线在某时刻发送的信号为 c_i，则对应接收天线 j 收到的信号为：

$$r_j = \alpha_{1,j}c_1 + \alpha_{2,j}c_2 + \cdots + \alpha_{N,j}c_N + \eta_j \tag{2}$$

当发射天线数为 N 时，接收天线 j 在每时刻收到的信号 r_j 中包含有 bN 比特的信息，将构成 $c_1, c_2, \cdots, c_N$ 的所有 bN 比特信息表示为：

$$\boldsymbol{b} = (b_1, \cdots, b_b, b_{b+1}, \cdots, b_{bN}) \tag{3}$$

其中 $b_{b(i-1)+1}, \cdots, b_{bi}$ 表示第 i 根发射天线发送的星座点 $c_i, i = 1, 2, \ldots, N$.

序列 $\boldsymbol{b}$ 中的第 l 比特信息 b_l 的对数似然比可通过下式计算

$$\Lambda(b_l) = \log \frac{\Pr[b_l = 1 \mid r_1, \cdots, r_M]}{\Pr[b_l = 0 \mid r_1, \cdots, r_M]} = \log \frac{\Pr[b_l = 1, r_1, \cdots, r_M]}{\Pr[b_l = 0, r_1, \cdots, r_M]}$$

$$= \log \frac{\sum_{\boldsymbol{b}: b_l = 1} \Pr[r_1, \cdots, r_M, \boldsymbol{b}]}{\sum_{\boldsymbol{b}: b_l = 0} \Pr[r_1, \cdots, r_M, \boldsymbol{b}]} = \log \frac{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 1} \Pr[r_1, \cdots, r_M, \boldsymbol{c}]}{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 0} \Pr[r_1, \cdots, r_M, \boldsymbol{c}]} \tag{4}$$

其中 $\boldsymbol{c} = (c_1, c_2, \cdots, c_N)$，$f(\boldsymbol{b})$ 为从 $\boldsymbol{b}$ 到 $\boldsymbol{c}$ 的映射，上式可进一步写为：

$$\Lambda(b_l) = \log \frac{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 1} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \Pr[c_1, \cdots, c_N]}{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 0} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \Pr[c_1, \cdots, c_N]} \tag{5}$$

假设各发送信号独立，可得

$$\Lambda(b_l) = \log \frac{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 1} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \prod_{i=1}^{N} \Pr[c_i]}{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 0} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \prod_{i=1}^{N} \Pr[c_i]} \tag{6}$$

由于使用了比特交织器，可认为构成发送信号的各个比特统计独立，即

$$\Pr[\boldsymbol{c}] = \prod_{k=1}^{b} \Pr[b_k] \tag{7}$$

由式(6)(7)以及比特外信息的概念，第 l 个比特的译码外信息可表示为：

$$\Lambda(b_l) = \log \frac{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 1} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \prod_{k=1, k \neq l}^{bN} \Pr[b_k]}{\sum_{\boldsymbol{c}: \boldsymbol{c} = f(\boldsymbol{b}), b_l = 0} \Pr[r_1, \cdots, r_M \mid c_1, \cdots, c_N] \prod_{k=1, k \neq l}^{bN} \Pr[b_k]}$$

$$= \log \frac{\sum_{c:c=f(b),b_l=1} \prod_{j=1}^{M} \Pr[r_j \mid c_1,\cdots,c_N] \prod_{k=1,k\neq l}^{bN} \Pr[b_k]}{\sum_{c:c=f(b),b_l=0} \prod_{j=1}^{M} \Pr[r_j \mid c_1,\cdots,c_N] \prod_{k=1,k\neq l}^{bN} \Pr[b_k]} \tag{8}$$

由

$$\Pr(r_1,\cdots,r_M \mid c_1,\cdots,c_N) = \frac{1}{\sqrt{\pi N_0}} \exp\left(-\frac{\sum_{j=1}^{M} \left| r_j - \sum_{i=1}^{N} \alpha_{i,j} c_i \right|}{N_0} \right) \tag{9}$$

式(8)可进一步写为:

$$\Lambda(b_l) = \log \frac{\sum_{c:c=f(b),b_l=1} \prod_{j=1}^{M} \exp\left(-\frac{\left| r_j - \sum_{i=1}^{N} \alpha_{i,j} c_i \right|^2}{N_0} \right) \prod_{k=1,k\neq l}^{bN} \Pr[b_k]}{\sum_{c:c=f(b),b_l=0} \prod_{j=1}^{M} \exp\left(-\frac{\left| r_j - \sum_{i=1}^{N} \alpha_{i,j} c_i \right|^2}{N_0} \right) \prod_{k=1,k\neq l}^{bN} \Pr[b_k]} \tag{10}$$

式(10)即为 SISO 解调器输出的关于第 l 个比特 b_l 的输出外信息，在迭代过程中作为外码译码器的输入先验信息。

由式(10)可知，SISO 解调器在计算每个比特的外信息时，需要考虑所有 2^{bN} 种可能的序列，即计算复杂度随发射天线数和调制阶数指数增加，在某些情况下，如发送天线数很大或调制阶数过高，会产生不可实现的高复杂度。

现有技术的垂直分层空时（VBLAST）迭代译码算法是一种复杂度低、实现容易的算法。

BLAST 接收机是基于干扰抑制与干扰抵消的联合技术。由于信噪比（SNR）最大的信号的检测质量相对最好，同时对其它信号的检测带来的影响最大，所以 VBAST 算法的基本思想是：先计算 SNR 最大的符号的估计值，然后从接收信号中消除该符号对其余符号检测的影响，再从修正的接收信号中进行 SNR 次最大的符号的检测，然后再消除该符号对其余符号检测的影响。如此类推，SNR 最小的符号最后进行检测。这种检测顺序使 SNR 最小

的符号检测受益最大，从而使整个干扰抵消方法的性能最优。

下面以迫零垂直分层空时（ZF-BLAST）算法为例，简要说明 VBLAST 算法检测处理的流程：

首先，将信号模型简化为：$R = HS + N$，

初始化 G 矩阵：

$$G_1 = H^+，\ i = 1 \tag{a}$$

为了确定第 i 次检测的发送信号 k_i，需要求得仍未被检测的所有发送信号的 SNR，选择 SNR 最大的一个作为第 i 次检测的信号，其中：

$$k_i = \arg\min_{j \notin \{k_1, k_2, \cdots, k_{i-1}\}} \left\| (G_i)_j \right\|^2 \tag{b}$$

选出第 i 次检测的信号 k_i 所对应的迫零矢量 w_{k_i}：

$$w_{k_i} = (G_i)_{k_i}^T \tag{c}$$

其中，$(G_i)_{k_i}$ 表示矩阵 G_i 的第 k_i 行

利用公式（d）计算第 i 次检测的发送信号 k_i 对应的符号估计值 y_{k_i}：

$$y_{k_i} = w_{k_i}^T R_i \tag{d}$$

利用公式（e）对 k_i 对应的符号估计值 y_{k_i} 进行量化：

$$\hat{s}_{k_i} = Q(y_{k_i}) \tag{e}$$

利用公式（f）从接收信号中将已检测的发送信号 k_i 抵消；利用公式（g）更新信道矩阵，将信道矩阵中已经检测的发送信号 k_i 对应的列置为零，其它列矢量保持不变：

$$R^{i+1} = R^i - \hat{s}_{k_i} \cdot (H)_{k_i} \tag{f}$$

$$G_{i+1} = H_{\bar{k}_i} \tag{g}$$

然后，将检测次数置为 i+1，返回公式（b）~（h）继续计算，直至 $i = N_T + 1$ 时停止计算，得到所有发送信号的估计值：$\hat{s}_{k_1}$，$\hat{s}_{k_2}$，…，$\hat{s}_{k_{N_T}}$。

VBLAST 算法虽然实现简单，但是性能不够好。由于该类算法利用公式

（e）对k_i对应的符号估计值y_{k_i}进行量化的判决过程是硬判，当SNR较高时，利用公式（f）进行的对消带来的效果是正向的；当SNR较低时，可能会出现错误判决，基于错误判决，利用公式（f）进行的对消就是负向的，所以错误判决对性能影响很大。

发明内容

本发明的主要目的在于提供一种空时比特交织调制系统的低复杂度迭代译码方法，在降低迭代译码算法复杂度的同时，获得较好的性能。

本发明的另一主要目的在于提供一种软输入软输出SISO解调器，在降低迭代译码算法复杂度的同时，获得较好的性能。

为了实现上述主要目的的第一方面，本发明提供了一种空时比特交织调制系统的迭代译码方法，该方法包括以下步骤：

A、确定当前检测的发送信号及其符号估计值；

B、利用步骤A确定的符号估计值，获取当前检测的发送信号的多于一个最可能的估计值；

其中：所述获取当前检测的发送信号的多于一个最可能的估计值包括：

将信号星座图划分成一个以上区域，并根据当前检测的发送信号的符号估计值，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值；

C、在检测完所有发送信号后，利用步骤B获取的所有发送信号的估计值，计算每个信息比特的外信息。

步骤A所述确定当前检测的发送信号及其符号估计值包括：

A1、利用预定的算法计算接收端所有未检测的发送信号的加权向量；所述预定的算法为迫零算法、或最小均方误差算法；

A2、根据步骤A1获得的加权向量，确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号；

A3、利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值。

所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值为：根据当前检

测的发送信号的符号估计值的相位，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值。

所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值为：根据当前检测的发送信号的符号估计值的实部和/或虚部的大小，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值。

所述选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值后，进一步包括：根据当前检测的发送信号所处的区域，将信号星座图划分为一个以上层，再根据需要增选的星座点个数，从各层中选择与该发送信号的符号估计值距离最近的星座点。

所述从各层中选择与该发送信号的符号估计值距离最近的星座点为：从各层中选择相同数量的与当前检测的发送信号的符号估计值距离最近的星座点；或者从与距离已选择的星座点越近的层中选择越多的星座点。

为了实现上述发明目的的另一方面，本发明提供了一种软输入软输出SISO解调器，包括：

外码译码器、软输入软输出SISO解调器、解交织器和交织器，其特征在于，所述SISO解调器包括：符号估计值模块、星座点估计模块以及外信息计算模块，

符号估计值模块，用于确定当前检测的发送信号及其符号估计值，并将确定的符号估计值发送给星座点估计模块；

星座点估计模块，用于根据接收到的符号估计值，获取并存储当前检测的发送信号的多于一个最可能的估计值，包括：将信号星座图划分成一个以上区域，并根据当前检测的发送信号的符号估计值，选择信号星座图中与当前检测的发送信号的符号估计值距离最近的多于一个星座点作为该发送信号的多于一个最可能估计值；并将存储的所有发送信号的估计值发送给外信息计算模块；

外信息计算模块，利用接收到的所有发送信号的估计值计算每个信息比

特的外信息。

所述符号估计值模块包括：加权向量模块和加权处理模块，

加权向量模块，用于利用预定的算法确定接收端所有未检测的发送信号的加权向量，并根据所有未检测的发送信号的加权向量确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号，并将选择的结果发送给加权处理模块；所述预定的算法为迫零算法、或最小均方误差算法；

加权处理模块，利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值，将其发送给星座点估计模块。

所述星座点估计模块包括：星座点确定模块，

星座点确定模块，用于将信号星座图划分成一个以上区域，并根据接收到的当前检测的发送信号的符号估计值选择最有可能的多于一个的星座点。

所述星座点估计模块进一步包括：分层模块，用于根据当前检测的发送信号所处的区域，将信号星座图划分为一个以上层，再根据需要增选的星座点个数，从各层中选择与当前检测的发送信号的符号估计值距离最近的星座点。

本发明提供的空时比特交织调制系统的迭代译码算法及接收系统，结合ML-SISO 及 VBLAST 接收机，通过改变备选信号向量集合的大小来降低迭代译码算法的复杂度。比如，当采用 QPSK 调制 4 根发射天线时，若采用传统迭代译码算法，需要计算所有 4^4=256 种四维信号向量；若采用本发明提供的译码算法，当每根天线仅选取三个可能信号点时，只需计算 3^4=81 种四维信号向量，复杂度仅为传统算法的 1/3 左右。事实上，随着发送天线数和调制阶数的增加，本发明所提算法在复杂度上面的优势会更加明显。

参见图 2，图 2 为两发两收 ST-BICM 系统在不同译码算法下的性能比较示意图。这里，两发两收 ST-BICM 系统仿真参数为：

信道编码采用 1/2 码率的 64 状态卷积码；信道假设为独立衰落，且接收端可准确估计信道衰落系数；交织器长度为 1024，调制方式为 QPSK 调制；译码算法为本发明中提出的基于 ZF 的迭代译码算法，基于 MMSE 的迭代译码算法和现有的基于 ML 的迭代译码算法，迭代次数分别为 1 次和 2 次。

图 2 中曲线 21、22、23 分别为本发明中提出的基于 ZF 的迭代译码算

法，基于 MMSE 的迭代译码算法和现有的基于 ML 的迭代译码算法迭代 1 次时的性能曲线；曲线 24、25、26 分别为本发明中提出的基于 ZF 的迭代译码算法，基于 MMSE 的迭代译码算法和现有的基于 ML 的迭代译码算法迭代 2 次时的性能曲线。

对于本发明提出的基于 ZF 和 MMSE 的迭代译码算法，本仿真中每层选择 3 个信号点来构成备选信号向量，共计 9 个二维信号向量，而基于 ML 迭代译码算法要考虑所有 16 个二维信号向量，因此本发明提出的迭代译码算法的复杂度与 ML 迭代译码算法相比，有大幅度的降低。而分析图 2 可知，本发明提出的迭代译码算法的性能和分集增益与 ML 迭代译码算法相比差别很小，在允许范围内。

另外，本发明的基于 MMSE 检测准则的译码算法性能优于基于 ZF 检测准则的译码算法，这主要是由于 ZF 准则本身只考虑了信道衰落系数矩阵的影响，没有考虑信道加性噪声，存在较大的错误传播。

总之，本发明的空时比特交织调制系统的迭代译码算法采用 MMSE 检测或 ZF 检测，与现有的 ML 迭代检测算法相比，其算法复杂度可降低接近一半，而性能的损失较小，在允许的范围内。而本发明的空时比特交织调制系统的迭代译码算法与现有技术的 VBLAST 译码算法比较，译码性能有很大的提高。因此，利用本发明的空时比特交织调制系统的迭代译码算法及接收系统可以在降低算法复杂度的同时，获得较好的性能。

附图说明

图 1（a）为 ST-BICM 系统的发射系统结构示意图；

图 1（b）为 ST-BICM 系统的接收系统结构示意图；

图 2 为两发两收 ST-BICM 系统在不同译码算法下的性能比较示意图；

图 3 为本发明空时比特交织调制系统的迭代译码算法的流程示意图；

图 4 为本发明空时比特交织调制系统的迭代译码算法的一个较佳实施例的流程示意图；

图 5 为 QPSK 信号星座图；

图 6 为本发明 16QAM 信号星座图的九个子方型区域示意图；

图 7 为本发明 16QAM 信号星座图的四个带状区域示意图；

图 8 为本发明以子方型区域 1 为例给出的分层示意图；

图 9 为本发明以子方型区域 2 为例给出的分层示意图；

图 10 为本发明以子方型区域 5 为例给出的分层示意图；

图 11 为本发明以带状区域 1 为例给出的分层示意图；

图 12 为本发明空时比特交织调制系统的接收系统的结构示意图。

具体实施方式

为使本发明的目的、技术方案和有益效果更加清楚明白，下面结合实施例和附图，对本发明进一步详细说明。

本发明提供的空时比特交织调制系统的迭代译码算法及接收系统，首先确定当前检测的发送信号及其符号估计值；根据当前检测的发送信号的符号估计值，确定该发送信号的多于一个最可能的估计值；在检测完所有发送信号后，利用获得的所有发送信号的估计值，计算每个信息比特的外信息。

参见图 3，图 3 为本发明空时比特交织调制系统的迭代译码算法的流程示意图。如图 3 所示，该流程包括以下步骤：

步骤 301，确定当前检测的发送信号及其符号估计值。

这里，确定当前检测的发送信号及其符号估计值包括：

a1、利用预定的算法计算接收端所有未检测的发送信号的加权向量。

a2、根据步骤 a1 获得的加权向量，确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号。

a3、利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值。

步骤 302，利用步骤 301 确定的符号估计值，获取当前检测的发送信号的多于一个最可能的估计值。

步骤 303，在检测完所有发送信号后，利用步骤 302 获取的所有发送信号的估计值，计算每个信息比特的外信息。

参见图 4，图 4 为本发明空时比特交织调制系统的迭代译码算法的一个较佳实施例的流程示意图。如图 4 所示，该流程包括以下步骤：

步骤 401：利用预定的算法确定接收端所有未被检测的发送信号的加权向量。这里，预定的算法可以为迫零检测算法或最小均方误差算法。

设接收端关于所有 N 个发送信号形成的加权向量构成矩阵 W 为：

$$W=(w_1,w_2,\ldots,w_N)^{\mathrm{T}} \tag{11}$$

按照迫零检测算法，矩阵 $W_{ZF}=H^{+}$，这里，H 为信道矩阵，H^{+} 为矩阵 H 的广义逆矩阵。

按照最小均方误差算法，矩阵 W 的选择应该使误差的均方值 $E\{\|c-Wr\|^2\}$ 最小，对应的基于 MMSE 准则的接收天线加权向量矩阵为：

$$W_{MMSE}=H^{\mathrm{H}}\left(HH^{\mathrm{H}}+\sigma^2 I\right)^{+} \tag{12}$$

其中，σ^2 是接收信号中包含的噪声，H^{H} 是矩阵 H 的共轭转置，$\left(HH^{\mathrm{H}}+\sigma^2 I\right)^{+}$ 是矩阵 $\left(HH^{\mathrm{H}}+\sigma^2 I\right)$ 的广义逆矩阵。将 W_{MMSE} 的行向量 w_i 作为接收天线对于发送信号 c_i 所形成的加权向量。

步骤 402：根据加权向量 W_{ZF} 或 W_{MMSE} 确定所有未被检测的各发送信号的信噪比（SNB），选择其中 SNB 最大的信号作为当前次的检测信号。

具体的方法是：计算加权向量 $w_i(i=1,2,\ldots,N)$ 矩阵的长度，长度最短的说明对应发送信号受噪声干扰最小，信噪比最大，则应作为当前次检测的发送信号。

步骤 403：利用步骤 402 选择的当前次检测信号的加权向量对每根接收天线上的发送信号进行加权处理，得到当前次检测信号的估计值。

具体处理方法为：将公式(1)写为矩阵形式：

$$\mathbf{r}=\mathbf{Hc}+\boldsymbol{\eta} \tag{13}$$

假设 $\mathbf{r}=(r_1,r_2,\ldots,r_M)^{\mathrm{T}}$ 是 M 维接收信号向量，$\mathbf{c}=(c_1,c_2,\ldots,c_N)^{\mathrm{T}}$ 是 N 维发送信

号向量，$\eta=(\eta_1,\eta_2,\ldots,\eta_M)^T$是$M$维加性噪声向量，$H$为$M\times N$的信道矩阵。在本发明中，假设接收端可准确估计信道信息。

将当前次检测信号的加权向量$\mathbf{w}_{k_i}^T$同乘以式(13)两端，得到当前次检测的发送信号的符号估计值为：

$$y_{k_i}=\mathbf{w}_{k_i}^T\mathbf{r}=c_{k_i}+v_{k_i} \tag{14}$$

步骤404，根据当前次检测的发送信号的符号估计值，确定该发送信号的N_i个最可能的估计值。

这里，N_i为大于1的整数。

具体处理方法为：利用公式(14)，选择信号星座图中与当前次检测信号的估计值y_{k_i}距离最近的N_i个点作为发送信号c_{k_i}的N_i个估计值$\{\hat{c}_{k_i,1},\hat{c}_{k_i,2},\cdots,\hat{c}_{k_i,N_i}\}$。

步骤405，更新接收信号向量和信道矩阵，从接收信号向量和信道矩阵中消掉已检测的发送信号。

具体处理方法为：利用公式（f）和（g），在接收信号中将已检测信号抵消，并将信道矩阵中已检测信号对应的列置为零。

步骤406，判断当前的检测次数是否小于发送信号的数目N，如果小于，则将检测次数加一后，返回步骤401；否则，执行步骤407。

上述步骤405也可以放在步骤406后执行，即先判断当前的检测次数是否小于发送信号的数目，当判断当前的检测次数小于发送信号的数目时，先更新接收信号向量和信道矩阵，然后再返回步骤401。

步骤407，将获得的所有发送信号的估计值构成符号序列集合，记为$\hat{c}$。

假设第i次检测的信号对应的估计值为N_i个，则当共有N个发送信号时，符号序列集合$\hat{c}$由$N_1\times N_2\times\cdots\times N_N$个$N$维符号序列构成。其中，$N_1$、$N_2$、…$N_N$都是大于1的整数。

步骤408，利用步骤407构成的符号序列集合$\hat{c}$，以及公式（10）计算每个信息比特的外信息。

本发明针对不同调制方式的星座图的特点提供了几种符号序列的确定原则，即上述步骤 404 所述根据当前次检测的发送信号的符号估计值，选择信号星座图中与估计值距离最近的多于一个星座点作为该发送信号的最可能估计值的确定原则。

下面分别给出 MPSK 信号星座图和 MQAM 信号星座图的符号序列确定原则。

MPSK 信号星座图均匀地分布在单位圆上，因此可根据消除了乘性衰落因子和已检测发送信号影响的第 i 个修正信号 y_i 的相位，来确定与修正信号 y_i 距离最近的几个点。这里，第 i 个修正信号 y_i 就是第 i 次检测的发送信号的符号估计值，用 $\angle y_i$ 表示修正信号 y_i 的相位。

以 QSPK 信号星座图为例说明 MPSK 信号星座图的符号序列确定原则。参见图 5，图 5 为 QPSK 信号星座图。

对于 QPSK 信号，若选择 N_i=2，则：

当 $-\frac{\pi}{4} \le \angle y_i \le \frac{\pi}{4}$ 时，选择信号点 0 和 3 作为两个最可能的备选信号点；

当 $\frac{\pi}{4} \le \angle y_i \le \frac{3\pi}{4}$ 时，选择信号点 0 和 1 作为两个最可能的备选信号点；

当 $\frac{3\pi}{4} \le \angle y_i \le \frac{5\pi}{4}$ 时，选择信号点 1 和 2 作为两个最可能的备选信号点；

当 $\frac{5\pi}{4} \le \angle y_i \le \frac{7\pi}{4}$ 时，选择信号点 2 和 3 作为两个最可能的备选信号点。

对于 QPSK 信号，若选择 N_i=3，则：

当 $0 \le \angle y_i \le \frac{\pi}{2}$ 时，选择信号点 0、1 和 3 作为三个最可能的备选信号点；

当 $\frac{\pi}{2} \le \angle y_i \le \pi$ 时，选择信号点 0、1 和 2 作为三个最可能的备选信号点；

当 $\pi \le \angle y_i \le \frac{3\pi}{2}$ 时，选择信号点 1、2 和 3 作为三个最可能的备选信号点；

当 $\frac{3\pi}{2} \le \angle y_i \le 2\pi$ 时，选择信号点 0、2 和 3 作为三个最可能的备选信号点。

MQAM 信号星座图基本是方型的，可根据修正信号的实部和虚部的大小来选择最可能的几个信号点。

以 16QAM 为例说明 MQAM 信号星座图的符号序列确定原则。该符号序列确定过程的具体步骤包括:

1）将 16 个信号点划分成 9 个子方型区域或四个带状区域。参见图 6 和图 7，图 6 为本发明 16QAM 信号星座图的九个子方型区域示意图；图 7 为本发明 16QAM 信号星座图的四个带状区域示意图。

2）根据修正信号的大小，选择最有可能的四个信号点。分两种情况:

第一种情况: 当 $|\mathrm{Re}(y)| \le 3d$ 且 $|\mathrm{Im}(y)| \le 3d$ 时，d 为图 6 和图 7 中小正方形的边长的一半。根据修正信号所处的子方型区域的位置选择最有可能的四个信号点。例如: 若 $-d \le \mathrm{Re}(y) \le d$ 且 $-d \le \mathrm{Im}(y) \le d$，则选择区域 5 内的四个信号点；若 $-d \le \mathrm{Re}(y) \le d$ 且 $d \le \mathrm{Im}(y) \le 3d$，则选择区域 6 内的四个信号点。

第二种情况: 当 $|\mathrm{Re}(y)| > 3d$ 或 $|\mathrm{Im}(y)| > 3d$ 时，选择图 7 中合适的带状区域内的四个信号点。例如: 若 $\mathrm{Re}(y) > 3d$，则选择带状区域 1 内的四个信号点；若 $\mathrm{Re}(y) < -3d$，则选择带状区域 3 内的四个信号点；若 $\mathrm{Im}(y) > 3d$，则选择带状区域 2 内的四个信号点；若 $\mathrm{Im}(y) < -3d$，则选择带状区域 4 内的四个信号点。

为改进迭代译码算法的性能，可进一步增选信号点，具体步骤如下:

3）根据修正信号所处的区域，将 16QAM 的信号星座图划分成若干个层次，然后确定所增选的信号点，下面分三种情况讨论:

第一种情况: 当修正信号处于子方型区域 1、3、7 或 9 时，除选择对应的四个信号点外，将剩余的 12 个信号点划分成两层，图 8 以子方型区域 1 为例给出了分层示意图。然后根据要求的增选信号点个数计算修正信号与各个子层中信号点的距离，选择距离最近的作为增选的信号点。例如，若要求增选的信号点个数小于等于 5，可只计算修正信号与第一层上信号点之间的距离即可。

第二种情况：当修正信号处于子方型区域2、4、6或8之中时，除选择对应的四个信号点外，同时将剩余的12个信号点划分成两层，图9以子方型区域2为例给出了分层示意图。然后根据要求的增选信号点个数计算修正信号与各个子层中信号点的距离，选择距离最近的作为增选的信号点。例如，若要求增选的信号点个数小于第一层上的信号点数，可只计算修正信号与第一层上信号点之间的距离，选出距离最近的信号点。

第三种情况：当修正信号处于子方型区域5之中时，将剩余的12个信号点划分成如图10所示的两层，然后计算修正信号与各个子层中信号点的距离，选择距离最近的作为增选的信号点。例如，若要求增选的信号点个数小于第一层上的信号点数，可只计算修正信号与第一层上信号点之间的距离，选出距离最近的信号点。

第四种情况：当$|\mathrm{Re}(y)|>3d$或$|\mathrm{Im}(y)|>3d$时，除选择对应的四个信号点外，同时将剩余的12个信号点划分成3层，图11以带状区域1为例给出了分层示意图，然后计算修正信号与各个子层中信号点的距离，选择距离最近的作为增选的信号点。例如，若要求增选的信号点个数小于第一层上的信号点数，可只计算修正信号与第一层上信号点之间的距离，选出距离最近的信号点。

这里，分层的目的是为了减小计算量，尽可能减少所求欧式距离的个数，所述的欧式距离即发射符号估计值与星座点之间的距离。首先将MQAM星座图分块，然后通过判断修正信号的实部虚部所处的位置，可以得到距离最近的4个星座点；如果所需要的星座点个数大于4，再将剩余的12个星座点分层。分层的原则是尽可能求最少的欧式距离。

本发明结合MMSE检测和ZF检测，提出了一种低复杂度的迭代译码算法，与现有ML迭代译码算法相比，复杂度可降低接近一半，而性能的损失在可接受的范围内。

为了进一步减少性能损失，本发明还提出了如下提高性能的方法：

由于最先检测出的几层信号点对性能的影响较大，因此在各层选择信号

点时，采用逐层减少信号点选择个数的方法来提高性能。比如，对于三根发送天线和 QPSK 调制的情况，可考虑 4、3、3 的检测模式，即在第一层保留所有 4 个信号点，在第二和第三层减少选择的信号点数，共考虑 36 个二维信号向量。

参见图 12，图 12 为本发明空时比特交织调制系统的接收系统的结构示意图，该系统包括：外码译码器、软输入软输出 SISO 解调器、解交织器和交织器。

其中，SISO 解调器包括：符号估计值模块 1201、星座点估计模块 1202 以及外信息计算模块 1203，

符号估计值模块 1201，用于确定当前检测的发送信号及其符号估计值，并将确定的符号估计值发送给星座点估计模块 1202；

星座点估计模块 1202，用于根据接收到的符号估计值，获取并存储当前检测发送信号的多于一个最可能的估计值，并将存储的所有发送信号的估计值发送给外信息计算模块 1203；

外信息计算模块 1203，利用接收到的所有发送信号的估计值计算每个信息比特的外信息。

所述符号估计值模块 1201 由加权向量模块和加权处理模块组成。

加权向量模块，用于利用预定的算法确定接收端所有未检测的发送信号的加权向量，并根据所有未检测的发送信号的加权向量确定所有未被检测的发送信号的信噪比，选择信噪比最大的发送信号作为当前检测的发送信号，并将选择的结果发送给加权处理模块。

加权处理模块利用选择的当前检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前检测的发送信号的符号估计值，将其发送给星座点估计模块 1202。

所述星座点估计模块 1202 包括星座点确定模块，该星座点确定模块用于将信号星座图划分成一个以上区域，并根据接收到的当前检测的发送信号的符号估计值选择最有可能的多于一个的星座点。

所述星座点估计模块 1202 还可以进一步包括分层模块，该分层模块用于根据当前检测的发送信号所处的区域，将信号星座图划分为一个以上层，再根据需要增选的星座点个数，从各层中选择与当前检测的发送信号的符号估计值距离最近的星座点。

由以上实施例可见，本发明提供的空时比特交织调制系统的迭代译码算法及接收系统，结合 ML-SISO 及 VBLAST 接收机，通过改变备选信号向量集合的大小，实现了在降低迭代译码算法复杂度的同时，获得较好的性能。

总之，以上所述仅为本发明的较佳实施例而已，并非用于限定本发明的保护范围。凡在本发明的精神和原则之内，所作的任何修改、等同替换、改进等，均应包含在本发明的保护范围之内。

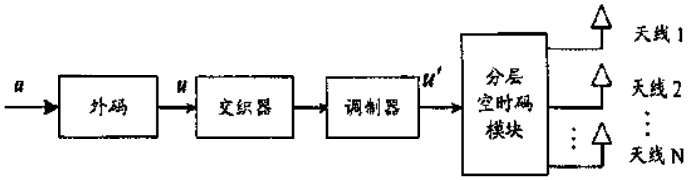

a
外码
u
交织器
调制器
u'
分层
空时码
模块
天线 1
天线 2
天线 N

图 1（a）

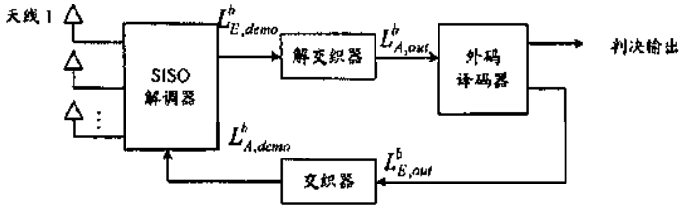

天线 1
SISO
解调器
$L^b_{E,demo}$
解交织器
$L^b_{A,out}$
外码
译码器
判决输出
$L^b_{A,demo}$
交织器
$L^b_{E,out}$

图 1（b）

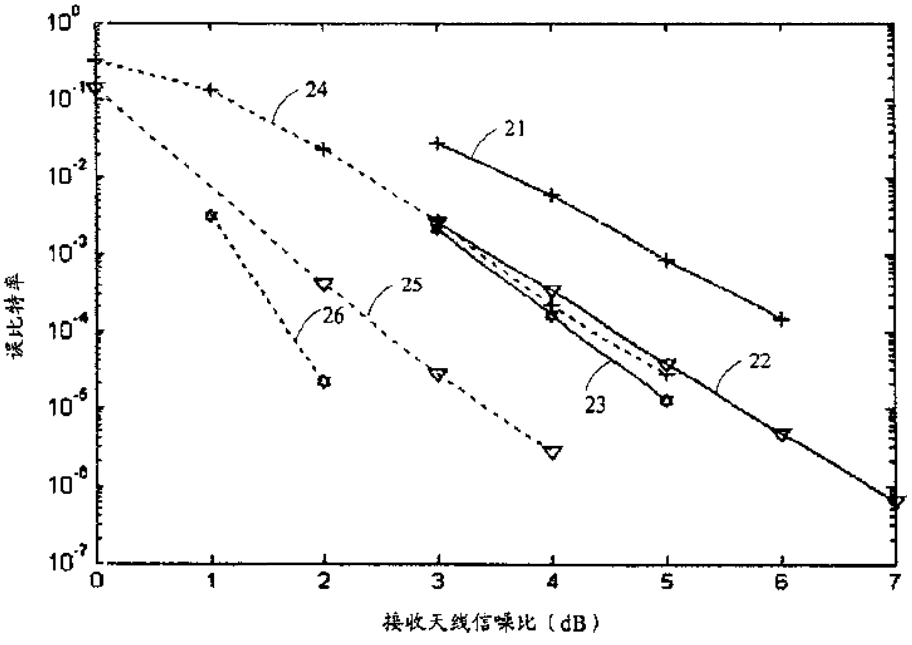

误比特率
接收天线信噪比（dB）
21
22
23
24
25
26

图 2

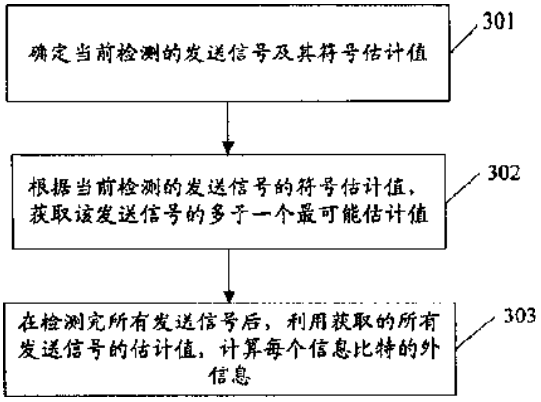

确定当前检测的发送信号及其符号估计值
301
根据当前检测的发送信号的符号估计值，获取该发送信号的多于一个最可能估计值
302
在检测完所有发送信号后，利用获取的所有发送信号的估计值，计算每个信息比特的外信息
303

图 3

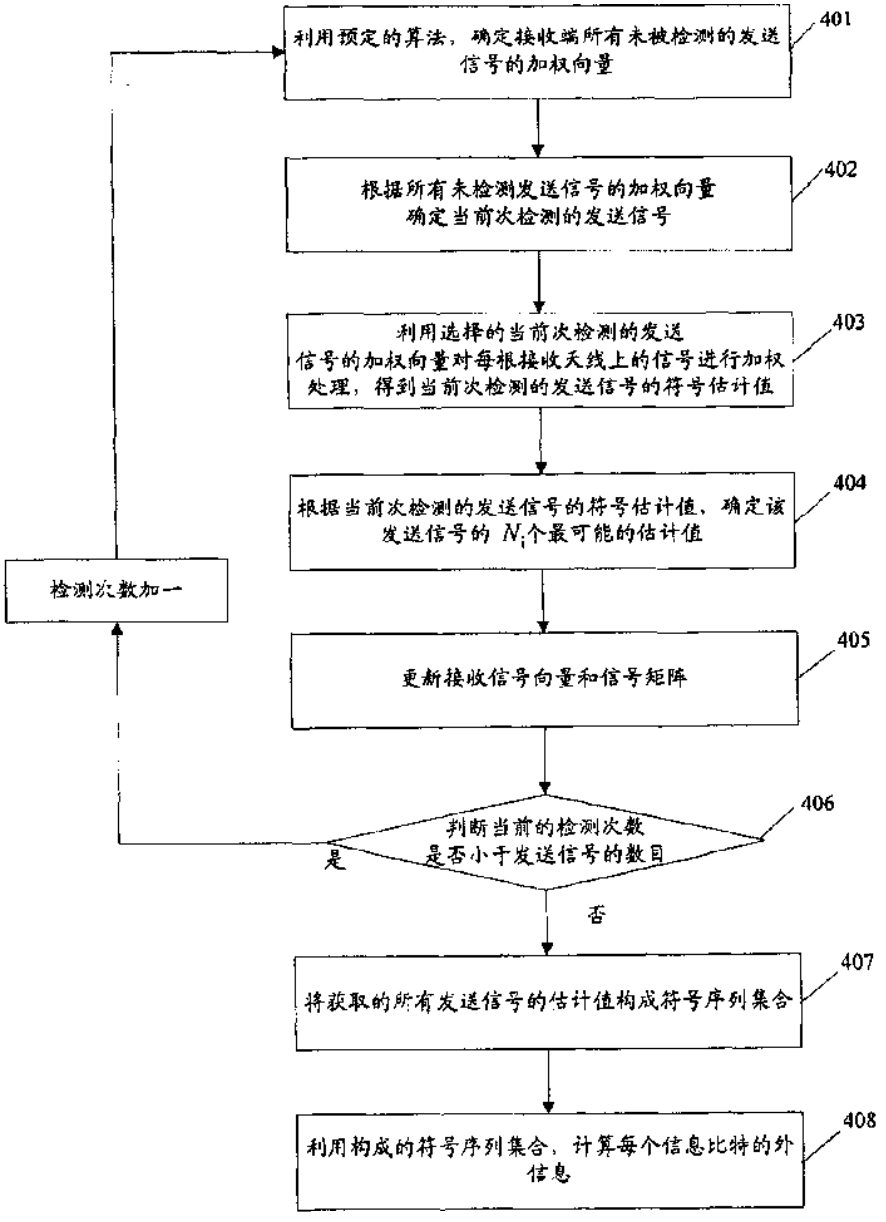

利用预定的算法，确定接收端所有未被检测的发送信号的加权向量
401
根据所有未检测发送信号的加权向量确定当前次检测的发送信号
402
利用选择的当前次检测的发送信号的加权向量对每根接收天线上的信号进行加权处理，得到当前次检测的发送信号的符号估计值
403
根据当前次检测的发送信号的符号估计值，确定该发送信号的 N_i 个最可能的估计值
404
更新接收信号向量和信号矩阵
405
判断当前的检测次数是否小于发送信号的数目
406
是
否
检测次数加一
将获取的所有发送信号的估计值构成符号序列集合
407
利用构成的符号序列集合，计算每个信息比特的外信息
408

图 4

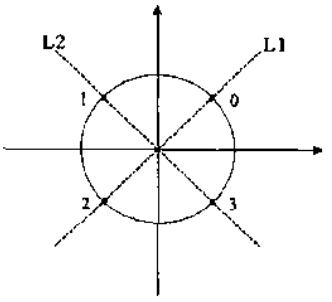


图 5

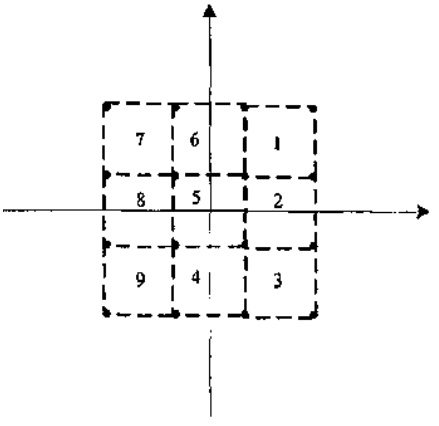


图 6

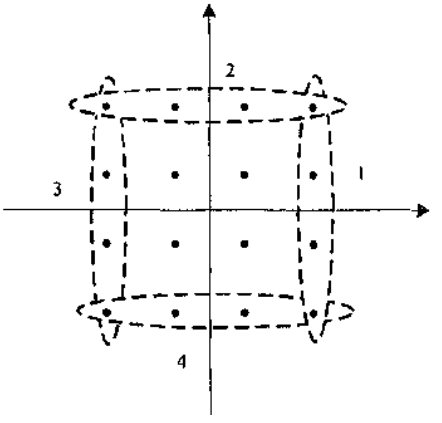


图 7

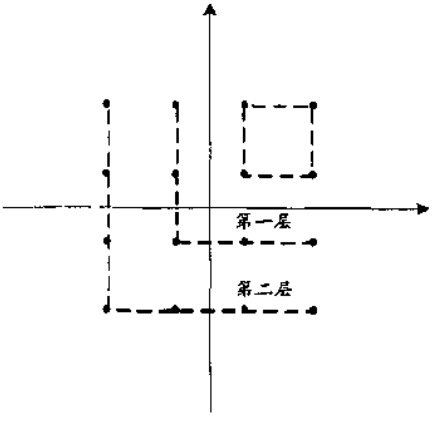


图 8

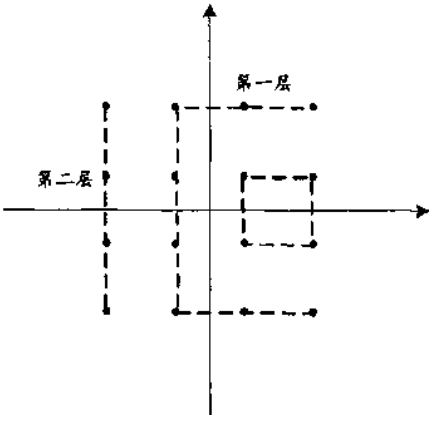


图 9

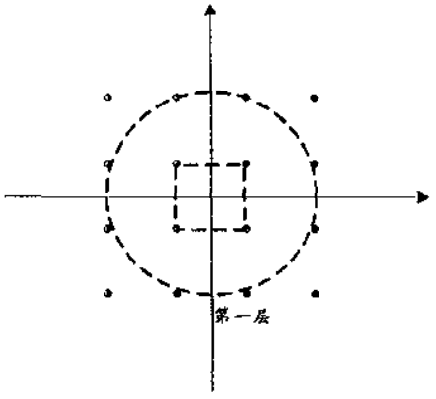

第一层

图 10

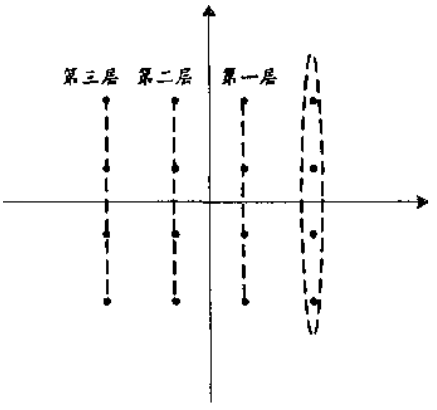

第三层
第二层
第一层

图 11

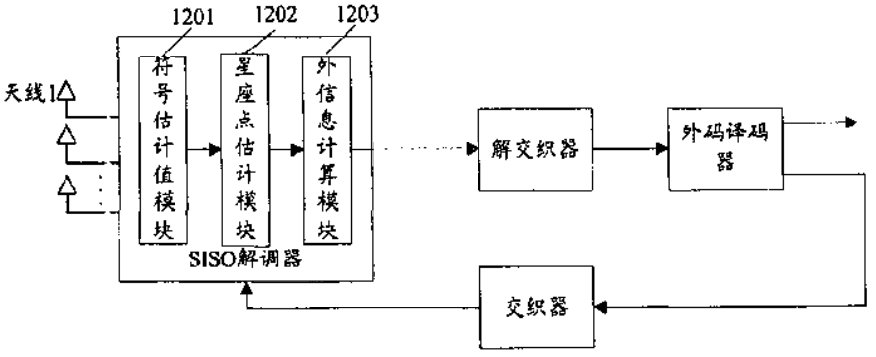

1201
1202
1203
天线1
符号估计值模块
星座点估计模块
外信息计算模块
SISO解调器
解交织器
外码译码器
交织器

图 12

JP4554610B2

Iterative decoding and equalization method for high-speed communication with multiple antenna channels between transmission and reception

Images (4)

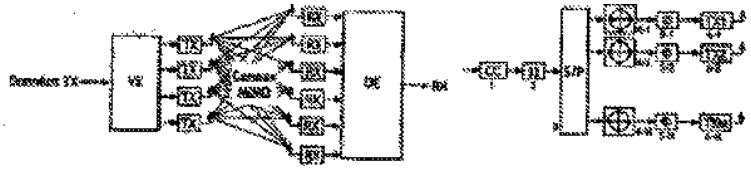


FIG.1

FIG.2

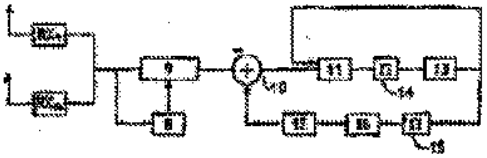


FIG.3

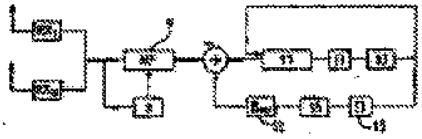


FIG.4

Classifications

H04L1/005 Iterative decoding, including iteration between signal detection and decoding operation

View 5 more classifications

Japan

Download PDF Find Prior Art Similar

Other languages

Japanese

Inventor

アントワーヌ・ベルテ

サミ・シュトゥル

ラファエル・ヴィゾズ

Worldwide applications

2003 ~~FR~~ 2004 WO EP JP US DE ~~AT~~ CN

Application JP2006524383A events

2003-08-29

Priority to FR0310290A

2004-08-06

Application filed by フランス・テレコム

2004-08-06

Priority to PCT/FR2004/002104

2007-03-01

Publication of JP2007504693A

2010-09-29

Application granted

2010-09-29

Publication of JP4554610B2

Status

Active

2024-08-06

Anticipated expiration

Info

- Patent citations (5)
- Cited by (20)
- Legal events
- Similar documents
- Priority and Related Applications

External links

Description

The present invention relates to the field of digital communications. The present invention more particularly relates to a method of iterative decoding and equalization for high speed communication over a plurality of transmit antenna channels, more particularly a frequency selective channel having a plurality of transmit and receive antennas. It is.

With respect to the development of antenna technology, communication models based on TDMA, CDMA, POFDM, and combinations thereof are systematically modified and extended to achieve MIMO channels. Two main types of space-time coding with different philosophies are envisioned:

The BLAST (Bell laboratory LAyered Space-Time architecture) transmission developed by Bell Laboratories aims to use spatial multiplexing to increase the bit rate that can be transmitted over the communication path. There are two ways to combine BLAST with error correction channel coding that gives two different hierarchical space-time coding models. In the first (ST-BICM) model, one encoding operation for concatenating various layers is spatio-temporal interleaving and spatial demultiplexing [Ref 1, LST-I, p. 1350]. In the second (MAC) model, the data is spatially demultiplexed and then each layer [Ref 1, LST-II, p. 1350] are independently encoded and interleaved.

Diversity schemes use space diversity to more effectively combat the effects of channel fading and improve performance. STTC schemes that improve coding, and STBC schemes that do not have a coding gain of 5 and are often combined with optimal trellis coded modulation known in the art as Gaussian channels (and their generalization) Some OD, LDC, etc.).

Recent theoretical contributions have demonstrated that improvements in spatial multiplexing and spatial diversity are linked by a compromise relationship. For example, the two above approaches cannot increase capacity at the same rate for arbitrarily high numbers M, N of transmit and receive antennas. The transmit diversity scheme is a MISO configuration {M, N} = {2, 1 which has a significant doubt on their validity for high bit rate

communication over MIMO channels using multiple antennas from an information theory standpoint. } Can only be shown to be optimized.

Only the ST-BICM approach guarantees a linear increase in capacity by min {M, N}.

This is why it is necessary to select this type of transmission and concentrate the design effort on a suitable receiver that guarantees performance as close as possible to the intrinsic limit. Unlike STBC, ST-BICM does not guarantee mutual orthogonality of data streams transmitted by various antennas.

For transmission over frequency selective MIMO channels, this coding strategy is used to cancel space domain multiple access interference (MAI) and time domain intersymbol interference (ISI). It therefore requires joint demodulation of the data at the receiver.

The joint data demodulation function, in particular, requires the receiver's precision because the ST-BICM inherently requires iterative processing between channel demodulation and decoding in order to obtain all of the diversity in the spatial domain. And complex elements.

"Iterative decoding on non-frequency selective MIMO channels" ST-BICM iterative decoding over frequency non-selective MIMO channels constitutes a special case of the problem to be solved under the assumption of two aspects of perfect ISI cancellation and is available for each symbol Recover all of the energetic energy.

In that case, the problem can be restated in the term “cancellation of MAI over a uniform MIMO channel in the presence of spatially colored noise” below. It always requires an APP calculation on the bits of the symbols of the transmitted vector (statistics on the data to be transmitted to the decoder).

Two methods are described in the literature.

-Accurate calculation of APP based on complete list (MAP).

Since the complexity of this method increases exponentially according to the number M of transmit antennas and is a polynomial function of the order Q of the constellation used, the complexity of this method is very expensive.

Approximate calculation of APP from candidate vector reduction list generated by sphere decoding (intrasphere decoding) algorithm [Ref 2].

The sphere decoding algorithm has at least two advantages: it preserves the MAP criterion and has a complexity in O (M3) that is not affected by the order Q of the constellation used. Note that the sphere decoding algorithm can be considered a variation of the Fano sequential decoding algorithm.

Approximate calculation of APP from the final candidate list of candidate vectors generated by the sequential stack algorithm.

Sphere decoding algorithms have been intensively studied for frequency non-selective MIMO channels. It is important to generalize to the situation of frequency selective MIMO channels.

A burdensome approach would be block iteration. This approach is a less interesting approach because it artificially increases the size of the search space (and thus increases complexity) and introduces redundant decoding of symbols.

The algorithm described in [Reference 2] has been derived for frequency non-selective MIMO channels. No conclusion has yet been reached on generalizing it to the situation of frequency selective MIMO channels.

The performance of these two algorithms is essentially determined by the quality of the generated candidate list, which is preferably a fixed size. In all respects, it is essential to include 1-bit optimal decoding and 0-bit optimal decoding.

"Iterative decoding over frequency selective MIMO channels" Several types of demodulators have been proposed that are inserted into the repetitive structure.

A demodulator with weighted inputs and outputs that applies a MAP criterion based on the BCJR algorithm.

The complexity of this receiver in O (QLxM) is clearly expensive for higher order modulation and MIMO channels with multiple inputs and high memory demand.

Suboptimal weighted input and output demodulation based on an efficient search algorithm in the trellis that is greatly reduced with respect to the number of states [ref. 3] and preceded by a minimum phase filter if necessary vessel.

This approach is still limited by its complexity in O (QLrxM). Here, Lr represents the reduced constraint length. In particular, the use of constellation with multiple states is excluded.

Replacing a non-linear demodulator composed of iterative linear cancellation of interference (see multi-user demodulation environment in [Ref. 4]) with a MIMO environment is generally a common point between different users and antennas. Is based. Their implementation therefore requires one filter per antenna and their performance is limited by the discontinuous MMSE per antenna approach, with the advantage that its complexity is a polynomial function of the number of transmit antennas. The S. Ariyavisitakul, "Turbo Space-Time Processing to Improve Wireless Channel Capacity," IEEE Trans. Commun., Vol. COM-48, no. 8, pp. 1347-1358, Aug. 2000. BM Hochwald, S. Ten Brink, "Achieving Near-Capacity on a Multiple-Antenna Channel," IEEE Trans. Commun., Vol. 51, no. 3, pp. 389-399, March 2003. R. Visoz, AO Berthet, "Iterative Decoding and Channel Estimation for Space-Time BICM over MIMO Block Fading Multipath AWGN Channel," IEEE Trans. Commun., Vol. COM-51, no. 8, pp. 1358-1367, Aug . 2003. X. Wang, HV Poor, "Iterative (Turbo) Soft-Interference Cancellation and Decoding for Coded CDMA," IEEE Trans. Commun., Vol. COM-47, no. 7, pp. 1046-1061, July 1999.

The object of the present invention is to propose a less complex advanced receiver for high bit rate digital transmission over a frequency selective channel with a plurality of transmit and receive antennas, in particular in the spatial domain. It is to propose a receiver that does not require a large amount of computational power while providing for inter-interference and inter-symbol interference processing in the time domain.

To achieve this object, the present invention proposes an iterative decoding and equalization apparatus for high bit rate communication over a frequency selective channel having a plurality of transmitting and receiving antennas. The apparatus comprises a decision feedback equalizer configured to receive data from different receive antennas, wherein the decision feedback equalizer is calculated from the output of the decoder supplied with the signal

by the decision means. A forward filter and a recursive backward filter supplied with weighted reconstruction data, and further means for subtracting the backward filter output data from the forward filter output data. Here, the subtracted data is supplied to the input of the determination means by the output of the decoder, and the determination means taking into account the spatial noise correlation in the output of the subtraction means is connected to the channel decoder having the weighted input and output. The statistics to be sent are generated and the decision means and the decoder are separated by a space-time interleaving process performed in the binary stage. Said device is characterized in that the forward and backward filters are iteratively adapted to minimize the mean square error at the output of the subtracting means.

A device of the aforementioned kind advantageously has the following supplementary features in all possible combinations, individually or technically.
The decision means at the output of the subtracting means of the equalizer is a spatial whitening type means, followed by a sphere decoder;
The determining means at the output of the equalizer subtracting means is a means of serial and / or parallel type (SIC / PIC) configured to cancel the remaining spatial interference at the output of the equalizer subtracting means; is there.
Spatial whitening is achieved at the output of the subtracting means of the equalizer.
-Spatial whitening is achieved by the determining means.
-Spatial whitening is achieved by forward and reverse filters.
Starting with an iteration, the forward filter becomes an adaptive filter and the backward filter becomes a similar adaptive filter with no center coefficient.

The present invention arises from a novel approach that is fundamentally different from the approach of considering an antenna as a different user. The transmitted signal is considered a T-dimensional modulation. Here, T is the number of transmission antennas. This modified aspect assumes that T-dimensional vector modulation is convoluted with the frequency selective channel at its input and provides adaptive information for the channel decoder. A T-dimensional modulation demodulator that can be generated (e.g., a sphere decoder) has a significant impact on the design of a receiver that is configured by a linear iterative equalizer connected thereafter. . This approach is less complicated because it requires one (vector) filter and

provides better performance because it allows the selection of vector demodulation criteria, It has two advantages over the prior art.

The invention likewise provides a high bit rate over a frequency selective channel having a plurality of transmit and receive antennas, characterized in that it comprises a receiver comprising an equalization and decoding device of the kind described above. Related to the system for communication.

The transmitting means is, for example, an advantageous ST-BICM type means.

The present invention further provides high bit rate communication over a frequency selective channel having multiple transmit and receive antennas using a decision feedback equalizer configured to receive data from different receive antennas. We propose an iterative decoding and equalization method. The decision feedback equalizer has a forward filter and a recursive backward filter that are supplied with weighted reconstruction data calculated from the output of the decoder supplied with the signal by the decision means, the forward direction Means are used for subtracting the output data of the backward filter from the output data of the filter. Here, the subtracted data is supplied to the input of the determination means by the output of the decoder, and the determination means taking into account the spatial noise correlation in the output of the subtraction means is connected to the channel decoder having the weighted input and output. The statistics to be sent are generated and the decision means and the decoder are separated by a space-time interleaving process performed in the binary stage. Said method is characterized in that the forward filter and the backward filter are iteratively adapted to minimize the mean square error at the output of the subtracting means.

Other features and advantages of the present invention will become apparent from the following description which is merely exemplary and not limiting with respect to the invention and should be read in conjunction with the accompanying drawings.

"BLAST technology and ST-BICM coding"
FIG. 1 is a general configuration diagram of a BLAST structure.

After space-time interleaving and spatial demultiplexing, data encoded by one vector encoder EV is transmitted by a plurality of transmit antennas TX. On the other side of the MIMO channel, the transmitted signals are received

by a plurality of receiving antennas RX which send them to decoding means DE (MIMO channel) whose data is restored at the output.

The following description relates to frequency selective MIMO channels. The presence of intersymbol interference increases the complexity of the receiver.

FIG. 2 shows a general communication model of ST-BICM.

In transmission, the data is error-corrected channel coded (1) (convolutional code, turbo code, LDPC code, etc.), interleaved in binary steps (2), spatially demultiplexed (3), and Modulated with respect to hierarchy (4-1 to 4-M). After shaping filter processing (5-1 to 5-M), the modulated data is sent to the transmitting antenna (6-1 to 6-M).

"Description of transmission processing and reception processing"
In transmission, the data undergoes ST-BICM processing.

This includes the following steps:
Receiving a digital data stream at a predetermined bit rate;
Applying error correction channel coding (1) to generate an encoded digital data stream;
Interleaving (2) the encoded digital data using a bit-level interleaver.
Spatial demultiplexing of the encoded and interleaved digital data streams to create a plurality of individual encoded digital data streams (M distinct streams or transmitter layers) called the transmitter layer (3) Step to perform.

The M distinguishable encoded digital data streams are then processed as follows.
Modulate (4-1 to 4-M) the encoded and interleaved digital data stream 15 according to a modulation scheme for obtaining a stream of modulated symbols.
Filter the modulated stream (5-1 to 5-M).
Send the modulated stream (6-1 to 6-M) via its own transmit antenna.

FIG. 3 shows a receiver of the transmission system.

It comprises a decision feedback equalizer defined by a forward filter 9, a subtractor 10, a decision algorithm 11, and a recursive backward filter 12.

The recursive backward filter 12 is supplied with weighted and reconstructed calculation data from the output of the decoder 13 to which the signal is supplied by the decision algorithm 11.

The forward filter 9 and the backward filter 12 are iteratively determined to minimize the mean square error (MSE) of the MIMO equalizer, that is, to minimize the error at the output of the subtractor 10. Is done.

To achieve this goal, they use an initial estimate 8 of the MIMO channel.

For example, the processing used is the type of processing shown in Appendix II.

In the light of the above processing, residual error and noise vector estimates are used, and forward and backward filters are calculated at the point of block calculation to minimize this vector. Caution must be taken.

This vector processing simplifies the calculation.

The input of the determination means 11 receives data from the output of the subtracter 10 and from the output of the decoder 13.

The algorithm can be a different kind of algorithm, in particular of the SIC / PIC type (serial and / or parallel cancellation of the remaining spatial interference at the output of the equalizer subtractor means-see Appendix III).

Alternatively, the algorithm can be a sphere decoder algorithm.

This type of algorithm has complexity in O (M^3). (Where M is the number of transmit antennas independent of Q.)

This allows the use of a modulation having multiple states for the purpose of increasing the bit rate.

The output of the judging means 11 is sent to a spatio-temporal deinterleaving process 14 which is performed in the binary stage between the judging means 11 and the decoder 13.

The output of the decoder 13 is a bit probability.

This probability is first sent to the input of the decision algorithm 11 and then sent to the weighted data reconstruction process 16 after performing spatio-temporal interleaving (15).

The reconstructed weighted data is sent to the input of the reverse filter 12.

The output of the reverse filter 12 constitutes weighted reproduction data.

Furthermore, the residual interference and noise injected at the input of the subtractor 10 and the corresponding errors are colored in both the time domain and the spatial domain.

Although temporal correlation has little effect on the processing performed by the equalizer, spatial correlation has an essential role.

This is why in one embodiment, spatial whitening is achieved using Cholesky factorization.

This spatial whitening is advantageously achieved at the output of the subtractor 10.

In one particular embodiment, it can be achieved by the sphere decoding algorithm 11 itself.

It can be equally integrated into the forward filter 9 and the backward filter 12.

Finally, in one embodiment, from a certain number of iterations, for example from the second iteration, the forward filter 9 is advantageously replaced with an adaptive filter.

Since the backward filter is estimated directly from the forward filter, convolution with the channel without the center coefficient of the adaptive filter thus occurs in this case.

This is shown in FIG. 4, where the forward filter 9 becomes an adaptive filter MF, while the backward filter $12B_{MF}$ becomes an adaptive filter convolved with a channel without a central coefficient.

"Appendix II"
"Communication model"

Consider P block channels with multiple inputs and outputs, T frequency selective transmit antennas with M memories, and R frequency selective receive antennas.

"A. Space-time binary interleaved code modulation" Suppose that C represents a linear code of length N_C and accepts the data vector "$u \in F_2^{\rho c N}$" and gives ρ_C for F_2 that generates the encoded word "$c \in F_2^N$". . If a convolutional code is used, the given ρ_C is assumed to include tail bits. The coded word is put into a well-designed binary interleaver ⊓, and its output matrix "$A \in F_2^{Tq * PL}$" is P matrices "$A^P \in F_2^{Tq * PL}$, p = 1, …, Divided into P ". The columns of the matrix A^P have T subvectors "$a^p{}_t$ [n] with stacked binary components" $a^p{}_{<t, j>}$ [n],... $A^p{}_{<t, j>}$ [n] ".] $\in F_2^q$, t = 1,…, T "(one for each input channel), a vector called" symbol label vector ""a^p [n] $\in F_2^{Tq}$, n = 1,… , L ". Here, "<t, j>" indicates the index "(t−1) q + j". In each matrix A^P, all the vectors "a^p [n]" are passed through bit coding with the symbol "μ: {0, 1} $^{qT} \rightarrow$ A" and the radix "| A | = 2^{qT} " The signal set "A 無 C^D " is modulated by a memoryless D-dimensional modulator. Without loss of generality, it is assumed that A is a simple product of T complex signal constellations C of the same order 2^q and that Gray coding is used for each signal constellation. After the transformation of the signal, this coded modulation process is also similar to the case where all the coded word complex symbols "P × L × T" are converted into P matrices "Xp $\in C^{T \times L}$, p = 1 , ..., P "are considered to be space-time coded modulation schemes. The column "x^p [n] $\in C^T$, n = 1,..., L" of the matrix "Xp" is called a "signal point arrangement symbol vector". The "symbol label vector" a^p [n] can be obtained from "x^p [n]" by a simple inversion of the bit coding with the symbol "μ^{-1} ". Classified into the general class of space-time codes, this architecture is called space-time bit-interleaved coded modulation (STBICM) [Ref 3]. From the ideal filtering conditions of the Nyquist limited band, the transmission given (in bits used per channel) is:

FIG. 1 shows a block diagram of this communication model.

"B. MIMO communication channel" It is assumed that "$H^p \in C^{R \times T \times (M + 1)}$" is a MIMO communication channel related to the block p. Here, it is assumed that "p = 1,..., P" and "H = {H^p}" is a set of all communication paths. Similarly, "$X^p \in C^{T \times L}$" and "$Y^p \in C^{R \times L}$" are assumed to be "signal point arrangement symbol matrix" and "communication path output matrix", respectively. The channel output of the

equivalent baseband vector at the discrete time “$y^{p}[n] \varepsilon C^{R}$” at time “n = 1,..., L” can be written as:

In the above equation, “$x^{p}[n] \in C^{T}$” is a “signal point constellation symbol vector” transmitted at time $\underline{n}$, in which the energy of each component is equal to “σ_{x}^{2}”. “$H^{p}[k] \in C^{R \times T}$” is a matrix of $\underline{k}$ taps of the impulse response of the communication path. “$W^{p}[n] \in C^{R}$” is an additional complex noise vector. The additional complex noise vector $w^{p}[n]$ is assumed to have zero and circularly symmetric complex Gaussian type averages that are independent of each other and follow the same distribution, and thus “pdf N $(0, \sigma^{2} I)$” It becomes. A channel H^{p} that is constant throughout the corresponding block has a finite impulse response (FIR) of length “M + 1” and its spatial symbol taps “$H^{p}[0],.^{p}[M]$ ”is a zero-mean random complex matrix“ R × T ”, and for equal power systems, the average power satisfies the normalization restrictions shown below.

Here, the operator “†” corresponds to a conjugate transpose operator. Assuming that the number of individual multiple channel components is mainly determined by large buildings and reflecting objects, the same channel memory for all possible “R × T” connections is reasonable. It is an assumption.

"Repeated block of MIMO interference cancellation"
“A. Theory and scheduling”
The iterative block of the MIMO interference canceller processes each received data block “Y^{p}, p = 1,..., P” using the random probabilistic information fed back by the output decoder separately. All described signals and elements are therefore indexed on a block-by-block basis. This dependency state is sometimes omitted to simplify the notation. During each iteration I, the linear forward filter F^{l} applied to the received symbol vector y [n] to generate a vector signal $y^{l}[n]$. A well-structured estimate $e^{l}[n]$ of the MAI and of the ISI that degrades x [n] is subtracted from the vector signal $y^{l}[n]$ to produce $z^{l}[n]$.

Given the preliminary knowledge (ancillary probability distribution) available on iteration I-1, the estimate of vector $e^{l}[n]$ is an adaptive decision trial vector $x^{l\ for}$ vector x [n]. Provided from the output of the backward filter B^{l} excited by [n].

The discussion will now consider the core of the MIMO turbo-equalizer, namely the derivation of the MIMO forward finite impulse response filter and the MIMO backward finite impulse response filter.

"B. Calculation of forward and reverse filters"
The description begins with several basic operations of the instantaneous output $y^{l}[n]$ of the forward filter F^{l}.

So $y^{l}[n]$ for each sample can be expanded as follows:

The equivalent matrix notation for convolution of two filters, filter H_{C} and filter F^{l} is as follows:

here,

Is a forward filter of order "LF = LF1 + LF2 + 1". here,

Is the Toeplitz diagonal channel matrix,

Is the transmitted vector. Introducing a combined filter:

And the following final expression is obtained:

Here, "$L_{G1} = L_{F1}$ ", "$L_{G2} = L_{F2} + M$", and "$L_{G} = L_{F} + M$".

By analogy, the backward filter is defined as follows:

The output of the scramble correction circuit can be written as follows:

The noise and the remaining scramble vector $v^{l}[n]$ (also referred to as a MIMO equalizer error vector) are expressed as follows:

here,

It is. Filter F^{l} and Filter B^{l} are calculated in blocks to minimize the mean square error of the MIMO equalizer under the restriction of "$G^{l}[0] = I$" and in the MIMO situation. The problem can be briefly formulated as follows:

MSE minimization is performed in two successive steps and is finally given as:

here,

And where

Furthermore, it is expressed as follows.

It should be emphasized that the noise plus the remaining scrambling vectors at the output of the subtractor is spatially and temporally correlated. Temporal correlation has no real impact on the following processing, but spatial correlation has an important role in this regard. Spatial white noise can be easily obtained by a simple Cholesky factorization of the correlation matrix $K^{I}{}_{v.}$

Proposal 2: Since the correlation matrix $K^{I}{}_{v}$ is defined as positive, Cholesky factorization can always be applied. This can be written as:

Here, L is a lower triangular matrix. Then, by applying "$F^{I,} = L^{-1} F^{I}$" and "$B^{I,} = L-^{1} B^{I}$" as the forward filter and the backward filter, the correlation of the scrambled vector added with noise is applied. The matrix is equal to the identity matrix.

Proof: Proof is self-evident.

The validity of the proposed algorithm is verified by the simulation of FIG.

In the following, the equalizer output is always considered with spatial whitening, as in the following equation:

Here, "$K_{v}{}^{I} = I$" and "$G^{I}[0] = L^{-1}$".

It is interesting to note that the MMSE criterion corresponds to maximizing the SNR (SNR) "tr $\{G^{I}[0]^{+} K_{v}{}^{I} G^{I}[0]\}$". The maximum SNR given for the adaptive filter is achieved for "$\sigma_{x}{}^{2} = \sigma_{x}{}^{2}$".

"C. Comprehensive APP decoder"
All stochastic quantities exchanged during the iteration are defined below. The purpose of the demodulation part of the MIMO equalizer is to provide incidental information about the bits of the symbols in matrix A according to the new communication model (10). Here, "$G^{I}[0]$" serves as a uniform MIMO channel equivalent to "T × T", and "$v^{I}[n]$" is the remaining interference with white noise added. is there. The log ratio APP for all bits of the symbol "$a_{<t, j>}[n]$" is defined as follows.

By simple abbreviation, the following equation is obtained.

here,

It is. The log ratio APP can be extended as follows.

Assume that perfect spatio-temporal interleaving allows spreading the probabilities previously added to the product of their bounded numeric components, as shown below.

Assuming that $v^{l}[n]$ has been spatially whitened by any means (eg, Cholesky factorization), the Euclidean metric can be used to estimate probabilities as follows:

In the general case, the correlation matrix $K^{l}{}_{v}$ noise vector, must be considered in assessing the probability as follows.

As established in the field of turbo demodulators, an incidental log probability rate is deduced from the equations.

All logarithmic incidental information samples provided by the MIMO demodulator are collected from all the blocks "p = 1,..., P", and after the spatio-temporal deinterleaving $^{-1}$, all the output decoders are Is relocated to a simple observation vector "$\zeta^{l,c} \in R^{N}$" as shown below on the assumption that a logarithmic incidental probability ratio for the coded word bits of

After space-time deinterleaving [pi, vector xi] ^{l,c} is one is intended for each block of data $A^{p,}$ disseminated into P matrices [pi ^{l,p} previous log probability rate. For each block "p = 1,..., P", the input for $\Pi^{l,p}$ is given by the following equation:

This means that the soft decision vector $x^{l}[n]$ can be rewritten as:

"D. Sub-optimal MIMO demodulation with sphere-decoded modified list" In "O (2^{qT})", the point count of the entire subset of signal constellations "$\Omega^{<0>}{}_{<t,\ j>}$" and "$\Omega^{<1>}{}_{<t,\ j>}$" whose radix varies is high. Optimal MIMO demodulation cannot be preserved when high bit rate communication scenarios occur because the next modulation and / or multiple transmit antennas quickly overpower. Careful analysis of the probability values shows that those large numbers are insignificant. Thus, as a significant contribution of the present invention, we have a much smaller subset "$L^{<0>}{}_{<t,\ j>}$" and

“$L^{<1>}$, also referred to as a list, which contains detailed point counts only with non-negligible probabilities. We propose to exchange for the point count of $_{<t, j>}$ ”. In that case, the logarithmic incidental probability ratio is as follows.

In the geometric representation, they include grid points in a sphere of radius r centered around a carefully selected point (eg, the unrestricted point ML, or the point ML itself). A modified version of the Sphere Decoder was used with some success to present these lists [Ref 2]. The radius r of the selected sphere controls the quality and complexity of the corresponding MIMO demodulator.

“E. PIC Demodulation Algorithm”
A sphere decoder can be considered too complex. One way to further reduce the decoder complexity is to generate an APP log independently for each dimension. For iteration I, the decision variables for component x are as follows:

Here, g_k is a column k of the matrix Gl [0]. In this case, assuming that $r^l{}_t$ [n] is a Gaussian variable having an average value x_t, the log “$\xi^l{}_{<t, j>}$ [n], j = 1,..., Q” of the ratio APP is calculated. And the variance is estimated as follows:

The first iterative demodulator can be improved using SIC techniques in the following manner.

Initialization For example, the components are classified with decreasing order as a function of their signal power to interference ratio (SIR).

here,

It is. In recursiveness, the number of iterations t is set from 1 to T (For t = 1 to T). The calculation performs the following:

Assuming that $r^l{}_t$ [n] is a Gaussian variable having an average value x_t, the log “$\xi^l{}_{<t, j>}$ [n], j = 1,. The variance is estimated as follows.

Compute an MMSE estimate of a Gaussian variable $r^l{}_t$ [n] with a known mean value x^t.

"Appendix III"
"List of references"

[1] S. Ariyavisitakul, "Turbo Space-Time Processing to Improve Wireless Channel Capacity," IEEE Trans. Commun., Vol. COM-48, no. 8, pp. 1347-1358, Aug. 2000.
[2] BM Hochwald, S. Ten Brink, "Achieving Near-Capacity on a Multiple-Antenna Channel," IEEE Trans. Commun., Vol. 51, no. 3, pp. 389-399, March 2003.
[3] R. Visoz, AO Berthet, "Iterative Decoding and Channel Estimation for Space-Time BICM over MIMO Block Fading Multipath AWGN Channel," IEEE Trans. Commun., Vol. COM-51, no. 8, pp. 1358- 1367, Aug. 2003.
[4] X. Wang, HV Poor, "Iterative (Turbo) Soft-Interference Cancellation and Decoding for Coded CDMA," IEEE Trans. Commun., Vol. COM-47, no. 7, pp. 1046-1061, July 1999 .

It is a figure which shows the concept of VBLAST. It is a figure which shows the general model of ST-BICM communication. FIG. 2 is a diagram illustrating the structure of an iterative receiver compatible with an embodiment of the present invention. FIG. 2 shows the structure of an iterative receiver that can be used, for example, from the second iteration or from the next iteration.

Explanation of symbols

DESCRIPTION OF SYMBOLS 1 Error correction channel coding 2 Interleaving 3 Spatial demultiplexing 4-1 to 4-M modulation 5-1 to 5-M shaping filter processing 6-1 to 6-M transmitting antenna 8 Initial estimated value 9 Forward filter 10 Subtraction 11 Determining algorithm 12 Recursive backward filter 13 Decoder 14 Space-time deinterleave process 15 Space-time interleave process 16 Weighted data reconstruction process
EV encoder
TX transmit antenna
DE decryption means
RX receiving antenna

Claims (15)

Hide Dependent

1. An iterative decoding and equalization apparatus for high bit rate communication over a frequency selective channel having a plurality of transmit and receive antennas, comprising:
   The apparatus comprises a decision feedback equalizer configured to receive data from different receive antennas,
   The decision feedback equalizer comprises a forward filter (9) and recursively supplied with weighted reconstruction data calculated from the output of the decoder (13) fed with a signal by the decision means (11) And a subtracting means (10) for subtracting the output data of the reverse filter (12) from the output data of the forward filter (9).
   The subtracted data is supplied to the input of the determination means (11) by the output of the decoder (13),
   The decision means (11) taking into account the spatial noise correlation at the output of the subtraction means (10) generates statistics that are sent to a channel decoder having weighted inputs and outputs;
   The determination means (11) and the decoder (13) are separated by a spatio-temporal interleaving process executed in a binary stage,
   Iterative decoding, characterized in that the forward filter (9) and the backward filter (12) are iteratively adapted to minimize the mean square error in the output of the subtracting means (10) And equalization equipment.
2. 2. The determining means (11) at the output of the subtracting means (10) of the equalizer is a spatial whitening type means, followed by a sphere decoder. Equipment.
   3. The determination means (11) at the output of the subtraction means (10) of the equalizer is configured to cancel the remaining spatial interference at the output of the subtraction means (10) of the equalizer. Device according to claim 1, characterized in that it is a means of parallel type (SIC / PIC).
   4. 4. A device according to claim 1, wherein spatial whitening is achieved at the output of the subtracting means (10) of the equalizer.

5. Device according to claim 4, characterized in that spatial whitening is achieved by the determination means (11).
6. Device according to claim 2, characterized in that spatial whitening is achieved by the forward filter (9) and the backward filter (12).
7. 7. A device according to any one of the preceding claims, characterized in that the forward filter (9) becomes an adaptive filter, starting from a certain iteration.
8. A system for high bit rate communication over a frequency selective channel having a plurality of transmit and receive antennas, comprising:
   A system comprising a receiver having the decoding and equalizing apparatus according to any one of claims 1 to 7.
9. 9. The system according to claim 8, comprising ST-BICM type transmission means.
10. Iterative decoding for high bit rate communication over frequency selective channels with multiple transmit and receive antennas using decision feedback equalizers configured to receive data from different receive antennas An equalization method,
    The decision feedback equalizer comprises a forward filter (9) and recursively supplied with weighted reconstruction data calculated from the output of the decoder (13) fed with a signal by the decision means (11) Having a reverse filter (12), and using a subtracting means (10) for subtracting the output data of the reverse filter (12) from the output data of the forward filter (9);
    The subtracted data is supplied to the input of the determination means (11) by the output of the decoder (13),
    The decision means (11) taking into account the spatial noise correlation at the output of the subtraction means (10) generates statistics that are sent to a channel decoder having weighted inputs and outputs;
    The determination means (11) and the decoder (13) are separated by a spatio-temporal interleaving process executed in a binary stage,
    Iterative decoding, characterized in that the forward filter (9) and the backward filter (12) are iteratively adapted to minimize the mean square error in the output of the subtracting means (10) And equalization method.

11. The method of claim 10, wherein recursive filtering of decision feedback equalization is similarly iteratively adapted to minimize mean square error at the input of the equalizer. .
12. 11. A method according to claim 9 or claim 10, wherein the subtracted data is subjected to spatial whitening.
13. 12. Method according to one of claims 9 to 11, characterized in that the decision algorithm (11) is a sphere decoder type algorithm.
14. 13. A method according to claim 11, wherein spatial whitening is achieved by means of a decision algorithm (11).
15. 15. A method according to any one of claims 10 to 14, characterized in that the forward filter (9) is an adaptive filter, starting with an iteration.

Patent Citations (5)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
US5414733A *1993-12-201995-05-09AdtranDecision feedback equalizer employing fixed ratio postcursor taps for minimizing noise and intersymbol interference in signals conveyed over high speed data service loop
EP1069722A2 *1999-07-122001-01-17Hughes Electronics CorporationWireless communication system and method having a space-time architecture, and receiver for multi-user detection
US7197084B2 *2002-03-272007-03-27Qualcomm IncorporatedPrecoding for a multipath channel in a MIMO system
US7292647B1 *2002-04-222007-11-06Regents Of The University Of MinnesotaWireless communication system having linear encoder
US7548598B2 *2003-04-072009-06-16Harris CorporationMethod and apparatus for iteratively improving the performance of coded and interleaved communication systems
* Cited by examiner, † Cited by third party

Cited By (20)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
US8611311B22001-06-062013-12-17Qualcomm IncorporatedMethod and apparatus for canceling pilot interference in a wireless communication system
US7782984B2 *2002-08-302010-08-24Alcatel-Lucent Usa Inc.Method of sphere decoding with low complexity and good statistical output
US8442441B22004-12-232013-05-14Qualcomm IncorporatedTraffic interference cancellation
US8422955B22004-12-232013-04-16Qualcomm IncorporatedChannel estimation for interference cancellation
US8406695B22004-12-232013-03-26Qualcomm IncorporatedJoint interference cancellation of pilot, overhead and traffic channels
US8472877B22005-10-242013-06-25Qualcomm IncorporatedIterative interference cancellation system and method
US8385388B2 *2005-12-062013-02-26Qualcomm IncorporatedMethod and system for signal reconstruction from spatially and temporally correlated received samples
FI20055715A0 *2005-12-302005-12-30Nokia CorpTurboekvaliseringsförfarande
US7974334B2 *2007-01-302011-07-05Texas Instruments IncorporatedSystems and methods for hybrid-MIMO equalization
FR2913161B1 *2007-02-272009-05-22Commissariat Energie AtomiqueMaximum reliability decoder for multi-source system with pulse and amplitude position modulation
US8234538B2 *2007-04-262012-07-31Nec Laboratories America, Inc.Ultra high-speed optical transmission based on LDPC-coded modulation and coherent detection for all-optical network
US8359522B22007-05-012013-01-22Texas A&M University SystemLow density parity check decoder for regular LDPC codes
US7911364B1 *2007-09-042011-03-22Marvell International Ltd.Interleaver for turbo equalization
US8385439B2 *2008-05-272013-02-26Nec Laboratories America, Inc.Polarization mode dispersion compensation in multilevel coded-modulation schemes using blast algorithm and iterative polarization cancellation

CN102340465B *2010-07-162014-11-26大唐移动通信设备有限公司
Frequency domain equalization method and equalization apparatus thereof
US8964901B2 *2011-01-072015-02-24Massachusetts Institute Of TechnologyAnalog/digital co-design methodology to achieve high linearity and low power dissipation in a radio frequency (RF) receiver
US8478335B22011-03-232013-07-02United States Of America As Represented By The Secretary Of The NavySystem and method for radio communication
US8566684B1 *2011-05-262013-10-22Sandia CorporationDecoding and optimized implementation of SECDED codes over GF(q)
US9337868B2 *2013-07-222016-05-10Nec CorporationIterative decoding for cascaded LDPC and TCM coding
FR3014621B12013-12-112016-01-01Thales SaIterative method of synthesis of digital filters for forming a signal
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
Li et al.2011Adaptive decision-feedback detection with constellation constraints for MIMO systems
Li et al.2011Multiple feedback successive interference cancellation detection for multiuser MIMO systems
Andrews2005Interference cancellation for cellular systems: a contemporary overview
US8761323B22014-06-24Impairment covariance and combining weight updates during iterative turbo interference cancellation reception
Kusume et al.2011IDMA vs. CDMA: Analysis and comparison of two multiple access schemes
JP5529810B22014-06-25Repetitive MIMO receiver using group-wise demapping
Baro et al.2000Improving BLAST performance using space-time block codes and turbo decoding
JP4609625B22011-01-12Interference cancellation method and apparatus
El Gamal et al.2000Iterative multiuser detection for coded CDMA signals in AWGN and fading channels
US7349496B22008-03-25Fast space-time decoding using soft demapping with table look-up
US7724832B22010-05-25MIMO decoding

EP1521387B12009-12-02Multi-turbo multi-user detector
Ariyavisitakul2000Turbo space-time processing to improve wireless channel capacity
El Gamal et al.2001A new approach to layered space-time coding and signal processing
US7903728B22011-03-08Equalize training method using re-encoded bits and known training sequences
JP3714910B22005-11-09Turbo receiving method and receiver thereof
JP4072539B22008-04-09Apparatus and method for receiving a signal in a communication system using a multiple-input multiple-output scheme
KR100778647B12007-11-22Combined channel coding and space-block coding in a multi-antenna arrangement
JP2014239445A2014-12-18Multiple antenna transmission with variable diversity gain
EP1404047B12013-08-28Iterative equalisation for MIMO transmission
US7333540B22008-02-19Equalisation apparatus and methods
EP1378089B12012-06-13Joint turbo decoding and equalization for MIMO transmission with intersymbol interference
US7672401B22010-03-02System and method for data communication over multi-input, multi-output channels
EP1540823B12008-01-16Power and confidence ordered low complexity soft turbo-mud with voting system
Witzke et al.2002Iterative detection of MIMO signals with linear detectors

Priority And Related Applications

Priority Applications (2)

ApplicationPriority dateFiling dateTitle
FR0310290A2003-08-292003-08-29Method and apparatus for iterative equalization and decoding for high-speed communications on multi-antenna channels in transmission and reception
PCT/FR2004/0021042003-08-292004-08-06Iterative decoding and equalizing method for high speed communications on multiple antenna channels during transmission and reception

Legal Events

DateCodeTitleDescription
2007-07-13A621Written request for application examination
Free format text: JAPANESE INTERMEDIATE CODE: A621

Effective date: 20070712

2010-02-02A977Report on retrieval
Free format text: JAPANESE INTERMEDIATE CODE: A971007

Effective date: 20100202

2010-06-04TRDDDecision of grant or rejection written
2010-06-16A01Written decision to grant a patent or to grant a registration (utility model)
Free format text: JAPANESE INTERMEDIATE CODE: A01

Effective date: 20100615

2010-06-17A01Written decision to grant a patent or to grant a registration (utility model)
Free format text: JAPANESE INTERMEDIATE CODE: A01

2010-07-22A61First payment of annual fees (during grant procedure)
Free format text: JAPANESE INTERMEDIATE CODE: A61

Effective date: 20100714

2010-07-23FPAYRenewal fee payment (event date is renewal date of database)
Free format text: PAYMENT UNTIL: 20130723

Year of fee payment: 3

2010-07-23R150Certificate of patent or registration of utility model
Free format text: JAPANESE INTERMEDIATE CODE: R150

2013-07-09R250Receipt of annual fees
Free format text: JAPANESE INTERMEDIATE CODE: R250

2014-07-08R250Receipt of annual fees

Free format text: JAPANESE INTERMEDIATE CODE: R250

2015-07-07R250Receipt of annual fees
Free format text: JAPANESE INTERMEDIATE CODE: R250

2016-07-12R250Receipt of annual fees
Free format text: JAPANESE INTERMEDIATE CODE: R250

2017-07-11R250Receipt of annual fees
Free format text: JAPANESE INTERMEDIATE CODE: R250

2019-07-09R250Receipt of annual fees
Free format text: JAPANESE INTERMEDIATE CODE: R250

(19)日本国特許庁(JP)　(12)特　許　公　報(B2)　(11)特許番号
特許第4554610号
(P4554610)
(45)発行日　平成22年9月29日(2010.9.29)　(24)登録日　平成22年7月23日(2010.7.23)

(51)Int.Cl.　F I
H04J 99/00 (2009.01)　H04J 15/00

請求項の数 15　(全 23 頁)

(21)出願番号　特願2006-524383(P2006-524383)
(86)(22)出願日　平成16年8月6日(2004.8.6)
(65)公表番号　特表2007-504693(P2007-504693A)
(43)公表日　平成19年3月1日(2007.3.1)
(86)国際出願番号　PCT/FR2004/002104
(87)国際公開番号　WO2005/025118
(87)国際公開日　平成17年3月17日(2005.3.17)
審査請求日　平成19年7月12日(2007.7.12)
(31)優先権主張番号　0310290
(32)優先日　平成15年8月29日(2003.8.29)
(33)優先権主張国　フランス(FR)

(73)特許権者　591034154
フランス・テレコム
フランス・７５０１５・パリ・プラス・ダルレ・６
(74)代理人　100064908
弁理士　志賀　正武
(74)代理人　100089037
弁理士　渡邊　隆
(74)代理人　100108453
弁理士　村山　靖彦
(74)代理人　100110364
弁理士　実広　信哉

最終頁に続く

(54)【発明の名称】送受信間の複数のアンテナ通信路による高速通信のための反復復号化と等化の方法

(57)【特許請求の範囲】
【請求項１】
複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレート通信のための反復復号化及び等化装置であって、
前記装置は、異なる受信アンテナからデータを受信するように構成される判定帰還型等化器を備え、
前記判定帰還型等化器が、判定手段（１１）によって信号を供給された復号器（１３）の出力から計算されて重み付けされた再構成データを供給される順方向フィルタ（９）及び再帰的逆方向フィルタ（１２）を有すると共に、更に前記順方向フィルタ（９）の出力データから前記逆方向フィルタ（１２）の出力データを減算するための減算手段（１０）を有し、
減算されたデータが、前記復号器（１３）の出力によって前記判定手段（１１）の入力に供給され、
前記減算手段（１０）の出力における空間時間相関を考慮する前記判定手段（１１）が、重み付けされた入力及び出力を有する通信路復号器に送られる統計値を生成し、
前記判定手段（１１）と前記復号器（１３）とが、２連続の段階で実行される時空間インタリーブ処理によって分離され、
前記順方向フィルタ（９）及び前記逆方向フィルタ（１２）が、前記減算手段（１０）の出力における平均二乗誤差を最小限にするように反復して適合される
ことを特徴とする反復復号化及び等化装置。

【請求項２】
前記等化器の前記減算手段（１０）の出力における前記判定手段（１１）が，空間白色化タイプの手段であり、その後にスフィア復号器が接続される
ことを特徴とする請求項１に記載の装置。
【請求項３】
前記等化器の前記減算手段（１０）の出力における前記判定手段（１１）が、前記等化器の前記減算手段（１０）の出力における残りの空間干渉をキャンセルするように構成される，直列及び／または並列タイプ（ＳＩＣ／ＰＩＣ）の手段である
ことを特徴とする請求項１に記載の装置。
【請求項４】
空間白色化が、前記等化器の前記減算手段（１０）の出力において達成される
ことを特徴とする請求項１から請求項３のいずれか１項に記載の装置。
【請求項５】
空間白色化が、前記判定手段（１１）によって達成される
ことを特徴とする請求項４に記載の装置。
【請求項６】
空間白色化が、前記順方向フィルタ（９）及び前記逆方向フィルタ（１２）によって達成される
ことを特徴とする請求項２に記載の装置。
【請求項７】
ある反復を発端として、前記順方向フィルタ（９）が適応フィルタになる
ことを特徴とする請求項１から請求項６のいずれか１項に記載の装置。
【請求項８】
複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレート通信のためのシステムであって、
請求項１から請求項７のいずれか１項に記載された復号化及び等化装置を有する受信機を備える
ことを特徴とするシステム。
【請求項９】
ＳＴ－ＢＩＣＭタイプの送信手段を備える
ことを特徴とする請求項８に記載のシステム。
【請求項１０】
異なる受信アンテナからデータを受信するように構成される判定帰還型等化器を使用した，複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレート通信のための反復復号化及び等化方法であって，
前記判定帰還型等化器が，判定手段（１１）によって信号を供給された復号器（１３）の出力から計算されて重み付けされた再構成データを供給される順方向フィルタ（９）及び再帰的逆方向フィルタ（１２）を有し、前記順方向フィルタ（９）の出力データから前記逆方向フィルタ（１２）の出力データを減算するための減算手段（１０）を使用すると共に、
減算されたデータが、前記復号器（１３）の出力によって前記判定手段（１１）の入力に供給され、
前記減算手段（１０）の出力における空間雑音相関を考慮する前記判定手段（１１）が，重み付けされた入力及び出力を有する通信路復号器に送られる統計値を生成し、
前記判定手段（１１）と前記復号器（１３）とが、２進数の段階で実行される時空間インタリーブ処理によって分離され、
前記順方向フィルタ（９）及び前記逆方向フィルタ（１２）が，前記減算手段（１０）の出力における平均二乗誤差を最小限にするように反復して適合される
ことを特徴とする反復復号化及び等化方法。
【請求項１１】

判定帰還型等化の再帰的フィルタ処理が、同様に、前記等化器の入力で平均二乗誤差を最小限にするように反復して適合される
ことを特徴とする請求項１０に記載の方法。
【請求項１２】
減算されたデータが、空間白色化を受ける
ことを特徴とする請求項９、または請求項１０に記載の方法。
【請求項１３】
判定アルゴリズム（１１）が、スフィア復号器タイプのアルゴリズムである
ことを特徴とする請求項９から請求項１１のいずれか１項に記載の方法。
【請求項１４】
空間白色化が、判定アルゴリズム（１１）によって達成される
ことを特徴とする請求項１１と請求項１２とを結合した請求項に記載の方法。
【請求項１５】
ある反復を発端として、前記順方向フィルタ（９）が適応フィルタになる
ことを特徴とする請求項１０から請求項１４のいずれか１項に記載の方法。
【発明の詳細な説明】
【技術分野】
【０００１】
本発明は、デジタル通信の分野に関係するものである。本発明は、更に特に、複数の送信アンテナ通信路、更に特に複数の送信アンテナ及び受信アンテナを有する周波数選択性通信路を介した高速通信のための反復復号化と等化の方法に関係するものである。
【背景技術】
【０００２】
アンテナ技術の開発に関して、ＴＤＭＡ、ＣＤＭＡ、ＯＦＤＭ、及びそれらの組み合わせに基づく通信モデルは、ＭＩＭＯ通信路を達成するために、系統的に修正されると共に、拡張される。異なる哲学を有する、以下のような２つの主要な時空間符号化の種類が想定される。
【０００３】
ベル研究所によって開発されたＢＬＡＳＴ（Bell laboratory LAyered Space-Time architecture）伝送は、通信路を介して送信され得るビットレートを増加するために、空間多重化を使用することを目的とする。ＢＬＡＳＴを、２つの異なる階層化時空間符号化モデルを与えるエラー訂正チャンネル符号化と結合する２つの方法がある。第１の（ＳＴ－ＢＩＣＭ）モデルにおいて、様々な階層を連結する１つの符号化動作は、時空間インタリーブ、及び空間逆多重化［参考文献１，ＬＳＴ－１，ｐ．１３９０］の前のデータに適用される。第２の（ＭＡＣ）モデルにおいて、データは、空間逆多重化されると共に、その後で、各階層［参考文献１、ＬＳＴ－１１、ｐ．１３５０］上で、独立して符号化されて、インタリーブされる。
【０００４】
ダイバシティ方式には、更に効果的に通信路のフェーディングの影響と戦うと共に、性能を向上させるために、空間ダイバーシティを使用する。符号化を改良するＳＴＴＣ方式、及び符号化利得が５にならず、多くの場合当該技術においてガウス通信路として知られている最適トレリス符号化変調と組み合わせられるＳＴＢＣ方式（及び、それらの一般化であるＯＤ，ＬＤＣ等）がある。
【０００５】
最近の理論上の貢献は、空間多重化及び空間ダイバーシティにおける改良が、妥協の関係によって連結されることを証明した。例えば、２つの上記アプローチは、送信アンテナと受信アンテナの任意に高い数Ｍ，Ｎに関して、同じ割合で容量を増大させ得ない。送信ダイバシティ方式は、情報理論の見地から、多数のアンテナを使用するＭＩＭＯ通信路を介した高ビットレートの通信に対するそれらの妥当性に著しく疑問があるＭＩＳＯ構成［Ｍ，Ｎ］＝［２，ｎ］においてのみ最適化されることが示され得る。

【０００６】
ＳＴ－ＢＩＣＭアプローチのみが、ｍｉｎ｛Ｍ、Ｎ｝によって、容量の線形の増大を保証する。
【０００７】
こういうわけで、この種類の伝送を選択すると共に、本質的に得ることができる限り近い性能を保証する適当な受信機に設計努力を集中させることが必要である。ＳＴＢＣと異なり、ＳＴ－ＢＩＣＭは、様々なアンテナによって送信されたデータストリームの相互の直交性を保証しない。
【０００８】
周波数選択性ＭＩＭＯ通信路上の伝送の場合、この符号化戦略は、空間領域の多元接続干渉（ＭＡＩ：multiple access interference）、及び時間領域の符号間干渉（ＩＳＩ：intersymbol interference）を消去するために、従って受信機におけるデータの共同の復調を必要とする。
【０００９】
共同のデータ復調機能は、特に、ＳＴ－ＢＩＣＭが、空間領域におけるダイバシティの全てを獲得するために、通信路復調化と復号化との間の反復処理を本来必要とするので、受信機の精密かつ複雑な要素を構成する。
【００１０】
「周波数非選択性ＭＩＭＯ通信路上の反復復号化」
周波数非選択性ＭＩＭＯ通信路上のＳＴ－ＢＩＣＭの反復復号化は、完璧なＩＳＩキャンセルの２つの面を有する仮定の下で解決されるべき問題の特別な場合を構成すると共に、各シンボルに関して利用可能なエネルギーの全てを回復する。
【００１１】
その場合に、問題は、以下の“空間的色付き雑音のある所での一様なＭＩＭＯ通信路上のＭＡＩのキャンセル”という用語に言い換えられ得る。それは、送信されたベクトルのシンボルのビットに関するＡＰＰの計算を常に必要とする（復号器に送信されるべきデータに関する統計）。
【００１２】
２つの方法が、文献で説明される。
【００１３】
－完全なリスト（ＭＡＰ）に基づくＡＰＰの正確な計算。
【００１４】
この方法の複雑さが、送信アンテナの数Ｍに従い指数的に増加すると共に、使用される信号点配置の次数Ｑの多項式関数であるので、この方法の複雑さは大変に高額である。
【００１５】
－スフィア復号（球内復号）化アルゴリズム［参考文献２］によって生成された候補ベクトルの削減リストからのＡＰＰの近似計算。
【００１６】
スフィア復号化アルゴリズムには、少なくとも、それはＭＡＰ基準を保存すると共に、使用される信号点配置の次数Ｑに影響を受けないＯ（Ｍ３）における複雑さを有するという、２つの利点がある。スフィア復号化アルゴリズムがファノ逐次復号化アルゴリズムの変形と見なされ得る点に注意が必要である。
【００１７】
－逐次スタックアルゴリズムによって生成された候補ベクトルの最終候補リストからのＡＰＰの近似計算。
【００１８】
スフィア復号化アルゴリズムは、周波数非選択性ＭＩＭＯ通信路に関して集中的に研究された。周波数選択性ＭＩＭＯ通信路の状況に一般化することが重要である。
【００１９】
負担の重いアプローチは、ブロック反復であろう。このアプローチは、それが検索スペ

ースのサイズを人工的に増大させる(従って複雑さを増大させる)と共に、シンボルの冗長な復号化を導入するので、関心が薄いアプローチである。

【００２０】

［参考文献２］で説明されたアルゴリズムは、周波数非選択性ＭＩＭＯ通信路に関して導き出されている。それを周波数選択性ＭＩＭＯ通信路の状況に一般化することに関して、まだ結論は出ていない。

【００２１】

これらの２つのアルゴリズムの性能は、好ましくは調整サイズである生成された候補のリストの品質によって本質的に決定される。あらゆる点で、１ビットの最適復号化、及び０ビットの最適復号化を含むことが、それにとって不可欠である。

【００２２】

『周波数選択性ＭＩＭＯ通信路上の反復復号化』

反復構造に挿入されるいくつかの種類の復調器が提案された。

【００２３】

－ＢＣＪＲアルゴリズムに基づいてＭＡＰ基準を適用する、重み付けされた入力及び出力を有する復調器。

【００２４】

Ｏ（ＱＬｘＭ）におけるこの受信機の複雑さは、高次の変調、及び多数の入力と高いメモリ需要を有するＭＩＭＯ通信路に関して、明らかに高額である。

【００２５】

－状態数［参考文献３］に関して非常に削減されたトレリスにおける効率的な検索アルゴリズムに基づくと共に、必要に応じて最小位相フィルタが前に配置される、次善の重み付けされた入力及び出力の復調器。

【００２６】

このアプローチは、Ｏ（ＱＬｒｘＭ）におけるその複雑さによってまだ制限される。ここで、Ｌｒは削減された拘束長を表す。特に、多数の状態を有する信号点配置の使用が除外される。

【００２７】

干渉の反復線形キャンセル（［参考文献４］のマルチユーザ復調の環境参照）から構成された非線形復調器を、ＭＩＭＯ環境に置き換えることは、一般的に異なる利用者とアンテナとの間の共通点に基づいている。それらの実施は、従ってアンテナ毎に１つのフィルタを必要とすると共に、それらの性能は、その複雑さが送信アンテナ数の多項式関数であるという利点を有する、アンテナアプローチ毎の不連続ＭＭＳＥによって制限される。

【非特許文献１】S. Ariyavisitakul, "Turbo Space-Time Processing to Improve Wireless Channel Capacity," IEEE Trans. Commun., vol. COM-48, no. 8, pp. 1347-1358, Aug. 2000.

【非特許文献２】B. M. Hochwald, S. Ten Brink, "Achieving Near-Capacity on a Multiple-Antenna Channel," IEEE Trans. Commun., vol. 51, no. 3, pp. 389-399, March 2003.

【非特許文献３】R. Visoz, A.O. Berthet, "Iterative Decoding and Channel Estimation for Space-Time BICM over MIMO Block Fading Multipath AWGN Channel," IEEE Trans. Commun., vol. COM-51, no. 8, pp. 1358-1367, Aug. 2003.

【非特許文献４】X. Wang, H.V. Poor, "Iterative (Turbo) Soft-Interference Cancellation and Decoding for Coded CDMA," IEEE Trans. Commun., vol. COM-47, no. 7, pp. 1046-1061, July 1999.

【発明の開示】

【発明が解決しようとする課題】

【００２８】

本発明の目的は、複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレートデジタル伝送のための、あまり複雑ではない高度な受信機を提案すると共に

，特に、空間領域におけるアンテナ間の干渉、及び時間領域におけるシンボル間の干渉の処理を提供する一方で，大きな計算パワーを特に必要としない受信機を提案することである。

【課題を解決するための手段】

【００２９】

この目的を達するために、本発明は、複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレート通信のための反復復号化及び等化装置を提案する。前記装置は、異なる受信アンテナからデータを受信するように構成される判定帰還型等化器を備え、判定帰還型等化器は、判定手段によって信号を供給された復号器の出力から計算されて重み付けされた再構成データを供給される順方向フィルタ及び再帰的逆方向フィルタを有すると共に、更に前記順方向フィルタの出力データから前記逆方向フィルタの出力データを減算するための手段を有する。ここで、減算されたデータは、復号器の出力によって判定手段の入力に供給され、減算手段の出力における空間干渉相関を考慮する判定手段は、重み付けされた入力及び出力を有する通信路復号器に送られる統計値を生成し、判定手段と復号器とが、２進数の段階で実行される時空間インタリーブ処理によって分離される。前記装置は、順方向フィルタ及び逆方向フィルタが、減算手段の出力で平均二乗誤差を最小限にするように反復して適合されるという点で特徴付けられる。

【００３０】

前述の種類の装置には、有利に，個々もしくは技術的に可能な全ての組み合わせにおける以下の補足的特徴がある。

－等化器の減算手段の出力における判定手段は、空間白色化タイプの手段であり，その後にスフィア復号器が接続される。

－等化器の減算手段の出力における判定手段は、等化器の減算手段の出力における残りの空間干渉をキャンセルするように構成される、直列及び／または並列タイプ（ＳＩＣ／ＰＩＣ）の手段である。

－空間白色化は，等化器の減算手段の出力で達成される。

－空間白色化は，判定手段によって達成される。

－空間白色化は、順方向フィルタ及び逆方向フィルタによって達成される。

－ある反復を発端として、順方向フィルタが適応フィルタになり、逆方向フィルタが中央係数がない同様の適応フィルタになる。

【００３１】

本発明は、アンテナを異なる利用者と考えるアプローチとは根本的に異なる、新奇なアプローチから生じる。送信された信号は、Ｔ－次元の変調と見なされる。ここで、Ｔは送信アンテナの数である。この変更された見地は、その入力においてＴ－次元のベクトル変調が周波数選択性通信路と巻き込まれているとみなされると共に、通信路復号器（チャンネルデコーダ）のための適応性のある情報を生成することが可能であるＴ－次元の変調の復調器（例えば、スフィア復号器）がそのあとに接続される線形反復等化器によって構成される受信機の設計に対してかなりの影響を与える。このアプローチには，それが１つの（ベクトルの）フィルタを必要とするので、それがあまり複雑ではないと共に、それがベクトル復調基準の選択を可能にするので更に良い性能を提供するという点で、従来技術に対する２つの利点を有する。

【００３２】

本発明は，同様に、前述の種類の等化及び復号化装置を含む受信機を備えるという点で特徴付けられた、複数の送信及び受信アンテナを有する周波数選択性通信路を介した高ビットレート通信のためのシステムに関係する。

【００３３】

送信手段は、例えば有利なＳＴ－ＢＩＣＭタイプの手段である。

【００３４】

本発明は，更に、異なる受信アンテナからデータを受信するように構成される判定帰還型等化器を使用した、複数の送信及び受信アンテナを有する周波数選択性通信路を介した

高ビットレート通信のための反復復号化及び等化方法を提案する。判定帰還型等化器は，判定手段によって信号を供給された復号器の出力から計算されて重み付けされた再構成データを供給される順方向フィルタ及び再帰的逆方向フィルタを有し、前記順方向フィルタの出力データから前記逆方向フィルタの出力データを減算するための手段を使用する。ここで、減算されたデータは、復号器の出力によって判定手段の入力に供給され、減算手段の出力における空間雑音相関を考慮する判定手段は、重み付けされた入力及び出力を有する通信路復号器に送られる統計値を生成し、判定手段と復号器とが、２進数の段階で実行される時空間インタリーブ処理によって分離される。前記方法は、順方向フィルタ及び逆方向フィルタが、減算手段の出力で平均二乗誤差を最小限にするように反復して適合されるという点で特徴付けられる。

【発明を実施するための最良の形態】

【００３５】

本発明の他の特徴、及び利点は、単に実例となり、そして発明に関して制限しないと共に、添付された図面と一緒に読まれるべきである以下の記述から明らかになる。

【００３６】

「ＢＬＡＳＴ技術，及びＳＴ－ＢＩＣＭ符号化」

図１は、ＢＬＡＳＴ構造の一般的な構成図である。

【００３７】

時空間インタリーブ、及び空間逆多重化の後で、１つのベクトルの符号化装置ENCで符号化されたデータは、複数の送信アンテナTXによって送信される。ＭＩＭＯ通信路の反対側において、送信された信号は、その出力においてはデータが復元される復号化手段DE（ＭＩＭＯ通信路）にそれらを送る、複数の受信アンテナRXによって受信される。

【００３８】

以下の記述は、周波数選択性ＭＩＭＯ通信路に関係する。符号間干渉の存在は、受信機の複雑さを増大させる。

【００３９】

図２は、ＳＴ－ＢＩＣＭの一般的な通信モデルを示す，

【００４０】

伝送において、データは、エラー訂正チャンネル符号化（１）（畳み込み符号、ターボ符号、ＬＤＰＣ符号等）され、２進数の段階でインタリーブ（２）され、空間逆多重化（３）され、そして，各端部に関して変調（４－１から４－Ｍまで）される。成形フィルタ処理（５－１から５－Ｍまで）したあと、変調されたデータは、送信アンテナ（６－１から６－Ｍまで）に送られる。

【００４１】

「送信処理及び受信処理の記述」

伝送において、データは，ＳＴ－ＢＩＣＭ処理を受ける。

【００４２】

これは、以下のステップを包含する。

－所定のビットレートでデジタルデータストリームを受信するステップ。

－符号化されたデジタルデータストリームを生成するためにエラー訂正チャンネル符号化（１）を適用するステップ。

－ビットの段階のインタリーバを用いて、符号化されたデジタルデータをインタリーブ（２）するステップ。

－送信機階層と呼ばれる複数の個別の符号化されたデジタルデータストリーム（Ｍ個の区別可能なストリームまたは送信機階層）を作成するために、符号化されインタリーブされたデジタルデータストリームを空間逆多重化（３）するステップ。

【００４３】

そして、Ｍ個の区別可能な符号化されたデジタルデータストリームは、以下のように処理される。

－変調されたシンボルのストリームを獲得するための変調方式に従って，符号化されイン

タリーブされたデジタルデータストリーム１５を変調（４－１から４－Ｍまで）する。
－変調されたストリームをフィルタ処理（５－１から５－Ｍまで）する。
－それ自身の送信アンテナ経由で、変調されたストリームを送信（６－１から６－Ｍまで）する。

【００４４】
図３は、伝送システムの受信機を示す。

【００４５】
それは，順方向フィルタ９、減算器１０、判定アルゴリズム１１、及び再帰的逆方向フィルタ１２によって定義される判定帰還等化器を備える。

【００４６】
再帰的逆方向フィルタ１２は、決定アルゴリズム１１によって信号が供給される復号器１３の出力から，重み付けされ再構成された計算データを供給される。

【００４７】
順方向フィルタ９及び逆方向フィルタ１２は、ＭＩＭＯ等化器の平均二乗誤差（ＭＳＥ）を最小限にするために、すなわち減算器１０の出力におけるエラーを最小限にするために、反復して決定される。

【００４８】
この目的を達するために、それらは，ＭＩＭＯ通信路の初期推定値８を使用する。

【００４９】
例えば、使用される処理は、別表IIで示された種類の処理である。

【００５０】
前述の処理に照らすと、残余誤差，及び雑音のベクトルの推定値が使用されると共に、順方向フィルタ及び逆方向フィルタは、このベクトルを最小限にするために、ブロック単位で計算される点に注意が必要である。

【００５１】
このベクトルの処理は，計算を単純化する。

【００５２】
判定手段１１の入力は、減算器１０の出力から、及び復号器１３の出力から、データを受け取る。

【００５３】
そのアルゴリズムは、異なる種類のアルゴリズム、特にＳＩＣ／ＰＩＣタイプ（等化器の減算器手段の出力における残りの空間干渉の直列及び／または並列のキャンセル－別表III参照。）のアルゴリズムであり得る。

【００５４】
代りに、そのアルゴリズムは，スフィア復号器アルゴリズムであり得る。

【００５５】
この種類のアルゴリズムは、Ｏ（M^3）における複雑さを有している。（ここで、Ｍは，Ｑから独立している送信アンテナの数である。）

【００５６】
これは、ビットレートを増加する目的で多数の状態を有する変調の使用を可能にする。

【００５７】
判定手段１１の出力は、前記判定手段１１と復号器１３との間の２連続の段階で実施される時空間デインターリーブ処理１４に送られる。

【００５８】
復号器１３の出力は，ビットの確率である。

【００５９】
この確率は、最初に判定アルゴリズム１１の入力に送られると共に、次に時空間インターリーブ（１５）を実行したあとで、重み付けされたデータの再構成処理１６に送られる。

【００６０】

このように再構成された、重み付けされたデータは、逆方向フィルタ１２の入力に送られる。

【００６１】

逆方向フィルタ１２の出力は、重み付けされた再生データを構成する。

【００６２】

更に、減算器１０の入力において注入される残りの干渉及び雑音と対応するエラーは、時間領域及び空間領域の両方で色付きである。

【００６３】

時間相関は等化器によって実行された処理に対して影響をほとんど与えないが、空間相関は本質的役割を有している。

【００６４】

こういうわけで、一実施例において，空間白色化がCholesky因数分解を用いて達成される。

【００６５】

この空間白色化は、減算器１０の出力で有利に達成される。

【００６６】

一つの特別な実施例において、それは、スフィア復号化アルゴリズム１１そのものによって達成されることができる。

【００６７】

それは、順方向フィルタ９及び逆方向フィルタ１２に、等しく統合されることができる。

【００６８】

最終的に、一実施例において，反復のある所定の回数から、例えば，第２番目の反復から、順方向フィルタ９は、適応フィルタと有利に交換される。

【００６９】

逆方向フィルタが直接順方向フィルタから推定されるので，適応フィルタの中央係数がない通信路との畳み込みが従ってこの場合に生じる。

【００７０】

これは，図４において示されており、順方向フィルタ９は、適応フィルタＭＦとなると共に、一方逆方向フィルタ１２$B_{M,p}$は、中央係数のない通信路と畳み込まれた適応フィルタとなる。

【００７１】

【表１】

『別表1』
「省略形のリスト」

ＡＰＰ	：経験的確率（A Posteriori Probability）
ＢＣＪＲ	：Bahl, Cocke, Jelinek, Raviv（アルゴリズム） （Bahl, Cocke, Jelinek, Raviv (algorithm)）
ＢＬＡＳＴ	：Bell Labs Layered Space Time
ＣＤＭＡ	：符合分割多元接続（Code Division Multiple Access）
ＩＣ	：干渉キャンセル（Interference Cancellation）
ＩＳＩ	：符号間干渉（InterSymbol Interference）
ＬＤＰＣ	：低密度パリティ検査（Low Density Parity Check）
ＭＡＩ	：多元接続干渉（Multiple Access Interference）
ＭＡＰ	：最大事後確率（Maximum A Posteriori (Probability)）
ＭＦ	：整合フィルタ（マッチドフィルタ）（Matched Filter）
ＭＩＭＯ	：複数入力複数出力（Multiple Input Multiple Output）
ＭＭＳＥ	：最小平均二乗誤差（Minimum Mean Square Error）
ＯＦＤＭ	：直交周波数分割多重方式 （Orthogonal Frequency Division Multiplex）
ＰＩＣ	：並列干渉キャンセル（Parallel Interference Cancellation）
ＳＩＣ	：直列干渉キャンセル（Serial Interference Cancellation）
ＳＩＳＯ	：単一入力単一出力（Single Input Single Output）
Ｓｏｆｔ－ＩＣ	：ソフト干渉キャンセル（Soft Interference Cancellation）
ＳＴＢＣ	：時空間ブロック符号（Space-Time Block Codes）
ＳＴ－ＢＩＣＭ	：時空間ビットインタリーブ符号化変調 （Space-Time Bit-Interleaved Coded Modulation）
ＳＴＴＣ	：時空間トレリス符号（Space-Time Trellis Codes）
ＴＤＭＡ	：時分割多元接続（Time Division Multiple Access）
ＷＭＦ	：白色化整合フィルタ（Whitening Matched Filter）

【００７２】
『別表II』
「通信モデル」
複数の入力及び出力を有するＰ個のブロック通信路、メモリがＭであるＴ個の周波数選択性送信アンテナ及びＲ個の周波数選択性受信アンテナを検討する。
【００７３】
「Ａ．時空間２進インターリーブ符号化変調」
仮に、Ｃが長さN_cの線形符号を示すと共に、データベクトル"$u \in F_2^{P \cdot (S-M)}$"を認めて、符号化ワード"$c \in F_2^S$"を生成する$F_2$に関して$\rho_c$を与えるとする。もし巻き込み符号が使用される場合，与えられたρ_cは、テールビットを含むと仮定される。符号化ワードは、上手く設計された２進インターリーバπに入れられ、その出力行列"$A \in F_2^{[S \times P]L}$"は，Ｐ個の行列"$A^\rho \in F_2^{[S \times P]L}$，$\rho = 1, \cdots, P$"に分割される。行列$A^\rho$の列は、スタックされた２進数成分"$a^\rho_{\langle t,1\rangle}[n], \cdots a^\rho_{\langle t,J\rangle}[n]$"を備えるＴ個のサブベクトル"$a^\rho_t[n] \in F_2^S$，$t = 1, \cdots, T$"（入力通信路毎に１個）を含む、"シンボルラベルベクトル"と呼ばれるベクトル"$a^\rho[n] \in F_2^{TS}$，$n = 1, \cdots, L$"である。ここで、"$\langle t,j\rangle$"は，インデックス"$(t-1)q+j$"を示す。各行列A^ρにおいて、全てのベクトル"$a^\rho[n]$"は、シンボル"$\mu : \{0, 1\}^M \to A$"を有するビット符号化を経由して、基数"$|A| = 2^M$"の信号のセット"$A \subset \mathbb{C}^D$"に関して、無記憶性のＤ一次元変調器よって変調される。一般性の損失なしで、Ａが同じ次数2^MのＴ個の複素信号点配置Ｃの単純な積であると共に，各信号点配置に関してグレイ符号化が使用されると仮定される。信号の変換の後で、この

符号化された変調プロセスは、同様に、全ての符号化ワード複素シンボル"P×L×T"か、P個の行列"Xp ∈ C^{T/L}, p = 1, ..., P"に分割される時空間符号化変調方式と考えられる。行列"Xp"の列"x^p [n] ∈ C^T, n = 1, ..., L"は、"信号点配置シンボルベクトル"と呼ばれる。"シンボルラベルベクトル"a^p [n]は、シンボル"μ^{-1}"によるビット符号化の単純な反転によって"x^p [n]"から獲得され得る。時空間符号の一般的なクラスに分類されて、このアーキテクチャは、時空間ビットインタリーブ符号化変調（ＳＴＢＩＣＭ：space-time bit-interleaved coded modulation）［参考文献３］と呼ばれる。ナイキスト（Nyquist）制限帯域の理想的なフィルタ処理条件から、（通信路毎に使用されるビットにおいて）与えられる送
信は、以下のようになる。

【００７４】
【数１】

$$\rho = qT\rho_c$$

【００７５】
図１は、この通信モデルの構成図を示す。
【００７６】
「Ｂ、ＭＩＭＯ通信路」
仮に、"H^p ∈ C^{R×T×(k+1)}"は、ブロックpに関するＭＩＭＯ通信路であるとする。ここで、"p = 1, ..., P"であると共に、"H = {H^p}"は全ての通信路のセットであるとする。同様に、仮に、"X^p ∈ C^{T×L}"、及び"Y^p ∈ C^{R×L}"は、それぞれ"信号点配置シンボル行列"、及び"通信路出力行列"とする。時刻"n = 1, ..., L"の離散時間"y^p [n] ∈ C^R"における同等のベースバンド内のベクトルの通信路出力は、以下のように書かれることができる。

【００７７】
【数２】

$$\mathbf{y}^p[n] = [\mathbf{H}(z)]\mathbf{x}^p[n] + \mathbf{w}^p[n] = \sum_{k=0}^{M} \mathbf{H}^p[k]\mathbf{x}^p[n-k] + \mathbf{w}^p[n] \qquad (2)$$

【００７８】
前述の方程式において、"x^p [n] ∈ C^T"は、各成分のエネルギーが"σ_x^2"に等しい、時刻nにおいて送信される"信号点配置シンボルベクトル"である。"H^p [k] ∈ C^{R×T}"は、通信路のインパルス応答のタップ数kの行列である。"w^p [n] ∈ C^R"は、付加的な複素ノイズベクトルである。付加的な複素ノイズベクトルw^p [n]は、等及び円対称である複素ガウスタイプの互いに独立で同一の分布に従う平均を有すると仮定され、従って、"pdf ~ (0, σ^2 I)"となる。対応するブロック全体で一定である通信路H^pは、長さ"M+1"の有限インパルス応答（ＦＩＲ）を有しており、その空間シンボルタップ"H^p [0], ..., H^p [M]"は、等平均のランダムな複素行列"R×T"であると共に、等しい電力システムの場合、その平均電力は以下に示す正規化の制限を満たす。

【００７９】
【数３】

$$\mathbf{E}\left[\mathrm{diag}\left\{\sum_{k=0}^{M} \mathbf{H}^p[k]\mathbf{H}^p[k]^{\dagger}\right\}\right] = T\mathbf{I} \qquad (3)$$

【００８０】
ここで、演算子"†"は、共役の転置演算子に対応する。個々の複数の通信路成分の数

が、大きな建造物及び反射している物体によって主として決定されると仮定すると、全ての"R×T"の可能性がある後続に関する同一通信路メモリは妥当な想定である。

【００８１】

「ＭＩＭＯ干渉キャンセルの反復ブロック」

「Ａ、理論、及びスケジューリング」

ＭＩＭＯ干渉キャンセラーの反復ブロックは、出力復号器によりフィードバックされたランダムな確率的情報を別々に使用して、各受信データブロック"Y^p, $p = 1, \ldots, P$"を処理する。説明された全ての信号及び要素は、従ってブロック単位でインデックスを付けられる。この依存状態は、表記法を簡素化するために時々省かれる。各反復Ｉの間、ベクトルy[n]の各受信されたシンボルに適用された線形順方向フィルタF^iは、ベクトル信号$y^i[n]$を生成する。ＭＡＩの、及びx[n]を悪化させるＩＳＩの適切に構成された推定値$e^i[n]$は、$z^i[n]$を生成するために、ベクトル信号$y^i[n]$から減算される。

【００８２】

【数４】

$$\mathbf{z}^i[n] = \tilde{\mathbf{y}}^i[n] - \mathbf{e}^i[n] \qquad (8)$$

【００８３】

反復Ｉ－１上で利用可能な予備的知識（付帯的な確率分布）を与えられて、ベクトル$e^i[n]$の推定値が、ベクトルx[n]に関する適応性のある判定試みベクトル$x^i[n]$によって惹起される逆方向フィルタB^iの出力から提供される。

【００８４】

説明は、次に、ＭＩＭＯターボ等化器の核心、すなわちのＭＩＭＯ順方向有限インパルス応答フィルタ、及びＭＩＭＯ逆方向有限インパルス応答フィルタの導出について考える。

【００８５】

「Ｂ．順方向フィルタ及び逆方向フィルタの計算」

その記述は、順方向フィルタF^iの瞬間的な出力$y^i[n]$のいくつかの基礎的な操作で始まる。

【００８６】

【数５】

$$\tilde{\mathbf{y}}^i[n] = \sum_{i=-L_{F_2}}^{L_{F_1}} \mathbf{F}^i[i]\,\mathbf{y}[n-i] \qquad (22)$$

【００８７】

【数６】

$$\mathbf{y}[n] = \sum_{k=0}^{M} \mathbf{H}[k]\,\mathbf{x}[n-k] + \mathbf{w}[n] \qquad (23)$$

【００８８】

なので、各サンプルの$y^i[n]$は、以下のように拡張されることができる。

【００８９】

【数７】

$$\tilde{\mathbf{y}}^i[n] = \sum_{i=-L_{F_2}}^{L_{F_1}} \sum_{k=0}^{M} \mathbf{F}^i[i]\,\mathbf{H}[k]\,\mathbf{x}[n-i-k] + \sum_{i=-L_{F_2}}^{L_{F_1}} \mathbf{F}^i[i]\,\mathbf{w}[n-i] \qquad (24)$$

【００９０】
２つのフィルタ，フィルタ$\mathbf{H}_c$及びフィルタ$\mathbf{F}^l$の畳み込みに関する同等の行列表記は、以下のようになる。
【００９１】
【数８】

$$\tilde{\mathbf{y}}^l[n] = \mathbf{F}^l\mathbf{H}_c\mathbf{x}_c[n] + \mathbf{F}^l\mathbf{w}_c[n] \qquad (25)$$

【００９２】
ここでは、
【００９３】
【数９】

$$\mathbf{F}^l = \left[\mathbf{F}^l[-L_{F_1}] \cdots \mathbf{F}^l[0] \cdots \mathbf{F}^l[L_{F_2}]\right] \in \mathbb{C}^{T\times RL_F} \qquad (26)$$

【００９４】
は、次数"LF = LF1 + LF2 + 1"の複方向フィルタである。ここで、
【００９５】
【数１０】

$$\mathbf{H}_c = \begin{bmatrix} \mathbf{H}[0] & \mathbf{H}[1] & \cdots & \mathbf{H}[M] & & & \\ & \mathbf{H}[0] & \mathbf{H}[1] & \cdots & \mathbf{H}[M] & & \\ & & \mathbf{H}[0] & \mathbf{H}[1] & & \mathbf{H}[M] & \\ & & & \ddots & & & \ddots \\ & & & & \mathbf{H}[0] & \mathbf{H}[1] & \cdots & \mathbf{H}[M] \end{bmatrix} \in \mathbb{C}^{RL_F\times T(L_F+M)} \qquad (27)$$

【００９６】
は、Toeplitz対角帯通信路行列であると共に、
【００９７】
【数１１】

$$\mathbf{x}_c[n] = \begin{bmatrix} \mathbf{x}[n+L_{F_1}] \\ \vdots \\ \mathbf{x}[n] \\ \vdots \\ \mathbf{x}[n-L_{F_2}-M] \end{bmatrix} \in \mathbb{C}^{T(L_F+M)} \qquad (28)$$

【００９８】
は、送信されたベクトルである。結合されたフィルタを導入すると、以下のようになる。
【００９９】

【数１２】

$$\mathbf{G}^i = \mathbf{F}^i\mathbf{H}_c = \left[\mathbf{G}^i[-L_{G_1}] \cdots \mathbf{G}^i[0] \cdots \mathbf{G}^i[L_{G_2}]\right] \in \mathbb{C}^{T\times TL_W} \qquad (30)$$

【０１００】
そして、以下の最終の表現が獲得される。
【０１０１】
【数１３】

$$\tilde{\mathbf{y}}^i[n] = \mathbf{G}^i\mathbf{x}_c[n] + \mathbf{F}^i\mathbf{w}_c[n] \qquad (31)$$

【０１０２】
ここで、“$L_{G1} = L_{F1}$”、“$L_{G2} = L_{F2} + N$”、そして“$L_W = L_F + N$”である。
【０１０３】
類推することによって、逆方向フィルタは、以下のとおりに定義される。
【０１０４】
【数１４】

$$\mathbf{B}^i = \left[\mathbf{B}^i[-L_{B_1}] \cdots \mathbf{0} \cdots \mathbf{B}^i[L_{B_2}]\right] \in \mathbb{C}^{T\times TL_W} \qquad (32)$$

【０１０５】
スクランブル補正回路の出力は、以下のように書くことができる。
【０１０６】
【数１５】

$$\mathbf{z}^i[n] = \sum_k \mathbf{G}^i[k]\mathbf{x}[n-k] - \sum_k \mathbf{B}^i[k]\tilde{\mathbf{x}}^i[n-k] + \sum_k \mathbf{F}^i[k]\mathbf{w}[n-k] \qquad (33)$$

【０１０７】
雑音及び残りのスクランブルベクトル$\mathbf{v}^i[n]$（ＭＩＭＯ等化器エラーベクトルとも呼ばれる）は、以下のように表される。
【０１０８】
【数１６】

$$\begin{aligned}\mathbf{v}^i[n] &= \mathbf{z}^i[n] - \mathbf{G}^i[0]\mathbf{x}[n] \\ &= \sum_{k\neq 0}\mathbf{G}^i[k]\mathbf{x}[n-k] - \sum_k \mathbf{B}^i[k]\tilde{\mathbf{x}}^i[n-k] + \sum_k \mathbf{F}^i[k]\mathbf{w}[n-k] \\ &= \sum_k^i \mathbf{G}^i[k]\mathbf{x}[n-k] - \sum_k \mathbf{B}^i[k]\tilde{\mathbf{x}}^i[n-k] + \sum_k \mathbf{F}^i[k]\mathbf{w}[n-k] \qquad (34)\end{aligned}$$

【０１０９】
ここでは、
【０１１０】

【数１７】

$$\underline{\mathbf{G}}^{l} = \left[\mathbf{G}^{l}\left[-L_{G_1}\right] \cdots \mathbf{0} \cdots \mathbf{G}^{l}\left[L_{G_2}\right]\right] \qquad (35)$$

【０１１１】
である。フィルタF^{l}、及びフィルタB^{l}は、"G^{l}[0] = I"という制限の下で、ＭＩＭＯ等化器の平均二乗誤差を最小限にするように、ブロック単位で計算され、ＭＩＭＯ状況における問題が、以下のように簡潔に公式化され得る。
【０１１２】
【数１８】

$$\left\{\mathbf{F}^{l}, \mathbf{B}^{l}\right\} = \arg \min_{\{\mathbf{F},\mathbf{B}\}/\mathbf{G}^{l}[0]=\mathbf{I}} \mathrm{tr}\, \mathbf{E}\left\{\mathbf{v}^{l}[n]\,\mathbf{v}^{l}[n-k]^{\dagger}\right\}$$
$$= \arg \min_{\{\mathbf{F},\mathbf{B}\}/\mathbf{G}^{l}[0]=\mathbf{I}} \mathrm{tr}\left\{\mathbf{K}_{\mathbf{v}}^{l}\right\}$$

【０１１３】
ＭＳＥの最小化は、２つの連続するステップにおいて実行され、そして最終的に以下のように与えられる。
【０１１４】
【数１９】

$$\mathbf{B}^{l} = \underline{\mathbf{G}}^{l} \qquad (37)$$

$$\mathbf{F}^{l} = \left(\mathbf{E}_{\Delta}\mathbf{H}_{\mathbf{c}}^{\dagger}\boldsymbol{\Phi}^{l-1}\mathbf{H}_{\mathbf{c}}\mathbf{E}_{\Delta}^{\dagger}\right)^{-1}\mathbf{E}_{\Delta}\mathbf{H}_{\mathbf{c}}^{\dagger}\boldsymbol{\Phi}^{l-1} \qquad (38)$$

【０１１５】
ここで、
【０１１６】
【数２０】

$$\mathbf{E}_{\Delta} = \underbrace{\left[\cdots \mathbf{0} \cdots\right.}_{L_{G_1}}\underbrace{\mathbf{I} \cdots \left.\mathbf{0} \cdots\right]}_{L_{G_2}} \in \mathbf{C}^{T \times TL_G} \qquad (39)$$

【０１１７】
であると共に、ここで、
【０１１８】
【数２１】

$$\boldsymbol{\Phi}^{l} = \left[\left(\sigma_{\mathbf{x}}^{2} - \sigma_{\mathbf{e}}^{2}\right)\mathbf{H}_{\mathbf{c}}\mathbf{H}_{\mathbf{c}}^{\dagger} + \sigma^{2}\mathbf{I}\right] \qquad (40)$$

【０１１９】
であり、更に以下のように表される。
【０１２０】

【数２２】

$$\sigma_{\mathbf{v}}^2 \simeq \frac{1}{LT}\sum_{n=1}^{L}\mathbf{z}^t[n]^t\mathbf{z}^t[n] \qquad (41)$$

【０１２１】
減算器の出力における残りのスクランブルベクトルを加えた雑音は、空間的及び時間的に相関があるということが強調されるべきである。時間相関は、以下の処理に対して実際の影響を与えないが、しかし、空間相関は、この点において重要な役割を持っている。空間白色雑音は、相関行列$\mathbf{K}_{\mathbf{v}}^t$の単純なCholesky因数分解によって容易に獲得されることができる。
【０１２２】
提案２：相関行列$\mathbf{K}_{\mathbf{v}}^t$は正であると定義されるので、Cholesky因数分解が常に適用できる。これは、以下のように書くことができる。
【０１２３】
【数２３】

$$\mathbf{K}_{\mathbf{v}}^t = \mathbf{L}\mathbf{L}^t \qquad (42)$$

【０１２４】
ここで、Lは下三角行列である。そして、順方向フィルタ、及び逆方向フィルタとして"$\mathbf{F}^t = \mathbf{L}^{-1}\mathbf{F}^t$"、及び"$\mathbf{B}^t = \mathbf{L}^{-1}\mathbf{B}^t$"を適用することによって、雑音を加えたスクランブルベクトルの相関行列は、単位行列に等しくなる。
【０１２５】
証明：証明は、自明である。
【０１２６】
【数２４】

$$\begin{aligned}\mathbf{K}_{\mathbf{v}}^t &= \left[(\sigma_{\mathbf{x}}^2-\sigma_{\mathbf{w}}^2)\mathbf{L}^{-1}\mathbf{B}^t\mathbf{B}^{tt}(\mathbf{L}^{-1})^t+\sigma^2\mathbf{L}^{-1}\mathbf{F}^t\mathbf{F}^{tt}(\mathbf{L}^{-1})^t\right]\\ &= \mathbf{L}^{-1}\left[(\sigma_{\mathbf{x}}^2-\sigma_{\mathbf{w}}^2)\mathbf{B}^t\mathbf{B}^{tt}+\sigma^2\mathbf{F}^t\mathbf{F}^{tt}\right](\mathbf{L}^{-1})^t=\mathbf{L}^{-1}\mathbf{K}_{\mathbf{v}}^t(\mathbf{L}^{-1})^t=\mathbf{I} \qquad (43)\end{aligned}$$

【０１２７】
図３のシミュレーションによって提案されたアルゴリズムの正当性を立証する。
【０１２８】
下記においては，等化器の出力は、以下の式のように、常に空間白色化と共に考察される。
【０１２９】
【数２５】

$$\mathbf{z}^t[n] = \mathbf{G}^t[0]\mathbf{x}[n] + \mathbf{v}^t[n]$$

【０１３０】
ここで、"$\mathbf{K}_{\mathbf{v}}^t = \mathbf{I}$"、及び"$\mathbf{G}^t[0] = \mathbf{L}^{-1}$"である。
【０１３１】
ＭＭＳＥ基準はＳＮＲ（ＳＮ比）"$\mathrm{tr}\{\mathbf{G}^t[0]^t\mathbf{K}_{\mathbf{v}}^t\mathbf{G}^t[0]\}$"を最大限にすることに対応する、ということに注目することは興味深い。適応フィルタに関して与えられた最大ＳＮＲ

が、“$\sigma_v^2 = \sigma_w^2$”に関して達成される。

【０１３２】

「Ｃ．包括的なＡＰＰ復号器」

反復の間に交換された全ての確率的な量は、以下で定義される。ＭＩＭＯ等化器の復調部分の目的は、新しい通信モデル（１０）に従った行列Ａのシンボルのビットに関する付帯的な情報を供給することである。ここで、“$G^t[0]$”は、“Ｔ × Ｔ”と同等の一様なＭＩＭＯ通信路の役割を果たすと共に、“$v^t[n]$”は、白色雑音を加えた残りの干渉である。シンボル“$a_{i,j}[n]$”の全てのビットに関する対数比率ＡＰＰは、以下のように定義される。

【０１３３】

【数２６】

$$\lambda^{Ld}_{(i,j)}[n] = \ln \frac{\Pr\left[a_{(i,j)}[n]=1 \,|\, \mathbf{z}^t[n]\right]}{\Pr\left[a_{(i,j)}[n]=0 \,|\, \mathbf{z}^t[n]\right]} \qquad (11)$$

【０１３４】

簡単な省略化によって、以下の式が得られる。

【０１３５】

【数２７】

$$\lambda^{Ld}_{(i,j)}[n] = \ln \frac{\sum_{\mathbf{a}\in\Omega^{(0)}_{(i,j)}} \Pr^t\left[\mathbf{x}[n]=\mu(\mathbf{a}) \,|\, \mathbf{z}^t[n]\right]}{\sum_{\mathbf{a}\in\Omega^{(0)}_{(i,j)}} \Pr^t\left[\mathbf{x}[n]=\mu(\mathbf{a}) \,|\, \mathbf{z}^t[n]\right]} \qquad (12)$$

【０１３６】

ここで、

【０１３７】

【数２８】

$$\Omega^{(\varepsilon)}_{(i,j)} = \left\{\mathbf{a}\in\mathbb{S}_2^{qT},\ a_{(i,j)}=\varepsilon\right\} \qquad (13)$$

【０１３８】

である。対数比率ＡＰＰは、以下のように拡張されることができる。

【０１３９】

【数２９】

$$\lambda^{Ld}_{(i,j)}[n] = \ln \frac{\sum_{\mathbf{a}\in\Omega^{(0)}_{(i,j)}} p\left(\mathbf{z}^t[n] \,|\, \mathbf{x}=\mu(\mathbf{a})\right) \Pr^t\left[\mathbf{a}[n]=\mathbf{a}\right]}{\sum_{\mathbf{a}\in\Omega^{(0)}_{(i,j)}} p\left(\mathbf{z}^t[n] \,|\, \mathbf{x}=\mu(\mathbf{a})\right) \Pr^t\left[\mathbf{a}[n]=\mathbf{a}\right]} \qquad (14)$$

【０１４０】

完璧な時空間インタリーブが、以下に示すように、それらの限界の数字成分の積に以前に加わった確率を広めることを認めると仮定する。

【０１４１】

【数３０】

$$\lambda_{(i,j)}^{ld}[n]=\ln\frac{\sum_{\mathbf{a}\in\Omega_{(i,j)}^{(1)}}p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}=\mu(\mathbf{a})\right)\prod_{(i,j)}\Pr^{l}\left[a_{(i,j)}[n]=a_{(i,j)}\right]}{\sum_{\mathbf{a}\in\Omega_{(i,j)}^{(0)}}p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}=\mu(\mathbf{a})\right)\prod_{(i,j)}\Pr^{l}\left[a_{(i,j)}[n]=a_{(i,j)}\right]}\qquad(15)$$

【０１４２】
$\nu^{l}[n]$が、あらゆる手段（例えば、Cholesky因数分解）によって空間白色化されたと仮定すると、ユークリッドのメトリックが、以下のように確率の評価に使用されることができる。

【０１４３】
【数３１】

$$p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}\right)\propto\exp\left\{-\frac{1}{2\sigma^{2}}\left\|\mathbf{z}^{l}[n]-\mathbf{G}^{l}[0]\mathbf{x}\right\|^{2}\right\}\qquad(16)$$

【０１４４】
一般的な場合において、雑音ベクトルの相関行列$\mathbf{K}_{\nu}^{l}$が、以下のように確率を評価する際に考慮されなければならない。

【０１４５】
【数３２】

$$p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}\right)\propto\exp\left\{-\left(\mathbf{z}^{l}[n]-\mathbf{G}^{l}[0]\mathbf{x}\right)^{T}\mathbf{K}_{\nu}^{-1}\left(\mathbf{z}^{l}[n]-\mathbf{G}^{l}[0]\mathbf{x}\right)\right\}\qquad(17)$$

【０１４６】
ターボ復調器の分野において確立されたように、付帯的な対数確率レートが、方程式から推論される。

【０１４７】
【数３３】

$$\xi_{(i,j)}^{ld}[n]=\ln\frac{\sum_{\mathbf{a}\in\Omega_{(i,j)}^{(1)}}p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}=\mu(\mathbf{a})\right)\prod_{(u,v)\neq(i,j)}\Pr^{l}\left[a_{(u,v)}[n]=a_{(u,v)}\right]}{\sum_{\mathbf{a}\in\Omega_{(i,j)}^{(0)}}p\left(\mathbf{z}^{l}[n]\middle|\mathbf{x}=\mu(\mathbf{a})\right)\prod_{(u,v)\neq(i,j)}\Pr^{l}\left[a_{(u,v)}[n]=a_{(u,v)}\right]}\qquad(18)$$

【０１４８】
ＭＩＭＯ復調器が提供する全ての対数の付帯的な情報サンプルは、全てのブロック"$p=1,\cdots,P$"から収集されて、時空間デインタリーブΠ^{-1}のあと、その出力復号器が全ての符号化ワードビットに関する対数の付帯的な確率比率を供給するという前提で、以下に示すような単純な観測ベクトル"$\zeta^{l,c}\in\mathbb{R}^{N}$"に再配置される。

【０１４９】
【数３４】

$$\xi^{l,c}[n]=\ln\frac{\Pr\left[c[n]=1\middle|\mathbf{c},\zeta^{l,c}/\left\{\zeta^{l,c}[n]\right\}\right]}{\Pr\left[c[n]=0\middle|\mathbf{c},\zeta^{l,c}/\left\{\zeta^{l,c}[n]\right\}\right]}\qquad(19)$$

【０１５０】
時空間デインタリーブΠのあと、ベクトル$\zeta^{l,c}$は、１つがデータ$\mathbf{c}$の各ブロックのた

めのものである、以前の対数確率レートのP個の行列$\Pi^{i,p}$に広められる。各ブロック"p = 1, …, P"に関して，$\Pi^{i,p}$に対する入力は、以下の方程式によって与えられる。

【０１５１】

【数３５】

$$\pi^{l}_{(i,j)}[n] = \ln \frac{\Pr^{l}\left[a_{(i,j)}[n]=1\right]}{\Pr^{l}\left[a_{(i,j)}[n]=0\right]} \tag{20}$$

【０１５２】

これは、軟判定ベクトル$x^{l}[n]$が、以下のように書き直され得ることを意味する。

【０１５３】

【数３６】

$$\bar{\mathbf{x}}^{l}[n] = \frac{1}{2}\sum_{\mathbf{x}\in C^{T}} \mathbf{x} \prod_{\{i,j\}} \left\{1 + \left(2\mu^{-1}_{(i,j)}(\mathbf{x}) - 1\right)\tanh\left(\frac{\pi^{l}_{(i,j)}[n]}{2}\right)\right\} \tag{21}$$

【０１５４】

「Ｄ．スフィア復号された修正リストによる準最適ＭＩＭＯ復調」

"$O(2^{\Pi^{T}})$"ではその基数が様々である信号点配置"$\Omega^{(0)}_{(i,j)}$"及び"$\Omega^{(1)}_{(i,j)}$"の全体のサブセットのポイント計数が、高次の変調、及び／または多数の送信アンテナのために急速に過出力になるので、最適ＭＩＭＯ復調は、高ビットレート通信シナリオが発生した場合は保持できない。確率値の注意深い分析は、それらの大きな数が取るに足らないということを示す。従って、本発明の有意の貢献として、我々は，詳細なポイント計数を、無視できない確率のみを含む、リストとも呼ばれるはるかに小さなサブセット"$L^{(0)}_{(i,j)}$"及び"$L^{(1)}_{(i,j)}$"のポイント計数に交換することを提案する。その場合に、対数の付帯的な確率比率は，以下のようになる。

【０１５５】

【数３７】

$$\xi^{l,d}_{(i,j)}[n] = \ln \frac{\sum_{\mathbf{a}\in\Omega^{(1)}_{(i,j)}} p\left(\mathbf{z}^{l}[n] \mid \mathbf{x}=\mu(\mathbf{a})\right) \prod_{\{u,v\}\neq\{i,j\}} \Pr^{l}\left[a_{(u,v)}[n]=a_{(u,v)}\right]}{\sum_{\mathbf{a}\in\Omega^{(0)}_{(i,j)}} p\left(\mathbf{z}^{l}[n] \mid \mathbf{x}=\mu(\mathbf{a})\right) \prod_{\{u,v\}\neq\{i,j\}} \Pr^{l}\left[a_{(u,v)}[n]=a_{(u,v)}\right]} \tag{18}$$

【０１５６】

幾何学的表現において、それらは，注意深く選択されたポイント（例えば，制限がないポイントML、またはポイントMLそれ自身）を中心とする半径$\underline{r}$の球における格子点を含む。スフィア復号器の変更バージョンは、これらのリスト［参考文献２］を示すために、いくらかの成功と共に使用された。選択された球の半径$\underline{r}$は，対応するＭＩＭＯ復調器の品質，及び複雑さを制御する。

【０１５７】

「Ｅ．ＰＩＣ復調アルゴリズム」

スフィア復号器は、複雑過ぎると考えられ得る。復号器の複雑さを更に減少させる１つの方法は，各次元に関して独立してＡＰＰのログを生成することである。反復Ｉに関して，成分$\underline{x}$に関する判定変数は、以下のようになる。

【０１５８】

【数３８】

$$r_t^l[n] = \left(\mathbf{g}_t^\dagger \mathbf{g}_t\right)^{-1} \mathbf{g}_t^\dagger \left(\mathbf{z}^l[n] - \sum_{k \neq t} \mathbf{g}_k \overline{x_k}\right)$$

【０１５９】

ここで、g_kは、行列G1[0]の列$\underline{k}$である。その場合に、$r^l_t[n]$が平均値x_tを有するガウス変数と仮定して、比率ＡＰＰのログ“$\xi^l_{t,j}[n]$, j= 1,…, q”が計算され、以下のように、分散が推定される。

【０１６０】

【数３９】

$$\hat{\sigma}_{t,l}^2 = \frac{1}{L}\sum_{n=1}^{L}\left|r_t^l[n]\right|^2 - \sigma_x^2$$

【０１６１】

第１の反復復調器は、以下の方法におけるＳＩＣ技術を用いて改良され得る。

【０１６２】

・初期化

例えば、それらの信号電力対干渉電力比（ＳＩＲ：signal to interference ratio）の関数として、次数を減少しながら成分を分類する。

【０１６３】

【数４０】

$$SIR(x_1) > SIR(x_2) > \cdots > SIR(x_T)$$

【０１６４】

ここで、

【０１６５】

【数４１】

$$SIR(x_i) = \frac{\mathbf{g}_i^\dagger \mathbf{g}_i}{\sum_{k \neq i} \mathbf{g}_k^\dagger \mathbf{g}_k + \sigma^2}$$

【０１６６】

である。再帰性は、繰り返し回数ｔを１からＴまでとする（For t = 1 to T）。計算は、以下を実行する。

【０１６７】

【数４２】

$$r_t^1[n] = \left(\mathbf{g}_t^\dagger \mathbf{g}_t\right)^{-1} \mathbf{g}_t^\dagger \left(\mathbf{z}^1[n] - \textstyle\sum_{k<t} \mathbf{g}_k \tilde{x}_k[n]\right)$$

【０１６８】

$r^1_{i,j}[n]$が平均値x_iを有するガウス変数と仮定して、比率ＡＰＰのログ“$\zeta^1_{(i,j)}[n]$，$j=1,\cdots,q$”を計算すると共に、以下のように、分散を推定する。

【０１６９】

【数４３】

$$\hat{\sigma}^2_{z,i} = \frac{1}{L}\sum_{n=1}^{L}\left|r^1_i[n]\right|^2 - \sigma^2_x$$

【０１７０】

平均値x^iが既知であるガウス変数$r^1_{i,j}[n]$のＭＭＳＥ推定を計算する。

【０１７１】

【数４４】

$$\hat{x}_i[n] = E\left\{x_i[n]\,\middle|\,r^1_i[n]\right\}$$

【０１７２】

『別表III』

『参考文献一覧』

[1] S. Ariyavisitakul, "Turbo Space-Time Processing to Improve Wireless Channel Capacity," IEEE Trans. Commun., vol. COM-48, no. 8, pp. 1347-1358, Aug. 2000.
[2] B. M. Hochwald, S. Ten Brink, "Achieving Near-Capacity on a Multiple-Antenna Channel," IEEE Trans. Commun., vol. 51, no. 3, pp. 389-399, March 2003.
[3] R. Visoz, A.O. Berthet, "Iterative Decoding and Channel Estimation for Space-Time BICM over MIMO Block Fading Multipath AWGN Channel," IEEE Trans. Commun., vol. COM-51, no. 8, pp. 1358-1367, Aug. 2003.
[4] X. Wang, H.V. Poor, "Iterative (Turbo) Soft-Interference Cancellation and Decoding for Coded CDMA," IEEE Trans. Commun., vol. COM-47, no. 7, pp. 1046-1061, July 1999.

【図面の簡単な説明】

【０１７３】

【図１】ＶＢＬＡＳＴの概念を示す図である。

【図２】ＳＴ－ＢＩＣＭ通信の一般的なモデルを示す図である。

【図３】本発明の一実施例に適合する反復受信機の構造を示す図である。

【図４】例えば第２番目の反復から、もしくは次の反復から使用され得る反復受信機の構造を示す図である。

【符号の説明】

【０１７４】

１　エラー訂正チャンネル符号化
２　インタリーブ
３　空間逆多重化
４－１～４－Ｍ　変調
５－１～５－Ｍ　成形フィルタ処理
６－１～６－Ｍ　送信アンテナ
８　初期推定値
９　順方向フィルタ
１０　減算器
１１　判定アルゴリズム
１２　再帰的逆方向フィルタ

１３　復号器
１４　時空間デインターリーブ処理
１５　時空間インターリーブ処理
１６　重み付けされたデータの再構成処理
EV　符号化装置
TX　送信アンテナ
DE　復号化手段
RX　受信アンテナ

【図１】

FIG.1

【図２】

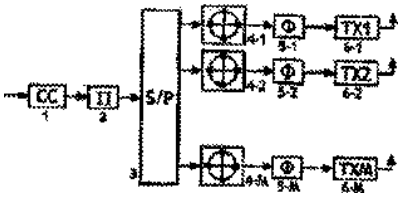


FIG.2

【図３】

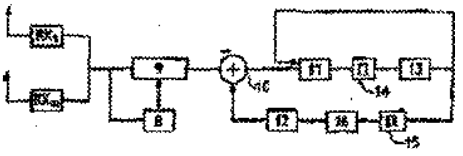


FIG.3

【図４】

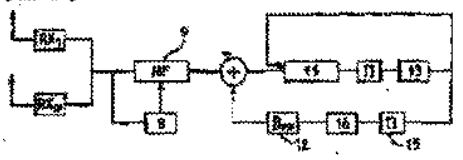


FIG.4

フロントページの続き

(72)発明者 ラファエル・ヴィソズ
フランス・Ｆ－９２１３０・イッシー－レ－ムリノー・ブールヴァール・ガンベッタ・２８
(72)発明者 サミ・シュトゥル
フランス・Ｆ－７５０１３・パリ・リュ・エスキロル・９
(72)発明者 アントワーヌ・ペシテ
フランス・Ｆ－９２２９０・シャトネー・マラブリ・アヴニュ・ジャン－ジョーレ・７２・パティ
マン・ア・エスカリエ・３

審査官 吉田 隆之

(56)参考文献 IEEE Transactions on Vehicular Technology , Vol.50 No.4, pp1176-1182

(58)調査した分野(Int.Cl., ＤＢ名)
H04J

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 1 670 168 B1

(12) **EUROPEAN PATENT SPECIFICATION**

(45) Date of publication and mention of the grant of the patent:
28.07.2010 Bulletin 2010/30

(51) Int Cl.:
H04L 1/06 *(2006.01)* ***H04B 7/06*** *(2006.01)*

(21) Application number: **04292980.2**

(22) Date of filing: **13.12.2004**

(54) **Method, system and device for transmitting uniformly distributed data in MIMO telecommunication systems**

Verfahren, System und Vorrichtung zur gleichmässig verteilten Datenübertragung in MIMO-Übertragungssystemen

Procédé, système et dispositif pour la transmission des données également distribuées dans les systèmes MIMO

(84) Designated Contracting States:
AT BE BG CH CY CZ DE DK EE ES FI FR GB GR HU IE IS IT LI LT LU MC NL PL PT RO SE SI SK TR

(43) Date of publication of application:
14.06.2006 Bulletin 2006/24

(73) Proprietors:
- **Mitsubishi Electric Information Technology Centre Europe B.V.**
  1119 NS Schiphol Rijk (NL)
  Designated Contracting States:
  FR
- **MITSUBISHI DENKI KABUSHIKI KAISHA**
  Chiyoda-ku
  Tokyo 100-8310 (JP)
  Designated Contracting States:
  AT BE BG CH CY CZ DE DK EE ES FI GB GR HU IE IS IT LI LT LU MC NL PL PT RO SE SI SK TR

(72) Inventors:
- **Brunel, Loïc**
  CS 10806
  35708 Rennes Cedex 7 (FR)
- **Gresset, Nicolas**
  CS 10806
  35708 Rennes Cedex 7 (FR)
- **Boutros, Joseph**
  75013 Paris (FR)

(74) Representative: **Saint-Marc-Etienne, Christophe Andréas**
Cabinet Le Guen & Maillet,
5, place Newquay,B.P. 70250
35802 Dinard Cedex (FR)

(56) References cited:
EP-A- 1 511 212 **US-B1- 6 356 528**

- **SANDHU ET AL: "Space-frequency bit-interleaved coded modulation for MIMO-OFDM/OFDMA systems" IEEE 802.16 BROADBAND WIRELESS ACCESS WORKING GROUP, 4 November 2004 (2004-11-04), XP002332928 Retrieved from the Internet: URL:http://www.ieee802.org/16/tge/contrib/C80216e-04_533.pdf> [retrieved on 2005-06-21]**
- **SALVEKAR ET AL: "Multiple-Antenna Technology in WiMAX Systems" INTEL TECHNOLOGY JOURNAL, vol. 8, no. 3, 20 August 2004 (2004-08-20), XP002332929 ISSN: 1535-864 Retrieved from the Internet: URL:ftp://download.intel.com/technology/it j/2004/volume08issue03/art05_multiantenna/vol8_art05.pdf> [retrieved on 2005-06-21]**

Note: Within nine months of the publication of the mention of the grant of the European patent in the European Patent Bulletin, any person may give notice to the European Patent Office of opposition to that patent, in accordance with the Implementing Regulations. Notice of opposition shall not be deemed to have been filed until the opposition fee has been paid. (Art. 99(1) European Patent Convention).

Printed by Jouve, 75001 PARIS (FR)

EP 1 670 168 B1

Description

[0001] The present invention relates to a method for transmitting data in a telecommunication system including at least one transmitter provided with a plurality of transmitting antennas and at least one receiver provided with at least one receiving antenna, which method includes a bit encoding step for generating encoded data bits, a bit interleaving step for permutating said encoded data bits, and a modulation step for producing symbols representative of the permutated bits, each symbol being represented by a predetermined number of bits intended to be transmitted over communication channels established between the transmitting and receiving antennas.

[0002] Telecommunication systems in which a plurality of antennas are used at a receiver end and/or at a transmitter end of a wireless link are called Multiple Input Multiple Output systems (further referred to as MIMO systems). MIMO systems have been shown to offer large transmission capacities compared to those offered by single antenna systems. In particular, MIMO capacity increases linearly with the number of transmitting or receiving antennas, whichever the smallest, for a given Signal-to-Noise Ratio and under favourable uncorrelated channel conditions. MIMO techniques are thus likely to be used in future wireless systems intended to provide large spectral efficiencies or, alternatively, reduce the transmitting power required for obtaining a spectral efficiency equivalent to that which is obtained in current telecommunication systems. Such MIMO techniques will very likely be combined with multi-carrier modulation techniques like OFDM (standing for Orthogonal Frequency Division Multiplex) and MC-CDMA (standing for Multi-Carrier-Code Division Multiple Access) techniques, whose use in future wireless systems is also considered.

[0003] A particular type of MIMO systems makes use of a Bit Interleaved Coded Modulation technique, further referred to as BICM, according to which the transmitter includes a channel encoder intended to apply an encoding, e.g. by means of a convolutional code or of a turbo code, to uncoded data bits, and to provide an encoded bit stream to an interleaver. This interleaver will then deliver permutated bits, which are to be divided into word sequences intended to be transformed into a series of coded symbols represented each by a plurality of bits, the bits representing a same symbol being intended to be transmitted during a same time interval, called symbol period, by respective transmitting antennas.

[0004] Transmitted symbols are to be decoded on the receiver end, which is usually performed in MIMO systems of the BICM type by means of an iterative space-time decoder, which decoder is intended to produce estimates of coded bits constituting the transmitted symbols. The spatial diversity induced by the use of multiple transmitting and receiving antennas eases such a decoding, since this diversity provides a larger amount of information than that which would be provided by a single signal transmitted through a single communication channel.

[0005] The inventors have observed that increasing the diversity of input data provided to a space-time decoder enables said decoder to converge towards more reliable estimates of the coded bits on the basis of which said data has been generated. This may be construed as obtaining better decoding performance by feeding the decoder with data having a higher quality, i.e. a richer content.

[0006] The highest diversity which may be perceived at the receiver end of a MIMO system is determined, on the one hand, by a space-related property of the system, i.e. by the number of receiving antennas, and, on the other hand by time-related properties of the system, i.e. the number of different communication channel states occurring during transmission of a given codeword or a minimum code distance. The minimum code distance is defined by a threshold number of bit differences between two codewords beneath which threshold said codewords are identical.

[0007] The maximum obtainable diversity may thus be expressed in the form of a product between the number of receiving antennas and the lowest value of the above-described time-related parameters.

[0008] The American patent US 6,356,528 discloses a method for transmitting data in a telecommunication system including at least one transceiver provided with a plural number of transmitting antennas. The method enhances the performance of a system utilizing interleaving and transmit diversity by shuffling the symbols transmitted on one channel with respect to the symbols transmitted through another channel. The goal of the shuffle is to make sure that even if the different transmission paths from the different transmitters become correlated, the performance degradation is minor.

[0009] The European patent application EP 1,511,212, published on 02.03.2005, discloses a system which includes a special bit interleaving step which includes a bit multiplexing sub-step for extracting from encoded data bits a plurality of bit sequences, a sequence interleaving sub-step for permutating the bits of each bit sequence produced by the demultiplexing sub-step, a sequence storing and segmenting sub-step for memorizing all permutated sequences produced by the sequence interleaving sub-step and a segment shifting substep for simultaneously applying circular shifts to segments of different permutated sequences produced by the sequence storing and segmenting sub-step.

[0010] The article of Sandhu et al., entitled "Space-frequency bit-interleaved coded modulation for MIMO-OFDM/OFDMA systems", IEEE 802.16 BRODABAND WIRELESS ACCESS WORKING GROUP, 4 November 2004, discloses a system which includes an interleaver in which linearly coded QAM symbols are placed in diagonals along the space-frequency codeword in order to extract full diversity at full rate.

[0011] Salvekar et a., "Multiple-Antenna Technology in WiMAX systems", Intel technology Journal, vol. 8, N°.

3, 20 August 2004, discloses a system in which the interleaver consists in serial-to-parallel multiplexing of incoming bits to antennas, interleaving on each antenna and forward circular shifting of the bits on each antenna by a cyclic lome shift.

[0012] Current interleaving techniques do not enable to use to its full extent the diversity theoretically offered by MIMO systems, since consecutive bit sequences delivered by known interleavers and intended to be transmitted during a same symbol period usually include bits which were adjacent to each other within the original encoded bit stream, which reduces the time-related diversity of the data sent to the receiver and in turn limits the performance of the space-time decoder included in this receiver.

[0013] The invention aims at solving the aforementioned problem, by providing a method for transmitting data in a MIMO system, which method involves an interleaving scheme enabling to use to its full extent the diversity theoretically offered by such a system, with respect to both space and time, of the data intended to be decoded at the receiver end of this system.

[0014] Indeed, a method according to the opening paragraph is characterized according to the invention in that, the communication channels featuring a predefined number Ni of channel inputs, the bit interleaving step includes:

> a bit demultiplexing step for dispatching a frame of encoded data bits into a number of bit sequences equal to the predefined number Ni of channel inputs,
> a sequence interleaving step for permutating the bits of each bit sequence produced by the demultiplexing step, and
> a sequence allocation step for forming groups of Nt interleaved bit sequences successively produced by the sequence interleaving step, all sequences of a same group being intended to be transmitted through Nt channels allocated to a same bit rank for all transmitting antennas.

[0015] The invention enables to compound the spatial diversity obtained by the use of multiple communication channels established between the transmitting and receiving antennas with a high diversity with respect to time of the data transmitted through said channels.

[0016] The invention ensures, by virtue of the demultiplexing step, an essentially homogeneous distribution of the encoded data bits over the different channel inputs, which ensures that successive bits will be transmitted over different channel realizations, and thus favours data diversity as perceived at the receiver end of a telecommunication system in which a method according to the invention is used.

[0017] The sequence interleaving step will preferably be carried out in such a way that Li successive bits belonging to a same demultiplexed sequence end up included in Li different and equal-sized subdivisions of said sequence after execution of the sequence interleaving step.

[0018] Such a preferred embodiment of the invention enables optimal de-correlation of successive uncoded bits, which might otherwise stay linked together in cases where the bit encoding technique makes use of a convolutional code, according to which technique a same uncoded information bit is involved in the generation of L.n successive encoded bits, n being the number of encoded bits delivered by a convolutional channel encoder when said encoder is fed with a given number k of successive uncoded bits and L being a code constraint length. The inventors have observed that the use of such a variant of the invention may enable an essentially homogeneous distribution on all channel inputs and a mapping on different symbol periods of a given number of consecutive encoded bits, which given number is a function of Li and Ni and will be greater than or equal to L.n provided Li is properly chosen. Furthermore, this variant of the invention may be implemented at a relatively low cost by means of a plurality of identical and easily built interleaving modules.

[0019] According to a first variant of the invention, the bit sequence interleaving step includes:

- a random bit permutation step for randomly permutating all bits included in each sequence produced by the demultiplexing step,
- a sequence storing and segmenting step for memorizing all permutated sequences produced by the random bit permutation step and dividing said sequences into segments including each a number of bits equal to the predefined number Ni of channel inputs,
- a segment permutation step for simultaneously applying permutations to segments of different permutated sequences produced by the sequence storing and segmenting step, so that two bits having identical positions within two respective segments included in any group of Ni segments end up having differentiated bit positions in corresponding respective permutated segments, and
- a concatenation step for reassembling all shifted segments related to a same initial permutated sequence into an interleaved sequence.

[0020] The segment permutation step carried out in this first variant of the invention enables to compensate for a vulnerability to errors which is heterogeneous from one bit of a mapping constellation to be used for transposing the encoded bits into modulated symbols. Indeed, if said constellation is represented graphically, it appears that, if one chooses an initial symbol represented by an initial point of the constellation and having for label a set of bits to be transmitted over the communication channels, it appears that modifying the value of a first or of a second one of said bits will lead to respective first or second points located at different first and second dis-

tances with respect to the initial point. The segment permutation step tends to redistribute the different bits to be transmitted over changing bit weights, and thus to homogenize the vulnerability to errors featured by the various bits.

[0021] In an advantageous embodiment of the invention, the segment permutation step will be carried out by simultaneously applying permutations to segments of different permutated sequences produced by the sequence storing and segmenting step, so that each shifted segment ends up shifted by one bit with respect to its adjacent segments.

[0022] This embodiment enables to implement the segment permutation step by means of simple components, such as shifting registers.

[0023] In a preferred embodiment of this first variant of the invention, one of the segments of a given permutated sequence produced by the interleaving step will remain unchanged during the segment permutation step, corresponding segments of each jth other permutated sequence being then simultaneously shifted by j bits.

[0024] This embodiment requires a limited number of sequence-shiftings while producing the above-mentioned advantages in terms of diversity, and is thus remarkable for its efficiency.

[0025] If the permutated sequences produced by the bit interleaving step are to be directly transmitted once the modulation step has been carried out, the predefined number Ni of channel inputs will be chosen equal to a product M.Nt between the number M of bits representative of a symbol and the number Nt of transmitting antennas.

[0026] Alternatively, if the modulation step is followed by a space-time encoding step in the course of which bits of K successive symbols are to be combined together before transmission by means of the Nt transmitting antennas, the predefined number Ni of channel inputs is chosen equal to M.K, where M is the number of bits representative of any given symbol.

[0027] According to a second variant of the invention, which may be used alternatively or cumulatively with the first variant described above when the communication channels are expected to feature Nc successive sets of physical properties, the encoded data bits produced by the encoding step form a superframe to be dispatched into Nc blocks ofNi sequences each in the course of the bit demultiplexing step.

[0028] Communication channels which are expected to successively feature different sets of communication conditions over successive symbol periods are called block-fading channels. Such block-fading channels are essentially invariant for the duration of each set of communication conditions, and the inventors have observed that bits transmitted during time slots corresponding to different blocks have in theory no chance of interfering with each other. This allows to summarily dispatch a superframe into blocks of sequences, the sequences of each block being then processed more cautiously since the bits included in sequences belonging to a same block will be able to interfere with each other during a same invariance period of the block-fading channels. This simple dispatching of the superframe followed by a more constraining block-per-block processing thus enables optimal allocation of processing ressources.

[0029] The dispatching of the superframe may be carried out simply by allocating any two consecutive bits of the superframe to two different transmitting antennas, but also by allocating any two consecutive bits of the superframe to two different blocks of sequences, both solutions ensuring that the bits included in the superframe will be uniformly distributed over all channel realizations.

[0030] According to a third variant of the invention, the bit demultiplexing step will not be carried out by simply placing each jth encoded bit of an initial frame, modulo Ni, in a corresponding bit sequence, but instead by placing two encoded bits separated by Ni-1 other encoded bits in different bit sequences.

[0031] Such a variant of the bit demultiplexing step enables to place consecutive encoded bits over different demultiplexed sequences according to a scheme which is not dictated by a structure of the code used for originally encoding the encoded bits. More particularly, if a specific bit has a high potential for being affected by an encoding error, said specific bit having then a period of occurrence dictated by the code structure, said periodic bit will not always be transmitted through a same channel input, the above described preferred embodiment of the bit demultiplexing step instead ensuring that different occurrences of this periodic bit will be sent through different channel inputs.

[0032] According to a preferred embodiment of this third variant, the bit demultiplexing step is carried out by placing an encoded bit of rank j in a bit sequence of rank j+IP((j-1)/Ni), modulo Ni, where IP((j-1)/Ni) represents an integer part of a ratio (j-1)/Ni.

[0033] According to one of its hardware-oriented aspects, the invention also relates to a telecommunication system including at least one transmitter provided with a plural number Nt of transmitting antennas and at least one receiver provided with at least one receiving antenna, which transmitter includes bit encoding means for generating encoded data bits, bit interleaving means for permutating said encoded data bits, and modulation means for producing symbols representative of the permutated bits, each symbol being represented by a predetermined number of bits intended to be transmitted over communication channels established between the transmitting and receiving antennas, which channels feature a predefined number Ni of channel inputs,

system characterized in that the bit interleaving means include:

- bit demultiplexing means for dispatching a frame of encoded data bits into a number of bit sequences equal to the predefined number Ni of channel inputs,
- sequence interleaving means for permutating the

bits of each bit sequence produced by the demultiplexing means, and

- sequence allocation means for forming groups of Nt interleaved bit sequences successively produced by the sequence interleaving means, all sequences of a same group being intended to be transmitted through Nt channels allocated to a same bit rank for all transmitting antennas.

[0034] According to a first variant of such a telecommunication system, which enables to place consecutive encoded bits over different symbol periods according to a scheme which is not dictated by a structure of the code used for originally encoding the encoded bits, the sequence interleaving means include:

- random bit permutation means for randomly permutating all bits included in each sequence produced by the demultiplexing step,
- sequence storing and segmenting means for memorizing all permutated sequences produced by the random bit permutation means and dividing said sequences into segments including each a number of bits equal to the predefined number Ni of channel inputs,
- segment permutation means for simultaneously permutating segments of different permutated sequences produced by the sequence storing and segmenting means, so that two bits having identical positions within two respective segments included in any group of Ni segments end up having differentiated bit positions in corresponding respective permutated segments, and
- concatenation means for reassembling all shifted segments related to a same initial permutated sequence into an interleaved sequence.

[0035] According to another of its hardware-related aspects, the invention also relates to a communication device provided with a plural number Nt of transmitting antennas, which transmitter includes bit encoding means for generating encoded data bits, bit interleaving means for permutating said encoded data bits, and modulation means for producing symbols representative of the permutated bits, each symbol being represented by a predetermined number of bits intended to be transmitted over communication channels established between the transmitting and receiving antennas, which channels feature a predefined number Ni of channel inputs,
device characterized in that the bit interleaving means include:

- bit demultiplexing means for dispatching a frame of encoded data bits into a number of bit sequences equal to the predefined number Ni of channel inputs,
- sequence interleaving means for permutating the bits of each bit sequence produced by the demultiplexing means, and
- sequence allocation means for forming groups of Nt interleaved bit sequences successively produced by the sequence interleaving means, all sequences of a same group being intended to be transmitted through Nt channels allocated to a same bit rank for all transmitting antennas.

[0036] The characteristics of the invention mentioned above, as well as others, will emerge more clearly from a reading of the following description given in relation to the accompanying figures, amongst which:

Fig. 1 is a block diagram showing a highly simplified MIMO telecommunication system in which the invention is to be used;
Fig.2 is a block diagram showing interleaving means comprised in a transmitter included in a MIMO telecommunication system according to the invention;
Fig.3 is a diagram showing how such an interleaver operates;
Fig.4 is a diagram showing a first splitting technique for dispatching a superframe into frames to be processed according to the invention;
Fig.5 is a diagram showing a second splitting technique for dispatching a superframe into frames to be processed according to the invention; and
Fig.6 is a diagram showing a preferred technique for carrying out a bit demultiplexing step according to a variant of the invention.

[0037] Fig.1 diagrammatically shows a telecommunication system including at least one transmitter TR and one receiver REC, intended to exchange signals through multiple communication channels CHNL established between a predetermined integer number Nt of transmitting antennas (ta1,ta2...taNt) and a predetermined integer number Nr of receiving antennas (ra1,ra2...raNr).

[0038] The transmitter TR shown in the example depicted here includes a channel encoder CHENC intended to apply an encoding, e.g. by means of a convolutional code or of a turbo code, to uncoded data bits Uncb, and to provide a binary stream of encoded data bits Tb to be transmitted. The transmitter TR includes interleaving means INTL intended to generate permutated bits Pb, such an interleaving being useful for a later processing on the receiver side, since it will allow to obtain uncorrelated data. The permutated bits Pb are to be divided into sequences of M.Nt bits each, which bit sequences are then mapped, i.e. transformed into a series of coded symbols Zk by a mapping and modulation module MAPMD, each symbol Zk thus being represented by M successive bits. Successive symbols Zk may then be fed via an unshown transmitting RF module to the Nt transmitting antennas (ta1,ta2...taNt), in which case the communication channels CHNL will feature a number Ni=M.Nt of inputs. Alternatively, the coded symbols Zk may be fed to an optional space-time encoder SPTENC, which is shown here in dashed lines and is to perform an additional

processing of the coded symbols Zk before their transmission. The optional space-time encoder SPTENC may, for example, carry out a space-time encoding step in the course of which K successive M-bit symbols Zk (for k=1 to K) are to be combined together before transmission by means of the Nt transmitting antennas (ta1,ta2... taNt), in which case the communication channels CHNL will feature a number Ni=M.K of inputs. A suitable combination of K successive M-bit symbols Zk may for example consist in a linear combination obtained by multiplying a vector representing said successive symbols Zk by a space-time encoding matrix.

[0039] The receiver REC shown in the example depicted here includes a space-time decoder SPTDEC intended to produce decoded data bits Decb which should ultimately correspond to the originally uncoded data bits Uncb. This space-time decoder SPTDEC includes a space-time detector DET intended to process data carried by signals received by means of the receiving antennas (ra1,ra2...raNr), and to produce likelihood values Rib related to estimates of the transmitted permutated bits Pb, which likelihood values are intended to be de-interleaved by a de-interleaver DINTL which is to output soft likelihood values Rb related to estimates of bits included in the binary stream of encoded data bits Tb. A bit decoder included in the receiver REC, further referred to as channel decoder CHDEC, is intended to generate the decoded data bits Decb on the basis of said likelihood values Rb.

[0040] According to a loop structure commonly used in the art, the space-time detector DET will make use of *a priori* information Pra generated in the course of previous decoding steps, and issued in the form of extrinsic information Exd by the channel decoder CHDEC through interleaving means INTR, which interleaving means are identical to the interleaving means INTL included in the transmitter TR.

[0041] The inventors have observed that increasing the diversity of the data transmitted through the multiple communication channels CHNL established between the transmitting and receiving antennas (ta1,ta2...taNt) and (ra1,ra2...raNr) enables the space-time decoder SPTDEC to converge towards more reliable estimates of the coded bits on the basis of which said data has been generated. The inventors have thus aimed at maximizing the time-related diversity of the data transmitted over the transmitting antennas (ta1,ta2...taNt).

[0042] Fig.2 depicts bit interleaving means INTL which enable to obtain such an increased diversity. In this specific embodiment of the invention, the bit interleaving means INTL include:

- bit demultiplexing means DMX for extracting from the encoded data bits Tb a plurality Ni of bit sequences Bsq1...BsqNi, each sequence Bsqi (for i=1 to Ni) corresponding to one channel input, i.e. to one bit to be transmitted by one of the Nt transmitting antennas described above, or to one of the M bits representing one of the K successive symbols to be combined together in the course of a later space-time encoding step, said bit demultiplexing means DMX being adapted for placing each jth encoded bit, modulo Ni, in a corresponding Njth bit sequence,
- sequence interleaving means SRI1 ... SRINi for permutating the bits of each bit sequence Bsq1 ...BsqNi produced by the demultiplexing means DMX, and
- sequence allocation means for forming groups of Nt interleaved bit sequences successively produced by the sequence interleaving means SRI1... SRINi, all sequences of a same group being intended to be transmitted through Nt channels allocated to a same bit rank for all transmitting antennas.

[0043] In this embodiment of the invention, the bit interleaving means INTL further include:

- sequence storing and segmenting means for memorizing all permutated sequences Isq1...IsqNi produced by the sequence interleaving means SRI1...SRINi and dividing each sequence Isqj into successive segments Psgj having each a predetermined length of Ni bits in this embodiment of the invention, and
- segment permutation means for simultaneously permutating segments Psg1...PsgNi of different permutated sequences Isq1...IsqNi produced by the sequence storing and segmenting means BUF1, so that two bits having identical positions within two respective segments included in any group of Ni segments, e.g. bits 25 and 26 in the first two segments of sequences Isq 1 and Isq2 included in a first group of Ni segments as shown in the next Figure, end up having differentiated bit positions in corresponding respective permutated segments, which is obtained in this example by applying circular shifts to said segments so that each resulting shifted segment Ssgj (for j=1 to Ni) ends up shifted by one bit with respect to its adjacent segments Ssgj-1 and Ssgj+1.

[0044] The sequence interleaving means SRI1... SRINi will preferrably be constituted by Ni identical interleaving modules arranged in parallel as shown here.

[0045] In this specific embodiment of the invention, the sequence storing and segmenting means are constituted by a first buffer BUF1 and the sequence permutation means include a parallel array of Ni circularly shiftable registers SRG1... SRGNi respectively intended to store successive series of segments Psgl...PsgNi simultaneously produced by said buffer BUF1. The contents of one of these registers, in this example SRG1, will remain unchanged, while the contents of each jth other register SRGj (for j=2 to Ni) will be submitted to a circular shift of (j-1) bits.

[0046] In this specific embodiment of the invention, the sequence allocation means included in the bit interleaving means INTL include a second buffer BUF2 which is

adapted to receive and to store successive series of shifted segments Ssg1 ... SsgNi and to reassemble all shifted segments Ssgj related to a same initial permutated sequence Isqj (for j=1 to Ni) into an interleaved sequence Psqj, the bit interleaving means INTL further including multiplexing means MX intended to receive all interleaved sequences Psqj delivered by the sequence allocation means BUF2 and to multiplex said interleaved sequences Psqj in order to form the stream of permutated bits Pb intended to be fed to the mapping and modulation module described in the previous Figure.

[0047] In alternative embodiments of the invention, the segments Ssg1 ... SsgNi may be sent in parallel to multiple antenna-specific mapping and modulation modules before their transmission by the respective antennas associated with said modules.

[0048] Fig. 3 illustrates the operation of the bit interleaving means described above, in a specific embodiment of the invention where the transmitter includes Nt=2 transmitting antennas, the symbols to be transmitted being mapped on two bits b1 and b2, so that M=2 and Ni=4.

[0049] During a bit demultiplexing step DMXS, a plurality of bit sequences of which only bit sequence Bsq1 is shown here, is extracted from a frame Tb of Si successive encoded data bits referenced 1, 2, 3, 4, 5, 6, 7, 8, 9... Si. The bits intended to form bit sequence Bsq1 are shown in white on this drawing, other bits intended to form part of three other sequences corresponding each to another one of the channel inputs being shown in a three different shades of grey. Each bit sequence Bsqj (for j=1 to Ni=4) thus includes Si/Ni bits referenced j, 4+j, 8+j, 12+j, 16+j...Si+j-Ni.

[0050] During a sequence interleaving step SRIS, the Si/Ni bits of each bit sequence Bsqj produced by the demultiplexing step DMXS are then permutated, only the permutation of bit sequence Bsq1 resulting in a permutated bit sequence Isq1 being shown here.

[0051] In the example depicted here, the sequence interleaving step SRIS is carried out in such a way that Li successive encoded bits 1, 5, 9, 13, 17, etc. end up included in Li different segments of length Ni after execution of the sequence interleaving step SRIS.

[0052] During a sequence storing and segmenting step each permutated sequence Isqj (for j=1 to 4) produced by the sequence interleaving step SRIS is then memorized in a buffer and divided into successive segments having each a predermined length equal to Ni=4 in this example.

[0053] During a segment permutation step SPS, circular shifts of j-1 bits are then simultaneously applied in parallel to Ni different segments Psgj (for j=2 to Ni=4) belonging each to one of the permutated sequences Isqj produced by the interleaving step SRIS, so that each resulting shifted segment Ssgj ends up shifted by one bit with respect to its adjacent segments Ssgj-1 and Ssgj+1. In this embodiment of the invention, each of the successive segments Psg1 belonging to the permutated sequence Isq1 produced by the interleaving step SRIS will remain unchanged during each segment permutation step SPS.

[0054] As can be seen in this Figure, the demultiplexing step DMXS and shifting step SRGS carried out in the method according to the invention enable to place consecutive encoded bits, for example the bits referenced 25, 26, 27 and 28, over different symbol periods. An allocation step ALS carried out by rearranging the contents of the above mentioned second buffer intended to store and concatenate all shifted segments ensures that groups G1 and G2 each including Nt=2 interleaved bit sequences successively produced by the sequence interleaving step are allocated to a same bit, respectively b1 and b2 for both transmitting antennas A1 and A2.

[0055] The allocation step enables an essentially homogeneous distribution of the encoded data bits over the various bits to be transmitted by the different transmitting antennas, which ensures that successive bits, for example the bits referenced 25, 26, 27 and 28, will be transmitted over different channels, and thus favours data diversity as perceived at the receiver end of a telecommunication system in which a method according to the invention is used.

[0056] The sequence interleaving step SRIS additionally enables optimal de-correlation of successive uncoded bits, which might otherwise stay linked together in cases where the bit encoding technique makes use of a convolutional code, according to which technique a same uncoded information bit is involved in the generation of L.n successive encoded bits, n being the number of encoded bits delivered by a convolutional channel encoder when said encoder is fed with a given number k of successive uncoded bits and L being the code constraint length. The inventors have observed that the use of such a variant of the invention may enable an essentially homogeneous distribution on all transmitting antennas and a mapping on different symbol periods of (Li-1)Ni+1 consecutive encoded bits if each segment has a length equal to Ni, the number (Li-1).Ni+1 being greater than or equal to L.n provided Li is properly chosen.

[0057] Fig.4 shows another embodiment of the invention where the communication channels are block-fading channels expected to feature Nc successive sets of physical properties, a superframe SF including Nc frames as described above being then intended to be dispatched in the course of a bit demultiplexing step DMXS into Nc blocks of Ni sequences, each sequence including Si bits and being in turn intended to be processed as described above. In the embodiment described here, Nc is chosen equal to 3 and Ni equal to 4 and the superframe SF is summarily dispatched into three blocks of sequences (Bsq1, Bsq4, Bsq7, Bsq10) (Bsq2, Bsq5, Bsq8, Bsq11) and (Bsq3, Bsq6, Bsq9, Bsq12) of Si/Ni bits each, the sequences of each block B1, B2 or B3 being then processed more cautiously by carrying out a sequence interleaving step SRIS, a segment permutation step SPS, and an allocation step ALS. Such a block-per-block processing is preferable since the bits included in sequences

belonging to a same block will be able to interfere with each other during a same invariance period of the block-fading channels. Such a simple dispatching of the superframe followed by a more constraining block-per-block processing thus enables optimal allocation of processing ressources.

[0058] The dispatching of the superframe SF is carried out in the example depicted in Fig.4 simply by allocating any two consecutive bits of the superframe SF to two different blocks of sequences B1, B2 or B3.

[0059] As shown in Fig.5, the dispatching of the superframe SF may also be carried out by allocating any two consecutive bits of the superframe to two different transmitting antennas, A1 or A2 in this example.

[0060] Both solutions ensure that the bits included in the superframe SF will be uniformly distributed over all channel realizations, which will enable to preserve the highest possible diversity at the receiver side.

[0061] It should be noted here that the bits included in the various sequences should preferrably first be dispatched over either the various blocks or the various antennas, before being dispatched over the various bits allocated to said antennas, since the channels corresponding to different blocks or to different antennas are theoretically more different from each other than channels corresponding to different bits to be transmitted over a same antenna and a same block. Attributing a higher priority to a dispatching of successive bits over different blocks or antennas thus enables to optimize the diversity obtained by virtue of the invention.

[0062] Fig.6 shows a preferred embodiment of the demultiplexing step DMXS which is not carried out by simply placing each jth encoded bit of an initial frame, modulo Ni, in a corresponding bit sequence, but instead by placing two encoded bits separated by Ni-1 other encoded bits in different bit sequences, which is obtained in this example by placing an encoded bit of rank j in a bit sequence of rank j+IP((j-1)/Ni), modulo Ni, where IP((j-1)/Ni) represents an integer part of a ratio (j-1)/Ni.

[0063] Such an embodiment of the bit demultiplexing step DMXS enables to place consecutive encoded bits over different demultiplexed sequences according to a scheme which is not dictated by a structure of the code used for originally encoding the encoded bits. More particularly, if a specific bit has a high potential for being affected by an encoding error, said specific bit having then a period of occurence dictated by the code structure, said periodic bit will not always be transmitted through a same channel input, the above described preferred embodiment of the bit demultiplexing step instead ensuring that different occurrences of this periodic bit will be sent through different channel inputs.

Claims

1. A method for transmitting data in a telecommunication system including at least one transceiver provided with a plural number Nt of transmitting antennas and at least one receiver provided with at least one receiving antenna, which method includes a bit encoding step (CHENC) for generating encoded data bits (Tb), a bit demultiplexing step (DMX) for dispatching a frame of encoded data bits into a number of bit sequences (Bsqi) equal to a predefined number of Ni channel inputs, a bit interleaving step (INTL) for permutating said encoded bits, and a modulation step (MAPMD) for producing symbols (Zk) representative of the permutated bits (Pb), each symbol (Zk) being represented by a predetermined number of bits and transmitted over Nt communication channels established between the transmitting and receiving antennas, which channels feature a predefined number Ni of channel inputs,
method **characterized in that** the bit interleaving step includes :

- a sequence interleaving step (SRi) which includes

  a random bit permutation step (SRIS) for randomly permutating all bits included in each sequence (Bsqi) produced by the demultiplexing step,
  a sequence storing and segmenting step (BUF1) for memorizing all permutated sequences (Isqi) produced by the random bit permutation step and dividing said sequences into segments (Psgj) including each a number of bits equal to the predefined number Ni of channels inputs, and
  a segment permutation step (SPS) for simultaneously applying permutations to segments (Psgj) of different permutated sequences (Isqi) produced by the sequence storing and segmenting step, so that two bits having identical positions within two respective segments included in any group of Ni segments end up having differentiated bit positions in corresponding respective permutated segments, and

- a sequence allocation step (ALS) for forming groups of interleaved bit sequences (Psq) successively produced by the sequence interleaving step, and for allocating all bit sequences of a same group to a same bit rank for all transmitting antennas.

2. A Method as claimed in claim 1, **characterized in that** the sequence interleaving step (SRi) further includes a concatenation step for reassembling all shifted segments related to a same initial permutated sequence into an interleaved sequence.

3. A method as claimed in claim 1 or 2, **characterized**

in that the segment permutations step (SPS) is carried out by simultaneously applying circular shifts to segments of different permutated sequences produced by the sequence storing and segmentaing step (BUF1), so that each shifted segment ends up shifted by one bit with respect to it adjacent segments.

4. A method as claimed in claim 3, **characterized in that** one of the segments of a given permutated sequence produced by the interleaving step remains unchanged during the segment permutation step, corresponding segments of each jth other permutated sequence being then simultaneously shifted by j bits..

5. A method as claimed in claims 1 to 4, **characterized in that** the communication channels being expected to feature Nc successive sets of physical properties, the encoded data bits produced by the encoding step form a superframe tobe dispatched into Nc blocks of Ni sequences each in the course of the bit demultiplexing step.

6. A method as claimed in claim 5, according to which any two consecutive bits of the superframe will be allocated to two different transmitting antennas.

7. A method as claimed in claim 5, according to which any two consecutive bits of the superframe will be allocated to two different blocks of sequences.

8. A method as claimed in any one of claims 1 to 7, according to which the predefined number Ni of channels inputs is chosen equal to a product M.Nt between the number M of bits representative of a symbol and the number Nt of transmitting antennas.

9. A method as claimed in any one of claims 1 to 7, according to which the modulation step being followed by a space-time encoding step in the course of which bits of K successive symbols are to be combined together before transmission by means of the Nt transmitting antennas, the predefined number Ni of channel inputs is chosen equal to M.K, where M is the number of bits representative of any given symbol.

10. A method as claimed in any one of claims 1 to 9, in which the bit demultiplexing step is carried out by placing two encoded bits separated by Ni-1 other encoded bits in different bit sequences.

11. A method as claimed in claim 10, in which the bit demultiplexing step is carried out by placing an encoded bit of rank j in a bit sequence of rank j+IP((j-1)/Ni), modulo Ni, where IP((j-1)/Ni) represents an integer part of ratio (j-1)/Ni.

12. A telecommunication system including:

- at least one transceiver provided with a plural number Nt of transmitting antennas and at least one receiver provided with at least one receiving antenna, which transmitter includes:
- bit encoding means (CHENC) for generating encoded data bits,
- bit demultiplexing means (DMX) for dispatching a frame of encoded data bits into a number of bit sequences (Bsqi) equal to a predefined number of Ni channel inputs,
- bit interleaving means (INTL) for permutating said encoded data bits, and
- modulation means for producing symbols (Zk) representative of the permutated bits (Pb), each symbol (Zk) being represented by a predetermined number of bits and transmitted over Nt communication channels established between the transmitting and receiving antennas, which channels feature a predefined number Ni of channel inputs,

system characerized in that the bit interleaving means include :

- sequence interleaving means (Sri) which includes :

  random bit permutation means (SRIS) for randomly permutating all bits included in each sequence (BSqi) produced by the demultiplexing means,
  sequence storing and segmenting means (BUF1) for memorizing all permutated sequences (Isqi) produced by the random bit permutation means and dividing said sequences into segments (Psgj) including each a number of bits equal to the predefined number Ni of channels inputs,
  segment permutation means (SPS) for simultaneously applying permutations to segments (Psgj) of different permutated sequences (Isqi) produced by the sequence storing and segmenting means, so that two bits having identical positions within two respective segments included in any group of Ni segments end up having differentiated bit positions in corresponding respective permutated segments, and

- sequence allocation means (ALS) for forming groups of interleaved bit sequences (Psq) successively produced by the sequence interleaving means, and for allocating all bit sequences of a same group to a same bit rank for all transmitting antennas.

13. A telecommunication system as claimed in claim 12, **characterized in that** said sequence interleaving means (SRi) further include concatenation means for reassembling all shifted segments related to a same initial permutated sequence into an interleaved sequence.

14. A communication device provided with a plural number Nt of transmitting antennas, which device includes bit encoding means (CHENC) for generating encoded data bits, bit demultiplexing means (DMX) for dispatching a frame of encoded data bits into a number of bit sequences (Bsqi) equal to a predefined number of Ni channel inputs, bit interleaving means (INTL) for permutating said encoded data bits, and modulation means for producing symbols (Zk) representative of the permutated bits (Pb), each symbol (Zk) being represented by a predetermined number of bits and transmitted over Nt communication channels established between the transmitting and receiving antennas, which channels feature a predefined number Ni of channel inputs, device **characterized in that** the bit interleaving means include :

- sequence interleaving means (Sri) which includes :

  random bit permutation means (SRIS) for randomly permutating all bits included in each sequence (BSqi) produced by the demultiplexing means,
  sequence storing and segmenting means (BUF1) for memorizing all permutated sequences (Isqi) produced by the random bit permutation means and dividing said sequences into segments (Psgj) including each a number of bits equal to the predefined number Ni of channels inputs,
  segment permutation means (SPS) for simultaneously applying permutations to segments (Psgj) of different permutated sequences (Isqi) produced by the sequence storing and segmenting means, so that two bits having identical positions within two respective segments included in any group of Ni segments end up having differentiated bit positions in corresponding respective permutated segments, and

- sequence allocation means (ALS) for forming groups of interleaved bit sequences (Psq) successively produced by the sequence interleaving means, and for allocating all bit sequences of a same group to a same bit rank for all transmitting antennas.

Patentansprüche

1. Verfahren zum Übertragen von Daten in einem Telekommunikationssystem mit wenigstens einem Transceiver, der mit einer Vielzahl Nt von Sendeantennen versehen ist, und wenigstens einem Empfänger, der mit wenigstens einer Empfangsantenne versehen ist, welches Verfahren einen Bit-Codierschritt (CHENC) zum Erzeugen codierter Datenbits (Tb), einen Bit-Demultiplexschritt (DMX) zum Absetzen eines Frames codierter Datenbits in eine Anzahl von Bitsequenzen (Bsqi) gleich einer vordefinierten Anzahl von Ni Kanaleingängen, einen Bit-Verschachtelungsschritt (INTL) zum Permutieren der codierten Bits und einen Modulationsschritt (MAPMD) zum Erzeugen von Symbolen (Zk), die die permutierten Bits darstellen, wobei jedes Symbol (Zk) durch eine vorbestimmte Anzahl von Bits dargestellt wird und über Nt Kommunikationskanäle übertragen wird, die zwischen den Sende- und Empfangsantennen eingerichtet sind, aufweist, welche Kanäle mit einer vordefinierten Anzahl Ni von Kanaleingängen ausgestattet sind,
wobei das Verfahren **dadurch gekennzeichnet ist, dass** der Bit-Verschachtelungsschritt folgendes enthält:

- einen Sequenz-Verschachtelungsschritt (Sri), der folgendes enthält:

  einen Zufalls-Bit-Permutationsschritt (SRIS) zum zufälligen Permutieren aller Bits, die in jeder durch den Demultiplexschritt erzeugten Sequenz (Bsqi) enthalten sind,
  einen Sequenz-Speicherungs- und -Segmentierschritt (BUF1) zum Speichern von allen durch den Zufalls-Bit-Permutationsschritt erzeugten permutierten Sequenzen (Isqi) und zum Aufteilen der Sequenzen in Segmente (Psgj), die jeweils eine Anzahl von Bits gleich der vordefinierten Anzahl Ni von Kanaleingängen enthalten, und
  einen Segment-Permutationsschritt (SPS) zum gleichzeitigen Anwenden von Permutationen auf durch den Sequenz-Speicherungs- und- Segmentierschritt erzeugte Segmente (Psgj) von unterschiedlichen permutierten Sequenzen (Isqi), so dass zwei Bits mit identischen Positionen innerhalb von zwei jeweiligen Segmenten, die in irgendeiner Gruppe von Ni Segmenten enthalten sind, dahin gelangen, differenzierte Bitpositionen in entsprechenden jeweiligen permutierten Segmenten zu haben, und

- einen Sequenz-Zuteilungsschritt (ALS) zum Ausbilden von Gruppen von durch den Se-

quenz-Verschachtelungsschritt aufeinanderfolgend erzeugten verschachtelten Bitsequenzen (Psq) und zum Zuteilen aller Bitsequenzen einer selben Gruppe zu einem selben Bitrang für alle Sendeantennen.

2. Verfahren nach Anspruch 1, **dadurch gekennzeichnet, dass** der Sequenz-Verschachtelungsschritt (SRi) weiterhin einen Verkettungsschritt zum erneuten Zusammensetzen aller verschobenen Segmente in Bezug auf eine selbe anfängliche permutierte Sequenz in eine verschachtelte Sequenz.

3. Verfahren nach Anspruch 1 oder 2, **dadurch gekennzeichnet, dass** der Segment-Permutationsschritt (SPS) durch gleichzeitiges Anwenden zirkularer Verschiebungen auf Segmente von durch den Sequenz-Speicherungs- und -Segmentierschritt (BUF1) erzeugten unterschiedlichen permutierten Sequenzen, so dass jedes verschobene Segment dahin gelangt, in Bezug auf benachbarte Segmente um ein Bit verschoben zu sein.

4. Verfahren nach Anspruch 3, **dadurch gekennzeichnet, dass** eines der Segmente einer durch den Verschachtelungsschritt erzeugten gegebenen permutierten Sequenz während des Segment-Permutationsschritts unverändert bleibt, wobei entsprechende Segmente jeder j-ten anderen permutierten Sequenz dann gleichzeitig um j Bits verschoben werden.

5. Verfahren nach den Ansprüchen 1 bis 4, **dadurch gekennzeichnet, dass**, während erwartet wird, dass die Kommunikationskanäle mit Nc aufeinanderfolgenden Gruppen von physikalischen Eigenschaften ausgestattet sind, die durch den Codierschritt erzeugten codierten Datenbits einen Superframe ausbilden, um im Verlauf des Bit-Demultiplexschritts jeweils in Nc Blöcke von Ni Sequenzen abgesetzt zu werden.

6. Verfahren nach Anspruch 5, gemäß welchem irgendwelche zwei aufeinanderfolgende Bits des Superframes zwei unterschiedlichen Sendeantennen zugeteilt werden.

7. Verfahren nach Anspruch 5, gemäß welchem irgendwelche zwei aufeinanderfolgende Bits des Superframes zwei unterschiedlichen Blöcken von Sequenzen zugeteilt werden.

8. Verfahren nach einem der Ansprüche 1 bis 7, gemäß welchem die vordefinierte Anzahl Ni von Kanaleingängen gleich einem Produkt M Nt zwischen der Anzahl M von Bits, die ein Symbol darstellen, und der Anzahl Nt von Sendeantennen gewählt wird.

9. Verfahren nach einem der Ansprüche 1 bis 7, gemäß welchem dem Modulationsschritt ein Raum-Zeit-Codierschritt folgt, in dessen Verlauf vor einem Senden mittels der Nt Sendeantennen Bits von K aufeinanderfolgenden Symbolen miteinander zu kombinieren sind, die vordefinierte Anzahl Ni von Kanaleingängen gleich M K gewählt wird, wobei M die Anzahl von Bits ist, die irgendein gegebenes Symbol darstellen.

10. Verfahren nach einem der Ansprüche 1 bis 9, wobei der Bit-Demultiplexschritt durch Platzieren von zwei codierten Bits, getrennt durch Ni-1 andere codierte Bits, in unterschiedlichen Bitsequenzen ausgeführt wird.

11. Verfahren nach Anspruch 10, wobei der Bit-Demultiplexschritt durch Platzieren eines codierten Bits eines Rangs j in einer Bitsequenz eines Rangs j+IP((j-1)/Ni) modulo Ni ausgeführt wird, wobei j+IP((j-1)/Ni) einen ganzzahligen Teil eines Verhältnisses (j-1)/Ni darstellt.

12. Telekommunikationssystem, das folgendes enthält:

- wenigstens einen Transceiver, der mit einer Vielzahl Nt von Sendeantennen versehen ist, und wenigstens einen Empfänger, der mit wenigstens einer Empfangsantenne versehen ist, welcher Sender folgendes enthält:

  - eine Bit-Codiereinrichtung (CHENC) zum Erzeugen codierter Datenbits,
  - eine Bit-Demultiplexeinrichtung (DMX) zum Absetzen eines Frames von codierten Datenbits in eine Anzahl von Bitsequenzen (Bsqi) gleich einer vordefinierten Anzahl von Ni Kanaleingängen,
  - eine Bit-Verschachtelungseinrichtung (INTL) zum Permutieren der codierten Datenbits, und
  - eine Modulationseinrichtung zum Erzeugen von Symbolen (Zk), die die permutierten Bits (Pb) darstellen, wobei jedes Symbol (Zk) durch eine vorbestimmte Anzahl von Bits dargestellt wird und über Nt Kommunikationskanäle gesendet werden, die zwischen den Sende- und Empfangsantennen eingerichtet sind, welche Kanäle mit einer vorbestimmten Anzahl Ni von Kanaleingängen ausgestattet sind,

wobei das System **dadurch gekennzeichnet ist, dass** die Bit-Verschachtelungseinrichtung folgendes enthält:

- eine Sequenz-Verschachtelungseinrichtung (Sri), die folgendes enthält:

eine Zufalls-Bit-Permutationseinrichtung (SRIS) zum zufälligen Permutieren aller Bits, die in jeder durch die Demultiplexeinrichtung erzeugten Sequenz (Isqi) enthalten sind,
eine Sequenz-Speicherungs- und -Segmentiereinrichtung (BUFI) zum Speichern von allen permutierten Sequenzen (Isqi), die durch die Zufalls-Bit-Permutationseinrichtung erzeugt sind, und zum Aufteilen der Sequenzen in Segmente (Psgi), die jeweils eine Anzahl von Bits gleich der vordefinierten Anzahl Ni von Kanaleingängen enthalten,
eine Sequenz-Permutationseinrichtung (SPS) zum gleichzeitigen Anwenden von Permutationen auf Segmente (Psgi) von unterschiedlichen permutierten Sequenzen (Isqi), die durch die Sequenz-Speicherungs- und -Segmentiereinrichtung erzeugt sind, so dass zwei Bits mit identischen Positionen innerhalb von zwei jeweiligen Segmenten, die in irgendeiner Gruppe von Ni Segmenten enthalten sind, dahin gelangen, differenzierte Bitpositionen in entsprechenden jeweiligen permutierten Segmenten zu haben, und

- eine Sequenz-Zuteilungseinrichtung (ALS) zum Ausbilden von Gruppen von verschachtelten Bitsequenzen (Psq), die durch den Sequenz-Verschachtelungsschritt aufeinanderfolgend erzeugt sind, und zum Zuteilen von allen Bitsequenzen einer selben Gruppe zu einem selben Bitrang für alle Sendeantennen.

13. Telekommunikationssystem nach Anspruch 12, **dadurch gekennzeichnet, dass** die Sequenz-Verschachtelungseinrichtung (SRi) weiterhin ein Verkettungseinrichtung zum erneuten Zusammensetzen aller verschobener Segmente in Bezug auf eine selbe anfängliche permutierte Sequenz in eine verschachtelte Sequenz enthält.

14. Kommunikationsvorrichtung, die mit einer Vielzahl Nt von Sendeantennen versehen ist, welche Vorrichtung eine Bit-Codiereinrichtung (CHENC) zum Erzeugen codierter Datenbits, eine Bit-Demultiplexeinrichtung (DMX) zum Absetzen eines Frames codierter Datenbits in eine Anzahl von Bitsequenzen (Bsqi) gleich einer vordefinierten Anzahl von Ni Kanaleingängen, eine Bit-Verschachtelungseinrichtung (INTL) zum Permutieren der codierten Bits und eine Modulationseinrichtung zum Erzeugen von Symbolen (Zk), die die permutierten Bits darstellen, wobei jedes Symbol (Zk) durch eine vorbestimmte Anzahl von Bits dargestellt wird und über Nt Kommunikationskanäle übertragen wird, die zwischen den Sende- und Empfangsantennen eingereichtet sind, aufweist, welche Kanäle mit einer vordefinierten Anzahl Ni von Kanaleingängen ausgestattet sind,
wobei die Vorrichtung **dadurch gekennzeichnet ist, dass** die Bit-Verschachtelungseinrichtung folgendes enthält:

- eine Sequenz-Verschachtelungseinrichtung (Sri), die folgendes enthält:

  eine Zufalls-Bit-Permutationseinrichtung (SRIS) zum zufälligen Permutieren aller Bits, die in jeder durch die Demultiplexeinrichtung erzeugten Sequenz (Isqi) enthalten sind,
  eine Sequenz-Speicherungs- und -Segmentiereinrichtung (BUFI) zum Speichern von allen permutierten Sequenzen (Isqi), die durch die Zufalls-Bit-Permutationseinrichtung erzeugt sind, und zum Aufteilen der Sequenzen in Segmente (Psgi), die jeweils eine Anzahl von Bits gleich der vordefinierten Anzahl Ni von Kanaleingängen enthalten,
  eine Sequenz-Permutationseinrichtung (SPS) zum gleichzeitigen Anwenden von Permutationen auf Segmente (Psgi) von unterschiedlichen permutierten Sequenzen (Isqi), die durch die Sequenz-Speicherungs- und- Segmentiereinrichtung erzeugt sind, so dass zwei Bits mit identischen Positionen innerhalb von zwei jeweiligen Segmenten, die in irgendeiner Gruppe von Ni Segmenten enthalten sind, dahin gelangen, differenzierte Bitpositionen in entsprechenden jeweiligen permutierten Segmenten zu haben, und

- eine Sequenz-Zuteilungseinrichtung (ALS) zum Ausbilden von Gruppen von verschachtelten Bitsequenzen (Psq), die durch den Sequenz-Verschachtelungsschritt aufeinanderfolgend erzeugt sind, und zum Zuteilen von allen Bitsequenzen einer selben Gruppe zu einem selben Bitrang für alle Sendeantennen.

Revendications

1. Une méthode pour transmettre des données dans un système de télécommunication incluant au moins un transmetteur équipé avec une pluralité Nt d'antennes de transmission et au moins un récepteur équipé avec au moins une antenne de reception, ladite méthode comprenant une étape de codage de bit (CHENC) pour générer des bits de données codés (Tb), une étape de démultiplage de bits (DMX) pour répartir une trame de bits de données codés

dans un nombre de séquences de bits égal à un nombre prédéfini Ni d'entrées canal, une étape d'entrelacement de bits (INTL) pour permuter lesdits bits codés, et une étape (MAPMD) pour produire des symboles (Zk) représentatifs des bits permutés (Pb), chaque symbole (Zk) étant représenté par un nombre prédéterminé de bits et transmis sur Ni canaux de communication établis entre les antennes de transmission et de réception, lesdits canaux définissant un nombre prédéfini Ni d'entrées canal,
méthode qui est **caractérisée en ce que** l'étape d'entrelacement de bits comprend :

- une étape d'entrelacement de séquence (SRi) qui comprend

  - une étape de permutation aléatoire de bit (SRIS) pour permuter aléatoirement tous les bits inclus dans chaque séquence (Bsqi) produite par l'étape de démultiplexage,
  - une étape de segmentation et de stockage de séquence (BUF1) pour mémoriser toutes les séquences permutées (Isqi) produites par l'étape de permutation aléatoire de bit et pour diviser lesdites séquences en segments (Psgj) incluant chacun un nombre de bits égal au nombre prédéfini Ni d'entrées canal, et
  - une étape de permutation de segments (SPS) pour appliquer simultanément des permutations aux segments (Psgj) de différentes séquences permutées (Isqi) produites par l'étape de segmentation et de stockage de séquence, de sorte que deux bits ayant des positions identiques à l'intérieur de deux segments respectifs inclus dans un quelconque groupe de Ni segments finissent par des positions de bit différenciées dans des segments permutés respectifs correspondants, et

- une étape d'allocation de séquence (ALS) pour former des groupes de Nt séquences de bits entrelacés (Psq) successivement produits par l'étape d'entrelacement de séquence, et pour allouer toutes les séquences de bits d'un même groupe à un même rang de bit pour toutes les antennes de transmission.

2. Méthode selon la revendication 1, **caractérisée en ce que** l'étape d'entrelacement de séquence (SRi) comprend une étape de concaténation pour réassembler tous les sgements décalés relatifs à une même séquence permutée initiale dans une séquence entrelacée.

3. Méthode selonla revendication 1 ou 2, **caractérisée en ce que** l'étape de permutation de segment (SPS) est exécutée en appliquant simultanément des décalages circulaires aux segments of différentes séquences permutées produites par l'étape de segmentation et de stockage de séquence (BUF1), de sorte que chaque segment décalé finit décalé d'un bit par rapport à ce bit de segments adjacents.

4. Méthod selon la revendication 3, **caractérisée en ce que** l'un des segments d'une séquence permutée donnée produite par l'étape d'entrelacement reste inchangé durant l'étape de permutation de segment, des segments correspondants à chaque jième autre sequence permutée étant alors simultanément décalés de j bits.

5. Méthode selon la revendication 1 à 4, **caractérisée en ce que** les canaux de communications étant attendus qu'ils se présentent comme Nc ensembles successifs de propriétés physiques, les bits de données codés produits par l'étape de codage forme une supertrame à répartir dans Nc blocs de Ni séquences chacun au cours de l'étape de démultiplexage de bits.

6. Méthode selon la revendication 5, dans laquelle deux quelconques bits consécutifs de la supertrame seront alloués à deux antennes de transmission différentes.

7. Méthode selon la revendication 5, dans laquelle deux quelconques bits consécutifs de la supertrame seront alloués à deux blocs différents de séquences.

8. Méthode selon l'une quelconque des revendications 1 à 7, dans laquelle le nombre prédéfini Ni d'entrées de canal est choisi égal au produit M.Nt entre le nombre M de bits représentatifs d'un symbole et le nombre NT d'antennes de transmission.

9. Méthode selon l'une quelconque des revendications 1 à 7, dans laquelle l'étape de modulation étant suivie d'une étape de codage temps-espace au cours de laquelle des bits de K symboles successifs sont combinés ensemble avant transmission par des moyens des Nt antennes de transmission, le nombre prédéfini Ni d'entrées canal est choisi égal à M.K, où M est le nombre de bits représentatifs d'un quelconque symbole donné.

10. Méthode selon l'une quelconque des revendications 1 à 9, dans laquelle l'étape de démultiplexage de bit est exécutée en plaçant deux bits codés séparés de Ni-1 autres bits codés dans des séquences de bits différentes.

11. Méthode selon la revendication 10, dans laquelle l'étape de démultiplexage de bit est exécutée en plaçant un bit codé de rang j dans une séquence de bits

de j+IP((j-1)/Ni), modulo Ni, où IP((j-1)/Ni) représente une partie entière du rapport (j-1)/Ni.

12. Un système de télécommunication comprenant :

- au moins un transmetteur équipé avec une pluralité Ni d'antennes de transmission et au moins un récepteur équipé avec au moins une antenne de reception,

ledit transmetteur comprend:

- des moyens de codage de bits (CHENC) générer des bits de données codés,
- des moyens de démultiplage de bits (DMX) pour répartir une trame de bits de données codés dans un nombre de séquences de bits (Bsqi) égal à un nombre prédéfini Ni d'entrées canal
- des moyens d'entrelacement de bits (INTL) pour permuter lesdits bits codés, et
- des moyens pour produire des symbôles (Zk) représentatifs des bits permutés (Pb), chaque symbôle (Zk) étant représenté par un nombre prédéterminé de bits et transmis sur des canaux de communication établis entre les antennes de transmission et de réception, lesdits canaux définissant un nombre prédéfini Ni d'entrées canal,

système qui est **caractérisé en ce que** les moyens d'entrelacement de bit comprennent :

- des moyens d'entrelacement de séquence (Sri) qui comprennent

  - des moyens de permutation aléatoire de bit (SRIS) pour permuter aléatoirement tous les bits inclus dans chaque séquence (Bsqi) produite par les moyens de démultiplexage,
  - des moyens de segmentation et de stockage de séquence (BUF 1) pour mémoriser toutes les séquences permutées (Isqi) produites par les moyens de permutation aléatoire de bit et pour diviser lesdites séquences en segments (Psgj) incluant chacun un nombre de bits égal au nombre prédéfini Ni d'entrées canal, et
  - des moyens de permutation de segments (SPS) pour appliquer simultanément des permutations aux segments (Psgj) de différentes séquences permutées (Isqi) produites par les moyens de segmentation et de stockage de séquence, de sorte que deux bits ayant des positions identiques à l'intérieur de deux segments respectifs inclus dans un quelconque groupe de Ni segments finissent par des positions de bit différenciées dans des segments permutés respectifs correspondants, et

- des moyens d'allocation de séquence (ALS) pour former des groupes de Nt séquences de bits entrelacés (Psq) sucessivement produits par l'étape d'entrelacement de séquence, et pour allouer toutes les séquences de bits d'un même groupe à un même rang de bit pour toutes les antennes de transmission.

13. Système de télécommunication selon la revendication 12, **caractérisé en ce que** lesdits moyens d'entrelacement de séquence (SRi) comprennent des moyens de concaténation pour réassembler tous les sgements décalés relatifs à une même séquence permutée initiale dans une séquence entrelacée.

14. Appareil de communication équipé avec une pluralité Nt d'antennes de transmission, lequel appareil inclut des moyens de codage de bits (CHENC) générer des bits de données codés, des moyens de démultiplage de bits (DMX) pour répartir une trame de bits de données codés dans un nombre de séquences de bits (Bsqi) égal à un nombre prédéfini Ni d'entrées canal, des moyens d'entrelacement de bits (INTL) pour permuter lesdits bits codés, et des moyens pour produire des symbôles (Zk) représentatifs des bits permutés (Pb), chaque symbôle (Zk) étant représenté par un nombre prédéterminé de bits et transmis sur des canaux de communication établis entre les antennes de transmission et de réception, lesdits canaux définissant un nombre prédéfini Ni d'entrées canal,
appareil qui est **caractérisé en ce que** les moyens d'entrelacement de bit comprennent :

- des moyens d'entrelacement de séquence (Sri) qui comprennent

  - des moyens de permutation aléatoire de bit (SRIS) pour permuter aléatoirement tous les bits inclus dans chaque séquence (Bsqi) produite par les moyens de démultiplexage,
  - des moyens de segmentation et de stockage de séquence (BUF1) pour mémoriser toutes les séquences permutées (Isqi) produites par les moyens de permutation aléatoire de bit et pour diviser lesdites séquences en segments (Psgj) incluant chacun un nombre de bits égal au nombre prédéfini Ni d'entrées canal, et
  - des moyens de permutation de segments (SPS) pour appliquer simultanément des permutations aux segments (Psgj) de différentes séquences permutées (Isqi) produites par les moyens de segmentation et de stockage de séquence, de sorte que deux bits ayant des positions identiques à l'intérieur de deux segments respectifs inclus dans un quelconque groupe de Ni seg-

ments finissent par des positions de bit différenciées dans des segments permutés respectifs correspondants, et

- des moyens d'allocation de séquence (ALS) pour former des groupes de Nt séquences de bits entrelacés (Psq) successivement produits par l'étape d'entrelacement de séquence, et pour allouer toutes les séquences de bits d'un même groupe à un même rang de bit pour toutes les antennes de transmission.

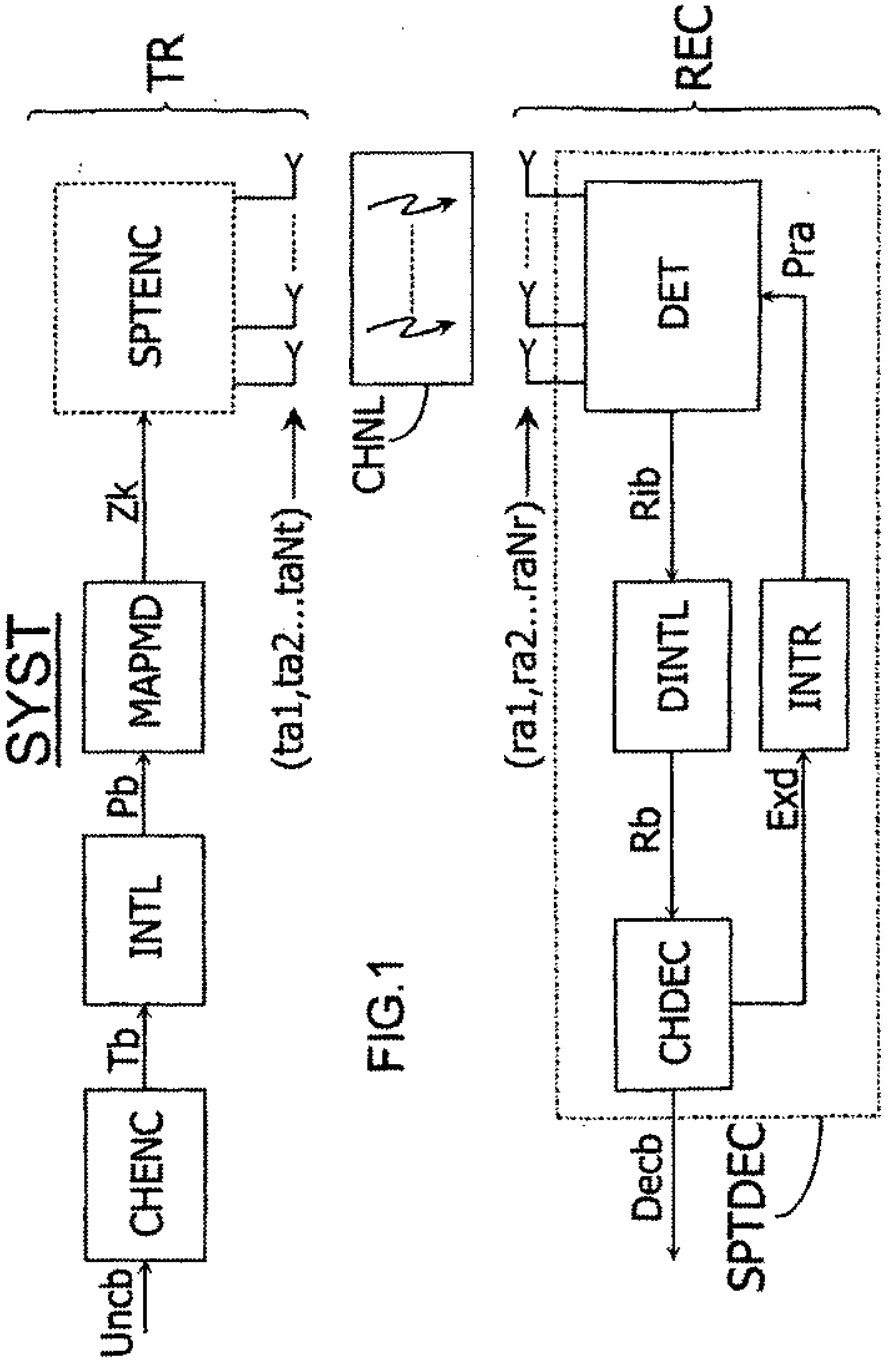

SYST
TR
REC
Uncb
CHENC
Tb
INTL
Pb
MAPMD
Zk
SPTENC
(ta1,ta2...taNt)
CHNL
(ra1,ra2...raNr)
DET
Pra
Rib
DINTL
INTR
Rb
Exd
CHDEC
Decb
SPTDEC

FIG.1

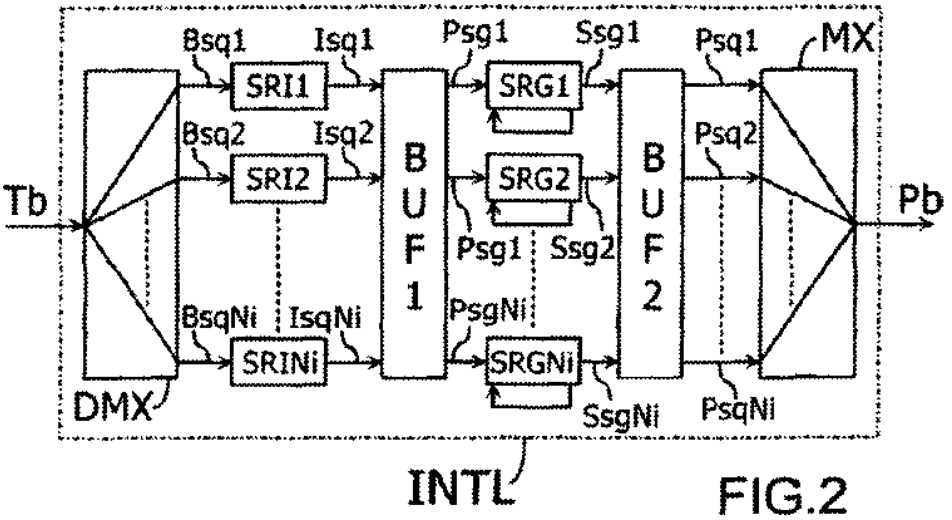

Bsq1
Isq1
Psg1
Ssg1
Psq1
MX
SRI1
SRG1
Bsq2
Isq2
Psq2
SRI2
SRG2
Tb
BUF1
BUF2
Pb
Psg1
Ssg2
BsqNi
IsqNi
PsgNi
SRINi
SRGNi
DMX
SsgNi
PsqNi
INTL

FIG.2

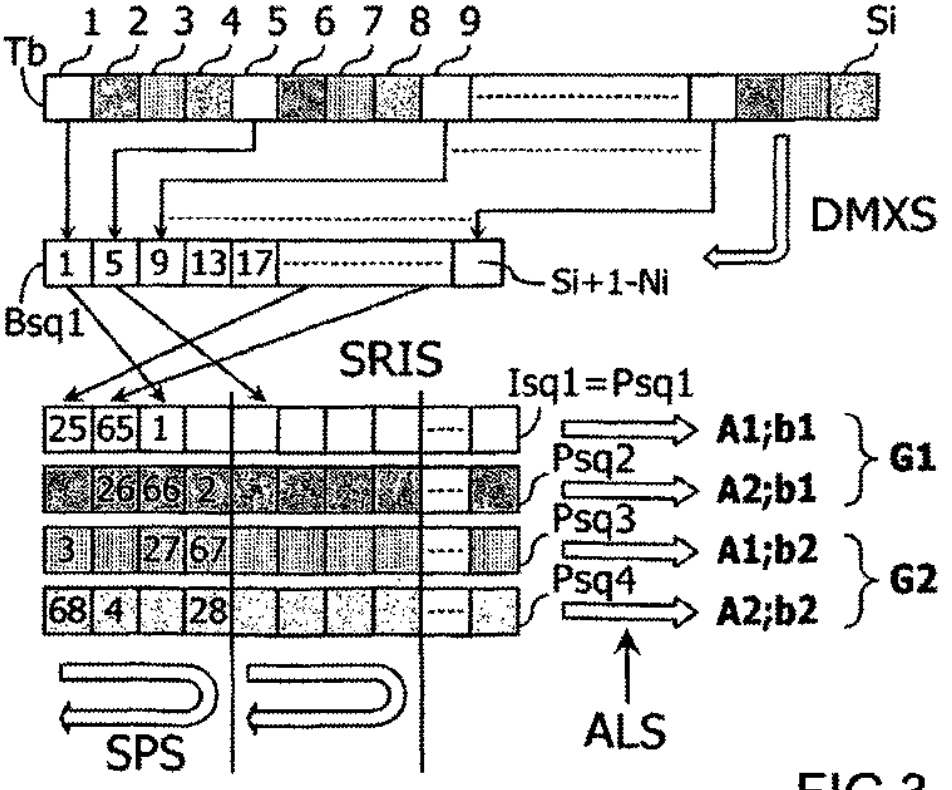

Tb
1 2 3 4 5 6 7 8 9
Si
DMXS
1 5 9 13 17
Si+1-Ni
Bsq1
SRIS
Isq1=Psq1
25 65 1
26 66 2
3 27 67
68 4 28
Psq2
Psq3
Psq4
A1;b1
A2;b1
A1;b2
A2;b2
G1
G2
SPS
ALS

FIG.3

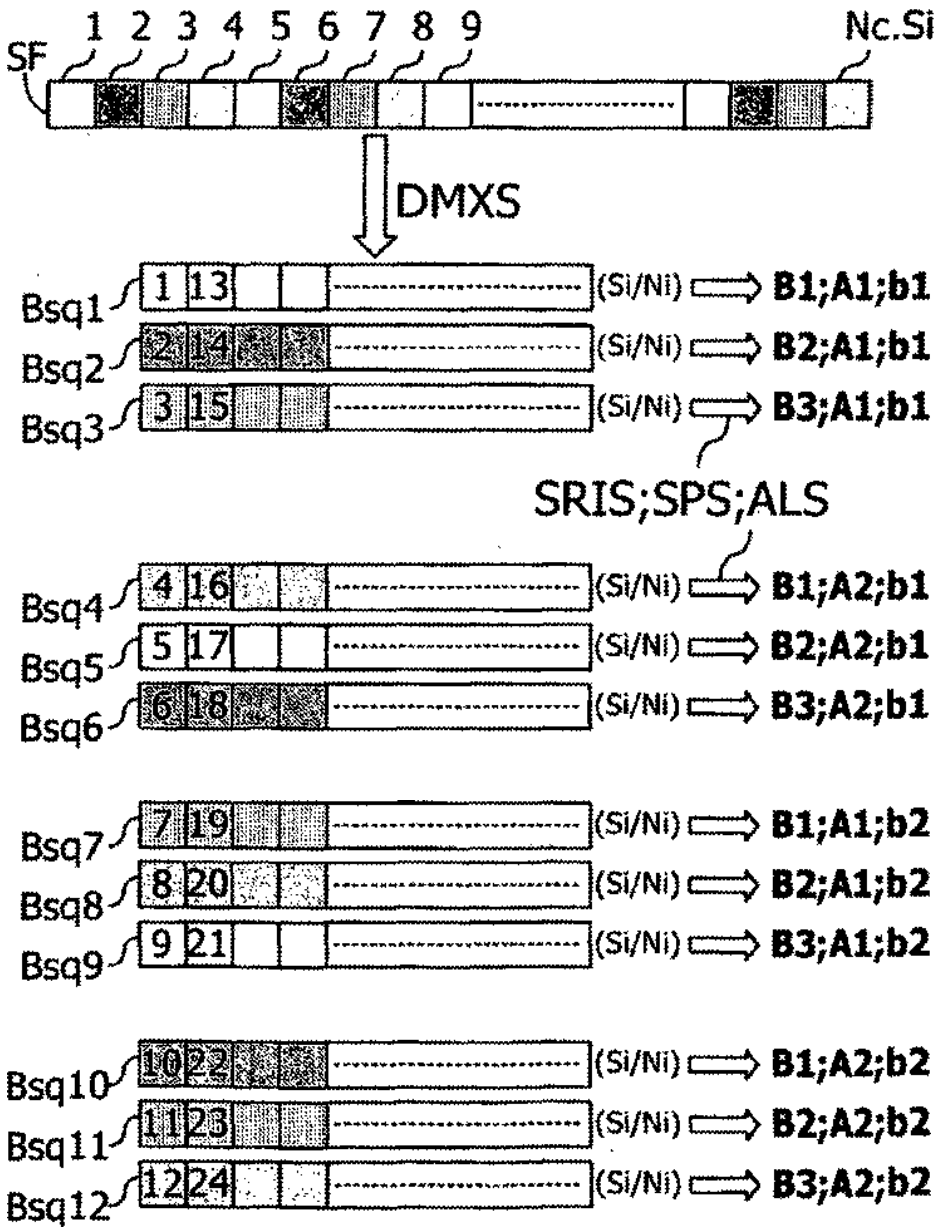

1 2 3 4 5 6 7 8 9
SF
Nc.Si
DMXS
Bsq1
Bsq2
Bsq3
(Si/Ni)
B1;A1;b1
B2;A1;b1
B3;A1;b1
SRIS;SPS;ALS
Bsq4
Bsq5
Bsq6
B1;A2;b1
B2;A2;b1
B3;A2;b1
Bsq7
Bsq8
Bsq9
B1;A1;b2
B2;A1;b2
B3;A1;b2
Bsq10
Bsq11
Bsq12
B1;A2;b2
B2;A2;b2
B3;A2;b2

FIG.4

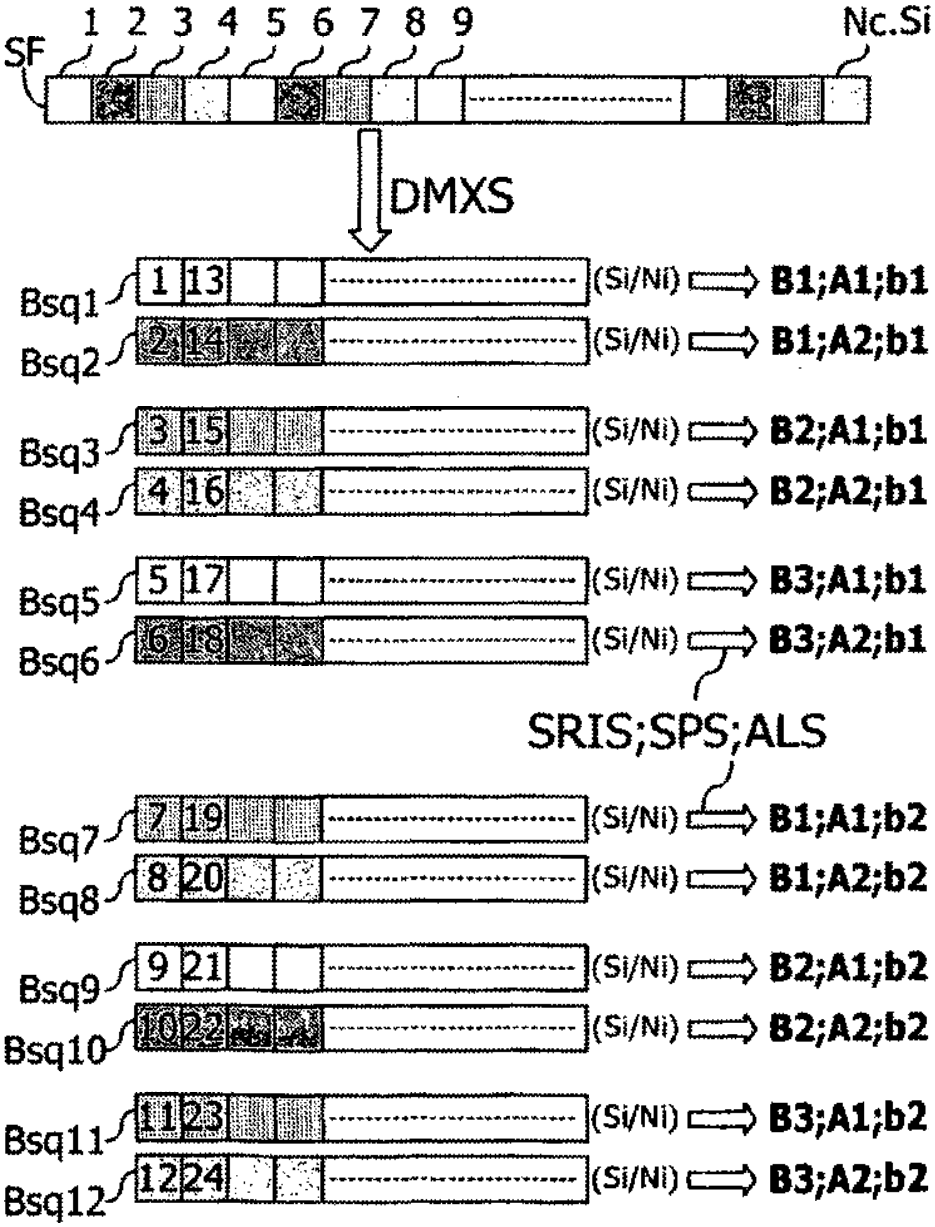

SF
1 2 3 4 5 6 7 8 9
Nc.Si
DMXS
Bsq1
Bsq2
Bsq3
Bsq4
Bsq5
Bsq6
Bsq7
Bsq8
Bsq9
Bsq10
Bsq11
Bsq12
(Si/Ni)
B1;A1;b1
B1;A2;b1
B2;A1;b1
B2;A2;b1
B3;A1;b1
B3;A2;b1
SRIS;SPS;ALS
B1;A1;b2
B1;A2;b2
B2;A1;b2
B2;A2;b2
B3;A1;b2
B3;A2;b2

FIG.5

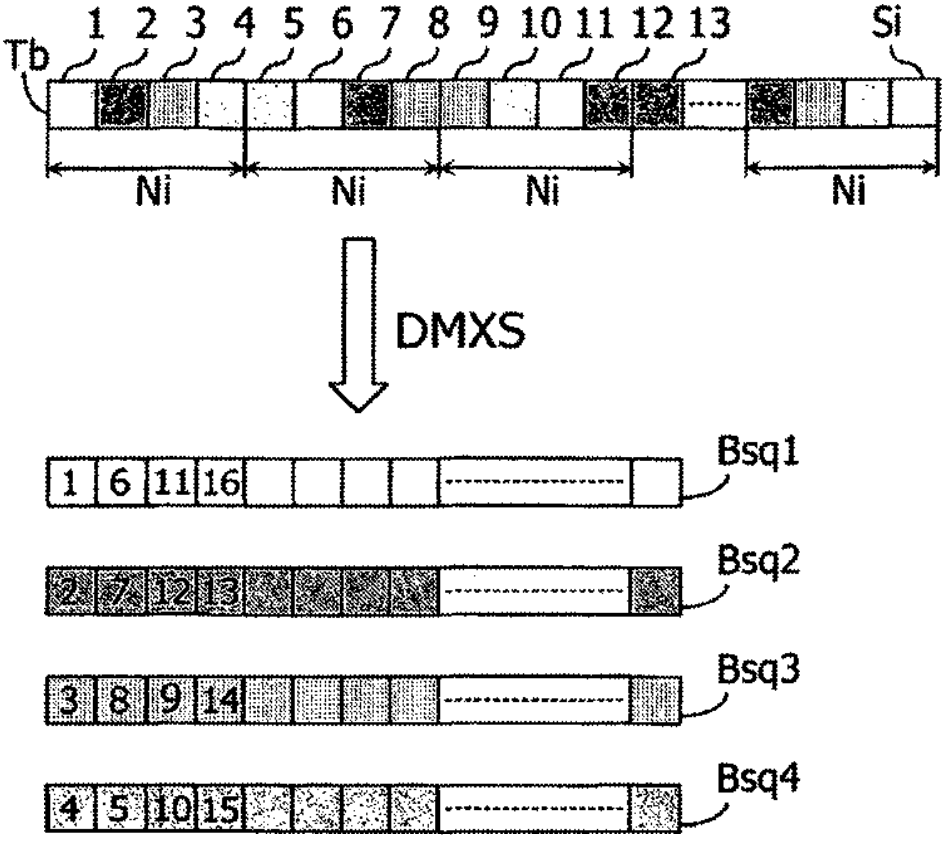

1 2 3 4 5 6 7 8 9 10 11 12 13
Tb
Si
Ni
Ni
Ni
Ni
DMXS
1 6 11 16
Bsq1
2 7 12 13
Bsq2
3 8 9 14
Bsq3
4 5 10 15
Bsq4

FIG.6

REFERENCES CITED IN THE DESCRIPTION

This list of references cited by the applicant is for the reader's convenience only. It does not form part of the European patent document. Even though great care has been taken in compiling the references, errors or omissions cannot be excluded and the EPO disclaims all liability in this regard.

Patent documents cited in the description

- US 6356528 B **[0008]**
- EP 1511212 A **[0009]**

Non-patent literature cited in the description

- **Salvekar.** Multiple-Antenna Technology in WiMAX systems. *Intel technology Journal*, 20 August 2004, vol. 8 (3 **[0011]**

CN101133558B

Method for transmitting data, method for receiving data, transmitter, receiver

Abstract

A method for transmitting data comprising a plurality of bits is described wherein the data is mapped to a plurality of modulation symbols, each modulation symbol comprising at least one more significant bit and at least one less significant bit, at least one parity bit is generated for the plurality of bits and the plurality of bits are mapped to more significant bits of the plurality of modulation symbols and the at least one parity bit is mapped to at least one less significant bit of the plurality of modulation symbols.

Classifications

H04L1/0041 Arrangements at the transmitter end
View 9 more classifications

China
Download PDF Find Prior Art Similar
Other languages
Chinese

Inventor
李援
孙素梅
吴岩

Worldwide applications
2006 CN JP US EP WO *TW*

Application CN2006800069000A events

2005-02-03

Priority to US64981805P

2005-02-03

Priority to US60/649,818

2006-02-02

Application filed by 新加坡科技研究局

2006-02-02

Priority to PCT/SG2006/000019

2008-02-27

Publication of CN101133558A

2010-10-06

Application granted

2010-10-06

Publication of CN101133558B

Info

- Patent citations (17)
- Non-patent citations (1)
- Cited by (28)
- Legal events
- Similar documents
- Priority and Related Applications

External links

- Espacenet
- Global Dossier
- Discuss

Description

Method, reflector and the receiver of the method for emission data, reception data

Technical field

The present invention relates to a kind of method of data, method, reflector, receiver and computer program of reception data launched.

Background technology

In communication system, use the method that detects and correct the mistake that takes place between transmission period to improve reliability.A kind of possibility is to use the low density parity check code (LDPC) described in [1], and it is the capacity approach coding scheme that has attracted many concerns at present.Be equally capacity approach the sign indicating number Turbo code compare, the LDPC sign indicating number has more performance under the situation of big block size, although its decoding complexity is less than the decoding complexity of Turbo code.

Another commercial benefit of LDPC is that the LDPC patent is expired, and therefore, company can use LDPC and needn't pay the expense of intellectual property.

Described the LDPC encoder and use (being called the LDPC coded modulation) with M-ary (M unit) modulator in [2], it provides high power and bandwidth efficiency.The M-ary modulator for example shines upon every log according to Gray $_2$M LDPC coded-bit is mapped to a constellation (constellation) symbol.This mapping is not introduced any stand-by period in transmission, and this is different from the Bit Interleave in the Bit Interleave coded modulation (BICM).By the optimal design of LDPC sign indicating number, the LDPC coded modulation can be near code modulated capacity limitation.

LDPC sign indicating number and LDPC coded modulation have extensively been introduced the next generation wireless communication system, so that carry out reliable and transmission efficiently with high data rate.Combine with quadrature amplitude modulation (QAM) and MIMO-OFDM (multi-input

multi-output-orthogonal frequency division multiplexing), LDPC has shown in high-throughput (HT) wireless local Access Network (WLAN) and has been better than Turbo code.It is still at the competitive candidate to the optional HT pattern in the proposal of IEEE 802.11n standard.The LDPC sign indicating number also has been proposed for the standard IEEE 802.16 of the broadband wireless access (BWA) that also needs high order modulation.Also have manyly, wherein might in QAM UWB system, use LDPC about LDPC being designed into the proposal in ultra broadband (UWB) radio communication.

In traditional Bit Interleave coded modulation (BICM) scheme of using convolution code or Turbo code, for coded-bit being replaced the diverse location in the distinct symbols, the bit interleaver between encoder and modulator is essential.

In the IEEE802.11a standard, bit interleaver is handled the displacement of two steps as follows:

Step 1:

Adjacent coded-bit is mapped on the non-adjacent modulation symbol, with the burst error of avoiding introducing by single symbol and channel relevancy.

When N represented codeword size, first displacement was defined as

$$i=(N/d)(k?mod?d)+floor(k/d),\quad k=0,\ 1,\ K,\ N-1,\quad (1)$$

Wherein d is an interleave depth, and floor (.) expression is no more than the maximum integer of this parameter.

Step 2:

Adjacent coded-bit alternately is mapped to the higher significant bit (MSB) of constellation and low significant bit (LSB), with the long-term operation of the LSB that avoids having low reliability.

Second displacement is defined as

$$j=s\times floor(i/s)+(i+N-floor(d\times i/N))mod?s,$$

i=0, 1, K, N-1 (2)

Wherein Determine by modulation order M.

Two kinds of methods for the use of LDPC coded modulation are described below.

First method is to use as mentioned above and replaces the LDPC coded-bit according to the bit interleaver of 802.11a, with reference to [2].

Second method is that the LDPC code word is mapped directly in the M-ary symbol.This explains below.

(N, K) the LDPC sign indicating number by its parity check matrix H=[$\underline{H}_1$ $\underline{H}_2$] provide K and N (wait to be encoded) the respectively number of information bit and the number of LDPC coded-bit wherein.Generator matrix is provided by following formula:

G ‾ = I ‾ H ‾ 1 T H ‾ 2 - T - - - (3)

Wherein $\underline{I}_{K*k}$It is the unit matrix of the systematic bits of generated codeword.Based on GCode word be

c ‾ = uG ‾ = u ‾ uH ‾ 1 T H &OverBar ; 2 - T = u ‾ p ‾ , - - - (4)

Wherein uBe information sequence (bits of user data vector).Code word cBy the systematic bits vector uWith the Parity Check Bits vector pConstitute.

According to second method, from code word cContinuous $\log_2 M$ bit is mapped directly in the M-ary constellation symbol.

Use first method, it is not optimum promptly using bit interleaver, and this is because compare with traditional BICM, does not need bit interleaver in the LDPC coded modulation.This is that the LDPC coded modulation is better than traditional code modulated advantage.

In addition, the shortcoming of first method is: it is not optimum that employed replacement rule among the IEEE 802.11a is applied to the LDPC coded modulation.This be because, in phase shift keying (PSK) and QAM, second displacement according to equation (2) alternately is mapped to higher

significant bit (MSB) and low significant bit (LSB) with adjacent coded-bit, this only is appropriate to nonsystematic code, as the convolution code of using according to IEEE 802.11a (seeing [3]).

Second method has similar shortcoming.When from code word cContinuous programming code bit when being mapped directly in PSK or the qam symbol, systematic bits and Parity Check Bits are placed among MSB and the LSB equably.

In [4], the result of study that provides shows that Turbo code is better than performance when the power that equates is assigned to whole codeword when system is assigned with different power levels with Parity Check Bits.For big block size, when systematic bits is protected well than Parity Check Bits, realize preferable performance, and this effect becomes stronger along with the increase of block size.The research of the unequal error protection in the LDPC sign indicating number is mainly concentrated on the structure (with reference to [5], [6]) of LDPC code word.Find that variable node degree (node degree) is big more usually, decoded average radiability is strong more.Yet this conclusion is only applicable to abnormal LDPC code.

An object of the present invention is to improve the performance of LDPC code modulating method.

This purpose realizes by the method, reflector, receiver and the computer program that have method according to the emission data of the feature of independent claims, receive data.

Summary of the invention

Provide a kind of emission to comprise the method for the data of a plurality of bits, wherein these data are mapped to a plurality of modulation symbols, each modulation symbol comprises at least one higher significant bit and at least one low significant bit, at least one Parity Check Bits generates at described a plurality of bits, and described a plurality of bit is mapped to the higher significant bit of a plurality of modulation symbols, and Parity Check Bits is mapped to the low significant bit of a plurality of modulation symbols.

A kind of reflector of the method according to above-mentioned emission data is provided in addition.

In addition, provide a kind of reception to comprise the method for the data of a plurality of bits, wherein a plurality of modulation symbols are received, each modulation symbol comprises at least one higher significant bit and at least one low significant bit, at least one Parity Check Bits that is used for described a plurality of bits is extracted from least one low significant bit of a plurality of modulation symbols, and described a plurality of bit is extracted from a plurality of modulation symbols.

A kind of receiver of the method according to above-mentioned reception data is provided in addition.

Description of drawings

Fig. 1 shows the communication system according to the embodiment of the invention.

Fig. 2 shows according to the pairing bit block of the modulation of the embodiment of the invention.

Fig. 3 shows the figure of the performance of the explanation embodiment of the invention.

Fig. 4 shows the figure of the performance of the explanation embodiment of the invention.

Embodiment

Illustrative ground, data bit (being systematic bits under LDPC (low-density checksum) coding situation) is mapped to the higher significant bit (MSB) of modulation symbol, and Parity Check Bits is mapped to the low significant bit (LSB) of modulation symbol.

It will be found that this unequal error protection (UEP) scheme performance is better than traditional method, particularly when it is applied to LDPC PSK or QAM (quadrature amplitude modulation).Performance gain is suitable with the gain of introducing by the frame sign that increases LDPC, sometimes even more remarkable than it, compares with using the present invention, and the frame sign of increase LDPC has increased the complexity and the memory cost of encoder.

With above-mentioned two kinds of art methods (method 1: use bit interleaver according to IEEE 802.11a; Method 2: the LDPC code word is mapped directly to the M-ary symbol) compares, use the present invention not sacrifice emission rate, and do not introduce additional wait time and extra process burden to reflector and receiver.Make data bit be mapped to MSB and Parity Check Bits is mapped to a plurality of data bits of LSB and the ordering of Parity Check Bits can be designed in advance and can be embedded in the generation of generator matrix and parity matrix, as following illustrate at LDPC.

The embodiment of the invention can embody from the subordinate claim.Effective to method, receiver and the computer program of reflector, reception data similarly around the embodiment that the method for emission data is described.

A plurality of modulation symbols can use at least one communication channel to send.In particular, the present invention is applicable to MIMO (multiple-input and multiple-output) system, for example MIMO OFDM (OFDM) system.

For example, at least one carrier signal is come modulated according to a plurality of modulation symbols.

In one embodiment, at least one Parity Check Bits generates according to the LDPC coding of a plurality of bits.

16QAM or 64QAM that data can according to quadrature amplitude modulation, for example use Gray to shine upon are mapped to a plurality of modulation symbols.In particular, the present invention can be applied to using all communication systems of LDPC coding QAM modulation.When using phase shift keying (PSK) for example 8PSK, 16PSK etc. and have higher significant bit and during other modulation scheme of low significant bit, also can use the present invention.

Can understand the present invention based on the following fact: find; even the LDPC sign indicating number has equal variable node degree; when highly being protected with Parity Check Bits or systematic bits is compared with the situation that Parity Check Bits is protected comparably; when systematic bits was highly protected, the LDPC sign indicating number provided more performance.

In the M-ary modulating system, the bit of distributing to MSB has better bit error rate (BER) performance than those bits of distributing to LSB, and perhaps of equal valuely, the bit among the MSB is than the higher SNR of the experience of the bit among the LSB.

Below with reference to accompanying drawing illustrative embodiment of the present invention is made an explanation.

Fig. 1 shows the communication system 100 according to the embodiment of the invention.

Communication system 100 comprises reflector 101 and receiver 102.

Reflector 101 comprises LDPC encoder 103 and M-ary modulator 104.The data for the treatment of to send to receiver 102 via communication channel 106 by reflector 101 are fed into LDPC encoder 103 with the form of first bit stream 105.103 pairs first bit streams of LDPC encoder 105 are carried out the LDPC coding, to generate LDPC coded bit stream 109.The LDPC coded-bit of LDPC coded bit stream 109 is combined into the code word that length is N, and is mapped in the M-ary constellation symbol by M-ary modulator 104.Used the Gray mapping in this embodiment, because it is better than anti-Gray mapping, collects and cut apart mapping (set partitioning mapping) etc.

Constellation symbol is launched into receiver via communication channel 106.

Receiver 102 comprises soft demodulator 107, and soft demodulator 107 is based on the bit generation bit metric of maximum-likelihood criterion at each emission, and the soft bit degree value that generates by this way is fed into iteration LDPC decoder 108.LDPC decoder 108 is used and long-pending message transmission (sum-product message-passing) algorithm comes to generate the reception data with the form of second bit stream 110.

Explain the LDPC coding of first bit stream 105 and the mapping that LDPC coded bit stream 109 arrives constellation symbol in more detail below.

LDPC encoder 103 is mapped to the code word that length is N with every K bit of first bit stream 105 cThis means LDPC encoder 103 enforcement (N, K) LDPC sign indicating numbers.This LDPC sign indicating number by its

parity check matrix H=[H 1 H 2] provide.Generator matrix is provided by following formula:

G ‾ = I ‾ H ‾ 1 T H ‾ 2 - T - - - (5)

Wherein I K * kIt is the unit matrix of the systematic bits of generated codeword.Based on GCode word cProvide by following formula:

c ‾ = uG ‾ = u ‾ uH ‾ 1 T H &OverBar ; 2 - T = u ‾ p ‾ , - - - (6)

Wherein uBe to be mapped to code word cThe vector of K bit, be also referred to as the systematic bits vector.Code word cBy the systematic bits vector uWith the vector that is called as the Parity Check Bits vector pConstitute.

For the code word c of the part of LDPC coded bit stream 109 is mapped to Individual constellation symbol.This mapping is corresponding to the layout of the bit block form shown in Fig. 2.

Fig. 2 shows according to the pairing bit block 200 of the modulation of the embodiment of the invention.

The height 201 of piece 200, promptly the number of the row 202 of piece 200 is M is a modulation order, for example for 8PSK, and M=8; For 16QAM, M=16; For 64QAM, M=64.

The width 203 of piece 200, promptly the number of the row 204 of piece 200 is N/s.This means that piece 200 comprises the N bit.Therefore, piece 200 can hold code word just cBit.

Each odd column of piece 200, promptly first row, the 3rd row etc. are corresponding to I (homophase) component.Each even column of piece 200, promptly secondary series, the 4th row etc. are corresponding to Q (quadrature) component.Per two row of piece 200 are mapped to a constellation symbol.For example, first row and the secondary series of piece 200 are mapped to a constellation symbol, wherein the first row assigned I component and secondary series appointment Q component.

In another embodiment, each row in the row of the first half of piece 200, promptly first row, secondary series etc. are up to Row are corresponding to

I (homophase) component.Each row in back half the row of piece 200, promptly Row, the Row wait up to the Row are corresponding to Q (quadrature) component.For example, first of piece 200 row and the Row are mapped to a constellation symbol, wherein the first row assigned I component and the Row are specified Q component.

Described row are mapped to constellation symbol as follows, and promptly the top row of piece 200 is corresponding to the higher significant bit (MSB) of I component and Q component, and the low row of piece 200 is corresponding to the low significant bit (LSB) of I component and Q component.For example, under the situation of 16QAM, first is listed in the higher significant bit that holds I component in first row, and holds the low significant bit of I component in second row.

Code word cBe written to piece 200 with line direction.This means code word cFirst bit corresponding to piece 200 first row first the row in bit, code word cSecond bit corresponding to secondary series first the row in bit ..., code word cThe N/s+1 bit corresponding to first row second the row in bit, or the like.

Suppose that the bit that is input to M-ary modulator 104 is written to piece 200 with column direction defaultly, promptly first Write first row from top to bottom, second Write secondary series from top to bottom, or the like, code word write with line direction (rather than column direction) cCorresponding to the code word that provides by following formula cTo code word Displacement:

c ‾ ~ = u G ~ ‾ - - - (7)

Wherein modified generator matrix Provide by following formula:

G ‾ ~ = GP ‾ . - - - (8)

Matrix P=[$p_{i,j}$] be according to code word cThe permutation matrix of displacement.It is provided by following formula:

When i=θ and j=μ, $P_{i,j}$=1; Otherwise, $P_{i,j}$=0 (9)

θ=0,1 wherein ..., N-1, and μ=N/s * (θ mod s)+floor (θ/s), θ=0,1 ..., N-1.

This means code word Be written to piece 200 with column direction, this is equivalent to line direction and writes code word c

This means in the LDPC sign indicating number code word cDisplacement be equivalent to generator matrix G be revised as Wherein PIt is permutation matrix.Corresponding parity matrix is With Can be designed in advance and be stored in reflector and the receiver respectively.Therefore, the displacement design (or of equal value, matrix POptimization) do not introduce any stand-by period and do not increase the processing complexity.

Code word cSystematic bits therefore be written to MSB and highly protected.This has caused high-performance.

In order to verify that above-mentioned UEP (unequal error protection) replacement scenario has caused high-performance, under the situation of 16QAM on the independent Rayleigh fading channel and 64QAM, with its performance and the code modulated two kinds of art methods of above-mentioned LDPC (method 1: use bit interleaver according to IEEE 802.11a; Method 2:LDPC code word is to the direct mapping of M-ary symbol) performance compare.

But a kind of high efficient coding abnormal LDPC code that practice significance is arranged, promptly partial differential (PD) LDPC sign indicating number is applied to emulation.Used 1/2 speed PD-LDPC (2304,1152).In this case, H_2Be low biconjugate angular moment battle array (low-dual-diagonal matrix), H_1Have row weight 6, and with progressive edge-growth (progressive edge-growth) (PEG) method design.

Utilize the 16QAM of Gray mapping and the constellation of 64QAM to follow IEEE 802.11a (seeing [3]).At the receiver place, the bit metric that generates based on maximum-likelihood criterion is passed in the iteration LDPC decoder of application and long-pending message pass-algorithm.The maximum number of iteration is set to 40.Fig. 3 shows the result.

Fig. 3 shows the figure of the performance of the explanation embodiment of the invention.

First Figure 30 1 shows bit error rate (BER) and frame error ratio (FER) performance at 64QAM.Second Figure 30 2 shows bit error rate (BER) and frame error ratio (FER) performance at 16QAM.

In Figure 30 1,302, BER and FER promptly increase towards the direction of corresponding y axle 303 from bottom to top.The ratio of code check and codeword size promptly increases towards the direction of respective x axle 304 from left to right, and provides with dB.

From table 1 at the performance gain that surpasses two kinds of art methods of advising as can be seen in the summary of 16QAM and 64QAM that displacement is realized.In 16QAM, method 1 has similar performance with method 2, and in 64QAM, method 2 performances are better than method 1.

Table 1

In addition, at 2 * 2 space division multiplexings (SM) MIMO-OFDM (multi-I/O OFDM) system that is HT WLAN (high throughput wireless local area network) system design, provided advise the comparison of displacement and the performance of method 2.Owing to can find out that in the above among the 64QAM on the Rayleigh fading channel, method 2 performances are better than method 1, the displacement of advising is only compared with method 2.

A kind of IEEE 802.11n channel model of reality is used to check, and receiver is considered actual frequency offset estimation, phase compensation and channel estimation (comprising the SNR estimation).Provided simulation parameter in the table 2, the result has been shown among Fig. 4.

Table 2

Fig. 4 shows the figure of the performance of the explanation embodiment of the invention.

First Figure 40 1 shows at the bit error rate (BER) of channel B and frame error ratio (FER) performance.Second Figure 40 2 shows bit error rate (BER) and frame error ratio (FER) performance at channel E.

To top similar, in Figure 40 1,402, BER and FER promptly increase towards the direction of corresponding y axle 403 from bottom to top.The ratio of code check and codeword size promptly increases towards the direction of respective x axle 404 from left to right, and provides with dB.

At channel B and channel E, the displacement of advising surpasses the BER gain of method 2 at BER=10^{-4}Be respectively 1.5dB and 1.1dB, and at channel B and channel E, PER gain at PER=10^{-2}Be respectively 0.46dB and 0.67dB.

Emulation further illustrates, the UEP that the advises mapping PER performance gain that surpasses directly mapping (method 2) increase with the increase of the row weight of LDPC.Yet higher row weight does not always cause PER performance preferably, exists one to trade off here.Increase the row weight and cause bigger free distance.On the other hand, it also increases short possibility and quantity of enclosing (short girth).

By emulation, it can also be seen that the LDPC code word is distributed to the difference in the channel with unequal SNR and caused performance inequality.For example, under the situation of two channels with different SNR, useful is, systematic bits is highly protected, and is about to all systematic bits and is assigned in the channel with higher SNR and with all Parity Check Bits and is assigned in the channel with low SNR.

In this document, quoted following publication:

[1]R.G.Gallager, Low-Density?Parity-Check?Codes.Cambridge, MA : MIT?Press, 1963

[2]J.Hou, P.H.Siegel, L.B.Milstein, and?H.D.Pfister, "Capacity-approaching?bandwidth-efficient?coded?modulation?schemes?based?on?low-density?parity-check?codes, "IEEE?Trans.Inform.Theory, vol.49, no.9, pp.2141-2155 , Sept.2003.

[3]IEEE?802.11a?wireless?LAN?medium?access?control?and?physical?layer?specifications : high-speed?physical?layer?in?the?5GHz?band.IEEE?std?802.11a-1999, Sep.1999

[4]A.H.S.Mohammadi, A.K.Khandani, "On?unequal?error?protection?on?turbo-encoder?output?bits, "Electronics?Lett., vol.33, no.4, pp.273-274, Feb.1997

[5]V.Kumar, O.Milenkovic,
"On?unequal?error?protection?LDPC?codes?based?on?Plotkin-type?Construction", IEEE?GlobeCom?2004, Dallas, TX, Nov.2004

[6]X.Tang, M.-S.Alouini, A.J.Goldsmith,
"Effect?of?channel?estimation?error?on?M-QAM?BER?performance?in?Rayleigh?fading", IEEE?Trans.Commun., vol.47, no.12, pp.1856-1864, Dec.1999

Claims (12)

Hide Dependent

1. an emission comprises the method for the data of a plurality of data bits, wherein:
-described data map is arrived a plurality of modulation symbols, each modulation symbol comprises at least one higher significant bit and at least one low significant bit;
-generate at least one Parity Check Bits at described a plurality of data bits;
-generated codeword, wherein said code word comprise described at least one Parity Check Bits and the described a plurality of data bit that is generated;
-described a plurality of data bits are mapped to the higher significant bit of described a plurality of modulation symbols, and described at least one Parity Check Bits is mapped at least one low significant bit of described a plurality of modulation symbols, wherein:
Described code word is with the arranged in form of bit block;
Described bit block comprises at least two bit-rows; And
Described code word is written to described bit block with line direction.
2. method according to claim 1 wherein uses at least one communication channel to launch described a plurality of modulation symbol.
3. method according to claim 1 also comprises according to described a plurality of modulation symbols and modulates at least one carrier signal.
4. method according to claim 2 also comprises according to described a plurality of modulation symbols and modulates at least one carrier signal.
5. method according to claim 1, wherein the LDPC coding according to described a plurality of data bits generates described at least one parity bits.
6. method according to claim 2, wherein the LDPC coding according to described a plurality of data bits generates described at least one parity bits.

7. method according to claim 3, wherein the LDPC coding according to described a plurality of data bits generates described at least one parity bits.
8. method according to claim 4, wherein the LDPC coding according to described a plurality of data bits generates described at least one parity bits.
9. according to any one described method in the claim 1 to 8, wherein described data map is arrived described a plurality of modulation symbols according to quadrature amplitude modulation.
10. reflector of launching data, described data comprise a plurality of data bits, wherein:
-described reflector comprises the map unit that is suitable for described data map is arrived a plurality of modulation symbols, and each modulation symbol comprises at least one higher significant bit and at least one low significant bit;
-described reflector comprises and is suitable for generating at least one Parity Check Bits and being suitable for generating first generation unit of the code word that comprises described at least one Parity Check Bits of being generated and described a plurality of data bits at described a plurality of data bits; And
-described map unit is suitable for described a plurality of data bits are mapped to the higher significant bit of described a plurality of modulation symbols, and is suitable for described at least one Parity Check Bits is mapped at least one low significant bit of described a plurality of modulation symbols, wherein:
Described code word is with the arranged in form of bit block;
Described bit block comprises at least two bit-rows; And
Described code word is written to described bit block with line direction.
11. a reception comprises the method for the data of a plurality of data bits, wherein:
-receiving a plurality of modulation symbols, each modulation symbol comprises at least one higher significant bit and at least one low significant bit;
-from least one low significant bit of described a plurality of modulation symbols, extract at least one Parity Check Bits that is used for described a plurality of data bits;
-described a plurality of the data bits of extraction from described a plurality of modulation symbols;
-generated codeword, wherein said code word comprise described at least one Parity Check Bits and the described a plurality of data bit that is extracted, and wherein:
Described code word is with the arranged in form of bit block;

Described bit block comprises at least two bit-rows; And
Described code word reads from described bit block with line direction subsequently.
12. a receiver that receives data, described data comprise a plurality of data bits, and wherein said receiver comprises:
-receiving element is suitable for receiving a plurality of modulation symbols, and each modulation symbol comprises at least one higher significant bit and at least one low significant bit;
-the first extraction unit is suitable for extracting at least one Parity Check Bits that is used for described a plurality of data bits from least one low significant bit of described a plurality of modulation symbols;
-the second extraction unit is suitable for extracting described a plurality of data bits from described a plurality of modulation symbols;
-maker unit is suitable for generated codeword, and wherein said code word comprises described at least one Parity Check Bits and the described a plurality of data bit that is extracted, and wherein:
Described code word is with the arranged in form of bit block;
Described bit block comprises at least two bit-rows; And
Described code word reads from described bit block with line direction subsequently.

Patent Citations (17)

Publication numberPriority datePublication dateAssigneeTitle
US20040001555A1 *2002-06-262004-01-01Arnauld TaffinMethod and apparatus for producing and processing sequences of modulation symbols
CN1492598A *2002-07-262004-04-28休斯电子公司Satellite communication system using low density parity check code
US6795947B1 *1999-10-072004-09-21The Regents Of The University Of CaliforniaParity check outer code and runlength constrained outer code usable with parity bits
CN1533067A *2003-03-182004-09-29华为技术有限公司Lattice code modulation method and multiple users receiver
US20040258177A1 *2003-06-132004-12-23Ba-Zhong ShenMulti-dimensional space Gray code maps for multi-dimensional phase modulation as applied to LDPC (Low Density Parity Check) coded modulation
Family To Family Citations
GB2180127B *1985-09-041989-08-23Philips Electronic AssociatedMethod of data communication

JPH05182368A *1992-01-071993-07-23Pioneer Electron CorpDetector for presence of information data in recording medium playing device
GB9421552D0 *1994-10-261994-12-14Philips Electronics Uk LtdImprovements in or relating to data communication
US6031874A *1997-09-262000-02-29Ericsson Inc.Unequal error protection in coded modulation schemes
JP2002064470A2000-08-222002-02-28Mega Chips CorpError control method
US7324472B2 *2001-10-152008-01-29Siemens AktiengesellschaftTransmission method
JP3808769B2 *2001-12-272006-08-16三菱電機株式会社LDPC code check matrix generation method
US7203887B2 *2002-07-032007-04-10The Directtv Group, Inc.Method and system for routing in low density parity check (LDPC) decoders
US7120849B2 *2002-11-052006-10-10Samsung Electronics Co., Ltd.Data storage medium having link zone and apparatus and method for recording/reproducing data on/from the data storage medium
US20040171359A1 *2003-02-282004-09-02Olav TirkkonenPower allocation in a communication system
US7457368B22003-06-052008-11-25Brooktree Broadband Holding, Inc.Multilevel channel coding in ADSL
US8527855B2 *2004-08-162013-09-03Koninklijke Philips N.V.Interleaving and parsing for MIMO-OFDM systems
* Cited by examiner, † Cited by third party

Non-Patent Citations (1)

Title
US 20040001555 A1,全文.
* Cited by examiner, † Cited by third party

Cited By (28)

Publication numberPriority datePublication dateAssigneeTitle
Family To Family Citations
US20090138785A1 *2006-03-172009-05-28Mitsubishi Electric CorporationCommunication device, decoding device, information transmission method, and decoding method

AU2007233563B2 *2006-04-032011-07-14National Ict Australia LimitedChannel estimation for rapid dispersive fading channels
US20070271493A1 *2006-05-162007-11-22Samsung Electronics Co., Ltd.Unequal error protection method and apparatus using the same
KR100987692B1 *2006-05-202010-10-13삼성전자주식회사Apparatus and method for transmitting/receiving signal in a communication system
US8194750B2 *2006-10-162012-06-05Samsung Electronics Co., Ltd.System and method for digital communication having a circulant bit interleaver for equal error protection (EEP) and unequal error protection (UEP)
JP5089339B2 *2006-11-022012-12-05パナソニック株式会社Transmission method, transmission apparatus, and reception method
US8005166B2 *2006-11-142011-08-23Samsung Electronics Co., Ltd.System for digital television broadcasting using modified 2/3 trellis coding
CN101188435B2007-01-082011-03-16中兴通讯股份有限公司A bit interleaving device and its method
US20100122138A1 *2007-04-172010-05-13Panasonic CorporationRadio Communication Device and Radio Communication Method
JP2009055228A *2007-08-242009-03-12Sony CorpWireless communication system, wireless communication apparatus, and wireless communication method
US8386903B2 *2007-10-312013-02-26Futurewei Technologies, Inc.Bit reverse interleaving methods for QAM modulation in a wireless communication system
TWI538415B *2007-11-262016-06-11Sony CorpData processing device and data processing method
KR101623468B1 *2008-03-032016-05-23라이 라디오텔레비지오네 이탈리아나 에스.페.아.Bit permutation patterns for ldpc coded modulation and qam constellations
ITTO20080472A1 *2008-06-162009-12-17Rai Radiotelevisione Italiana SpaMethod of processing digital signals and transmission and reception system that implements said method
US20090322961A1 *2008-06-272009-12-31Limberg Allen LeroySystem for digital television broadcasting using modified 2/3 trellis coding
JP5235629B2 *2008-11-282013-07-10株式会社日立製作所Encoding and modulation method and decoding method for wireless communication apparatus

US8806288B2 *2009-03-132014-08-12Sharp Laboratories Of America, Inc.Systems and methods for providing unequal error protection code design from probabilistically fixed composition codes
JP5445829B2 *2009-05-292014-03-19ソニー株式会社Receiving apparatus, receiving method, program, and receiving system
JP4810620B2 *2010-08-102011-11-09パナソニック株式会社Receiving apparatus and receiving method
JP5753678B2 *2010-11-252015-07-22シャープ株式会社wireless transmission device
CN102638431A *2011-02-102012-08-15富士通株式会社Bit dispensing equipment, transmitter, bit distribution method and power distribution method
CN103023603B *2011-09-202015-07-08澜起科技（上海）有限公司Method for realizing bit-interleaved coded modulation based on low-density parity-check (LDPC) matrix
US9094125B2 *2012-05-242015-07-28Nec Laboratories America, Inc.Multidimensional coded-modulation for high-speed optical transport over few-mode fibers
EP3493408B1 *2012-07-272020-06-17Sun Patent TrustTransmission method
US9722634B2 *2013-10-142017-08-01Nec CorporationMultiple component codes based generalized low-density parity-check codes for high-speed optical transport
CN108322285A *2017-01-162018-07-24华为技术有限公司Sending method, method of reseptance and the device of data
US10523400B22017-11-202019-12-31Nokia Technologies OyMulti-code probabilistic signal shaping using frequency-division multiplexing
US10091046B12017-11-202018-10-02Nokia Technologies OyJoint use of probabilistic signal shaping and forward error correction
* Cited by examiner, † Cited by third party, ‡ Family to family citation

Similar Documents

PublicationPublication DateTitle
US20180069614A12018-03-08System and Method for Mapping Symbols for MIMO Transmission
US10680752B22020-06-09Wireless communication base station apparatus, wireless communication terminal, communication method of a wireless

communication base station apparatus and communication method of a wireless communication terminal
US9634797B22017-04-25Method of transmitting a digital signal for a semi-orthogonal MS-marc system, and a corresponding program product and relay device
US8286062B22012-10-09Method and apparatus of encoding and decoding data using low density parity check code in a wireless communication system
JP2014042335A2014-03-06Method and apparatus for channel decoding in system using low-density parity-check code
Grimm et al.1998Further results on space-time coding for Rayleigh fading
US8190981B22012-05-29Apparatus and method for transmitting and receiving data in a communication system using low density parity check codes
KR100567698B12006-04-05A satellite communication system utilizing low density parity check codes
CA2540263C2011-10-11Method and apparatus for space-time coding using lifting low density parity check codes in a wireless communication system
US8683288B22014-03-25Variable modulation with LDPC (low density parity check) coding
US8392793B22013-03-05Code design and implementation improvements for low density parity check codes for multiple-input multiple-output channels
US8144801B22012-03-27Satellite communication system utilizing low density parity check codes
US8464120B22013-06-11Method and system for data transmission in a multiple input multiple output (MIMO) system including unbalanced lifting of a parity check matrix prior to encoding input data streams
US7600180B22009-10-06Iterative metric updating when decoding LDPC (low density parity check) coded signals and LDPC coded modulation signals
KR101102396B12012-01-05Method of matching codeword size and transmitter therefor in mobile communications system
EP2254250B12015-05-27Bit permutation patterns for LDPC coded modulation and 64QAM constellations
CA2572514C2017-07-11Advanced mimo interleaving
KR101321023B12013-10-23Multidimensional constellations for coded transmission
US8559552B22013-10-15Dual carrier modulation precoding
CN101072062B2011-05-25Apparatus and method for transmitting/receiving a signal in a communication system using multiple input multiple output scheme

CN101374032B2013-05-01Wireless communication system, wireless communication device, and wireless communication method
CN101778074B2012-10-03Multimedia broadcast wireless signal single-frequency network robust transmission method
US8527855B22013-09-03Interleaving and parsing for MIMO-OFDM systems
US8031793B22011-10-04Apparatus using concatenations of signal-space codes for jointly encoding across multiple transmit antennas, and employing coordinate interleaving
EP2025063B12018-09-26Method of encoding/decoding using low density check code matrix

Priority And Related Applications

Priority Applications (3)

ApplicationPriority dateFiling dateTitle
US64981805P2005-02-032005-02-03*US Provisional Application*
US60/649,8182005-02-03
PCT/SG2006/0000192005-02-032006-02-02Method for transmitting data, method for receiving data, transmitter, receiver, and computer program products

Legal Events

DateCodeTitleDescription
2008-02-27PB01Publication
2008-02-27C06Publication
2008-04-23SE01Entry into force of request for substantive examination
2008-04-23C10Entry into substantive examination
2010-10-06GR01Patent grant
2010-10-06C14Grant of patent or utility model

(19)中华人民共和国国家知识产权局

(12)发明专利

(10)授权公告号 CN 101133558 B
(45)授权公告日 2010.10.06

(21)申请号 200680006900.0
(22)申请日 2006.02.02
(30)优先权数据
60/649,818 2005.02.03 US
(85)PCT申请进入国家阶段日
2007.09.03
(86)PCT申请的申请数据
PCT/SG2006/000019 2006.02.02
(87)PCT申请的公布数据
WO2006/083233 EN 2006.08.10
(73)专利权人 新加坡科技研究局
地址 新加坡新加坡城
(72)发明人 李援 孙素梅 吴岩
(74)专利代理机构 北京集佳知识产权代理有限公司 11227
代理人 杨生平 潘士霖
(51)Int.Cl.
H03M 13/09(2006.01)
H04L 1/00(2006.01)
H03M 13/00(2006.01)
(56)对比文件
CN 1533067 A,2004.09.29, 全文.
US 20040258177 A1,2004.12.23, 全文.
CN 1492598 A,2004.04.28, 全文.
US 20040001555 A1, 全文.
US 6795947 B1,2004.09.21, 全文.
审查员 孟宪超

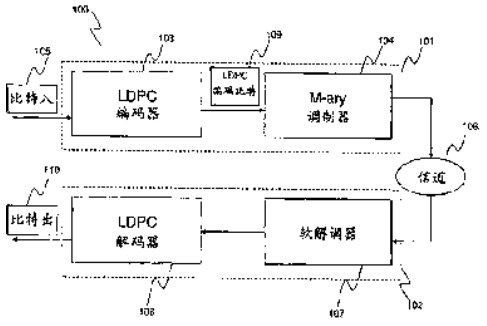
权利要求书 2 页 说明书 8 页 附图 4 页

(54)发明名称

发射数据的方法、接收数据的方法、发射器和接收器

(57)摘要

描述了一种发射包括多个比特的数据的方法，其中该数据被映射到多个调制符号，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特，至少一个奇偶校验比特针对所述多个比特而生成，并且所述多个比特被映射到多个调制符号的较高有效比特，而至少一个奇偶校验比特被映射到多个调制符号的至少一个较低有效比特。



CN 101133558 B

1. 一种发射包括多个数据比特的数据的方法，其中：

-将所述数据映射到多个调制符号，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特；

-针对所述多个数据比特生成至少一个奇偶校验比特；

-生成码字，其中所述码字包括所生成的所述至少一个奇偶校验比特和所述多个数据比特；

-所述多个数据比特被映射到所述多个调制符号的较高有效比特，并且所述至少一个奇偶校验比特被映射到所述多个调制符号的至少一个较低有效比特，其中：

所述码字以比特块的形式布置；

所述比特块包括至少两个比特行；并且

所述码字以行方向被写到所述比特块。

2. 根据权利要求 1 所述的方法，其中使用至少一个通信信道来发射所述多个调制符号。

3. 根据权利要求 1 所述的方法，还包括根据所述多个调制符号来调制至少一个载波信号。

4. 根据权利要求 2 所述的方法，还包括根据所述多个调制符号来调制至少一个载波信号。

5. 根据权利要求 1 所述的方法，其中根据所述多个数据比特的 LDPC 编码来生成所述至少一个奇偶检验比特。

6. 根据权利要求 2 所述的方法，其中根据所述多个数据比特的 LDPC 编码来生成所述至少一个奇偶检验比特。

7. 根据权利要求 3 所述的方法，其中根据所述多个数据比特的 LDPC 编码来生成所述至少一个奇偶检验比特。

8. 根据权利要求 4 所述的方法，其中根据所述多个数据比特的 LDPC 编码来生成所述至少一个奇偶检验比特。

9. 根据权利要求 1 至 8 中任何一项所述的方法，其中根据正交幅度调制来将所述数据映射到所述多个调制符号。

10. 一种发射数据的发射器，所述数据包括多个数据比特，其中：

-所述发射器包括适于将所述数据映射到多个调制符号的映射单元，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特；

-所述发射器包括适于针对所述多个数据比特生成至少一个奇偶校验比特并且适于生成包括所生成的所述至少一个奇偶校验比特和所述多个数据比特的码字的第一生成单元；并且

-所述映射单元适于将所述多个数据比特映射到所述多个调制符号的较高有效比特，并且适于将所述至少一个奇偶校验比特映射到所述多个调制符号的至少一个较低有效比特，其中：

所述码字以比特块的形式布置；

所述比特块包括至少两个比特行；并且

所述码字以行方向被写到所述比特块。

11. 一种接收包括多个数据比特的数据的方法，其中：

- 接收多个调制符号，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特；

- 从所述多个调制符号的至少一个较低有效比特中提取用于所述多个数据比特的至少一个奇偶校验比特；

- 从所述多个调制符号中提取所述多个数据比特；

- 生成码字，其中所述码字包括所提取的所述至少一个奇偶校验比特和所述多个数据比特，并且其中：

所述码字以比特块的形式布置；

所述比特块包括至少两个比特行；并且

所述码字随后以行方向从所述比特块中读取。

12. 一种接收数据的接收器，所述数据包括多个数据比特，其中所述接收器包括：

- 接收单元，适于接收多个调制符号，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特；

- 第一提取单元，适于从所述多个调制符号的至少一个较低有效比特中提取用于所述多个数据比特的至少一个奇偶校验比特；

- 第二提取单元，适于从所述多个调制符号中提取所述多个数据比特；

- 生成器单元，适于生成码字，其中所述码字包括所提取的所述至少一个奇偶校验比特和所述多个数据比特，并且其中：

所述码字以比特块的形式布置；

所述比特块包括至少两个比特行；并且

所述码字随后以行方向从所述比特块中读取。

发射数据的方法、接收数据的方法、发射器和接收器

技术领域

[0001] 本发明涉及一种发射数据的方法、接收数据的方法、发射器、接收器和计算机程序产品。

背景技术

[0002] 在通信系统中，使用检测并纠正传输期间发生的错误的方法来提高可靠性。一种可能性是使用[1]中所描述的低密度奇偶校验码(LDPC)，其是目前吸引了许多关注的容量逼近编码方案。与同样是容量逼近码的Turbo码相比较，LDPC码在大的块大小的情况下有更好的性能，尽管其解码复杂性小于Turbo码的解码复杂性。

[0003] LDPC的另一个商业优点是LDPC专利已经期满，因此，公司可以使用LDPC而不必支付知识产权的费用。

[0004] 在[2]中描述了LDPC编码器与M-ary(M元)调制器一起使用(称为LDPC编码调制)，其提供了高的功率和带宽效率。M-ary调制器例如根据Gray映射将每$\log_2 M$个LDPC编码比特映射到一个星座(constellation)符号。此映射不在传输中引入任何等待时间，这不同于比特交织编码调制(BICM)中的比特交织。通过LDPC码的最优设计，LDPC编码调制可以接近编码调制的容量极限。

[0005] LDPC码和LDPC编码调制已被广泛引入下一代无线通信系统，以便于以高数据速率进行可靠和高效的传输。与正交幅度调制(QAM)和MIMO-OFDM(多输入多输出-正交频分复用)相结合，LDPC在高吞吐量(HT)无线局域接入网(WLAN)中已表现出优于Turbo码。它还是在对IEEE 802.11n标准的提议中的任选HT模式的有竞争力的候选者。LDPC码还已被提议用于也需要高阶调制的宽带无线接入(BWA)的标准IEEE 802.16。还有许多关于将LDPC设计到超宽带(UWB)无线通信中的提议，其中有可能在QAM UWB系统中使用LDPC。

[0006] 在应用卷积码或Turbo码的传统比特交织编码调制(BICM)方案中，为了将编码比特置换到不同符号中的不同位置，在编码器和调制器之间的比特交织器是必需的。

[0007] 在IEEE802.11a标准中，比特交织器通过如下步骤处理两步置换：

[0008] 步骤1：

[0009] 将相邻的编码比特映射到非相邻的调制符号上，以避免由单个符号和信道相关性引入的突发错误。

[0010] 当N表示码字大小时，第一置换被定义为

[0011] $i = (N/d)(k \bmod d) + \mathrm{floor}(k/d), k = 0, 1, K, N-1, (1)$

[0012] 其中d是交织深度，floor(.)表示不超过该参数的最大整数。

[0013] 步骤2：

[0014] 将相邻的编码比特交替地映射到星座的较高有效比特(MSB)和较低有效比特(LSB)，以避免具有低可靠性的LSB的长期运转。

[0015] 第二置换被定义为

[0016] $j = s \times \mathrm{floor}(i/s) + (i + N - \mathrm{floor}(d \times i/N)) \bmod s,$

[0017] $i = 0, 1, K, N-1$ (2)

[0018] 其中 $s = \max(\frac{\log_2 M}{2}, 1)$，由调制阶 M 确定。

[0019] 下面描述两种供 LDPC 编码调制使用的方法。

[0020] 第一种方法是使用如上所述根据 802.11a 的比特交织器来置换 LDPC 编码比特，参考 [2]。

[0021] 第二种方法是将 LDPC 码字直接映射到 M-ary 符号中。这在下面解释。

[0022] (N,K) LDPC 码由其奇偶校验矩阵 $H = [H_1 \ H_2]$ 给出，其中 K 和 N 分别是（待被编码的）信息比特的数目和 LDPC 编码比特的数目。生成矩阵由下式给出：

[0023]

$$G = [I \ H_1^T H_2^{-T}] \quad (3)$$

[0024] 其中 $I_{k\times k}$ 是生成码字的系统比特的单位矩阵。基于 G 的码字是

[0025]

$$c = uG = [u \ uH_1^T H_2^{-T}] = [u \ p], \quad (4)$$

[0026] 其中 u 是信息序列（用户数据比特向量）。码字 c 由系统比特向量 u 和奇偶校验比特向量 p 构成。

[0027] 根据第二种方法，来自码字 c 的连续 $\log_2 M$ 个比特被直接映射到 M-ary 星座符号中。

[0028] 使用第一种方法，即使用比特交织器不是最优的，这是因为与传统的 BICM 相比，在 LDPC 编码调制中不需要比特交织器。这是 LDPC 编码调制优于传统编码调制的优点。

[0029] 此外，第一种方法的缺点是：将 IEEE 802.11a 中所使用的置换规则应用于 LDPC 编码调制不是最优的。这是因为，在相移键控 (PSK) 和 QAM 中，根据等式 (2) 的第二置换将相邻的编码比特交替地映射到较高有效比特 (MSB) 和较低有效比特 (LSB)，这仅合适于非系统码，如根据 IEEE 802.11a 使用的卷积码（见 [3]）。

[0030] 第二种方法具有类似的缺点。当来自码字 c 的连续编码比特被直接映射到 PSK 或 QAM 符号中时，系统比特和奇偶校验比特被均匀地放置到 MSB 和 LSB 中。

[0031] 在 [4] 中，给出的研究结果显示，Turbo 码当系统和奇偶校验比特被分配不同的功率电平时比当相等的功率被分配给整个码字时性能更好。对于大的块大小，当系统比特比奇偶校验比特保护得好时实现较好的性能，并且此效应随着块大小的增大而变得更强。对 LDPC 码中的不等错误保护的研究主要集中于 LDPC 码字的构造（参考 [5]、[6]）。已发现，通常可变节点度 (node degree) 越大，解码后的平均透辐射能力越强。然而，这个结论仅适用于不规则 LDPC 码。

[0032] 本发明的一个目的是提高 LDPC 编码调制方法的性能。

[0033] 该目的通过具有根据独立权利要求的特征的发射数据的方法、接收数据的方法、发射器、接收器和计算机程序产品来实现。

发明内容

[0034] 提供了一种发射包括多个比特的数据的方法，其中该数据被映射到多个调制符

号，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特，至少一个奇偶校验比特针对所述多个比特而生成，并且所述多个比特被映射到多个调制符号的较高有效比特，而奇偶校验比特被映射到多个调制符号的较低有效比特。

[0035] 此外，提供了一种根据上述发射数据的方法的发射器。

[0036] 此外，提供了一种接收包括多个比特的数据的方法，其中多个调制符号被接收，每个调制符号包括至少一个较高有效比特和至少一个较低有效比特，用于所述多个比特的至少一个奇偶校验比特从多个调制符号的至少一个较低有效比特中被提取，并且所述多个比特从多个调制符号中被提取。

[0037] 此外，提供了一种根据上述接收数据的方法的接收器。

附图说明

[0038] 图 1 示出了根据本发明实施例的通信系统。

[0039] 图 2 示出了根据本发明实施例的调制所对应的比特块。

[0040] 图 3 示出了说明本发明实施例的性能的图。

[0041] 图 4 示出了说明本发明实施例的性能的图。

具体实施方式

[0042] 说明性地，数据比特（在 LDPC（低密度奇偶校验）编码情况下为系统比特）被映射到调制符号的较高有效比特 (MSB)，而奇偶校验比特被映射到调制符号的较低有效比特 (LSB)。

[0043] 下面将表明，这种不等错误保护 (UEP) 方案性能优于传统的方法，特别是当其应用于 LDPC PSK 或 QAM（正交幅度调制）时。性能增益与通过增大 LDPC 的帧大小而引入的增益相当，有时甚至比其更显著，与使用本发明相比，增大 LDPC 的帧大小增大了编码器和解码器的复杂性和存储器成本。

[0044] 与上述两种现有技术方法（方法 1：使用根据 IEEE 802.11a 的比特交织器；方法 2：将 LDPC 码字直接映射到 M-ary 符号）相比，使用本发明不牺牲发射速率，并且不向发射器和接收器引入额外等待时间和额外处理负担。使得数据比特被映射到 MSB 而奇偶校验比特被映射到 LSB 的多个数据比特和奇偶校验比特的排序可以被预设计并且可以被嵌入到生成矩阵和奇偶校验矩阵的生成中，如下面针对 LDPC 示出的那样。

[0045] 本发明实施例可体现自从属权利要求。围绕发射数据的方法描述的实施例类似地对发射器、接收数据的方法、接收器和计算机程序产品有效。

[0046] 多个调制符号可以使用至少一个通信信道来发送。特别而言，本发明适用于 MIMO（多输入多输出）系统，例如 MIMO OFDM（正交频分复用）系统。

[0047] 例如，至少一个载波信号根据多个调制符号来被调制。

[0048] 在一个实施例中，至少一个奇偶校验比特根据多个比特的 LDPC 编码来生成。

[0049] 数据可以根据正交幅度调制、例如使用 Gray 映射的 16QAM 或 64QAM 来被映射到多个调制符号。特别而言，本发明可以被应用于使用 LDPC 编码 QAM 调制的所有通信系统。当使用相移键控 (PSK) 例如 8PSK、16PSK 等以及存在有较高有效比特和较低有效比特的其它调制方案时，也可以应用本发明。

[0050] 可以基于如下事实来领会本发明：已发现，即使 LDPC 码具有相等的可变节点度，当与奇偶校验比特被高度保护或者系统比特和奇偶校验比特被同等地保护的情形相比，系统比特被高度保护时，LDPC 码提供更好的性能。

[0051] 在 M-ary 调制系统中，分配给 MSB 的比特比分配给 LSB 的那些比特具有更好的比特错误率 (BER) 性能，或者等价地，MSB 中的比特比 LSB 中的比特经历更高的 SNR。

[0052] 下面参考附图对本发明的说明性实施例进行解释。

[0053] 图 1 示出了根据本发明实施例的通信系统 100。

[0054] 通信系统 100 包括发射器 101 和接收器 102。

[0055] 发射器 101 包括 LDPC 编码器 103 和 M ary 调制器 104。待由发射器 101 经由通信信道 106 发送到接收器 102 的数据以第一比特流 105 的形式被供给到 LDPC 编码器 103。LDPC 编码器 103 对第一比特流 105 执行 LDPC 编码，以生成 LDPC 编码比特流 109。LDPC 编码比特流 109 的 LDPC 编码比特被组合成长度为 N 的码字，并通过 M-ary 调制器 104 被映射到 M-ary 星座符号中。在这个实施例中应用了 Gray 映射，因为它优于反 Gray 映射、集分割映射 (set partitioning mapping) 等。

[0056] 星座符号经由通信信道 106 被发射到接收器。

[0057] 接收器 102 包括软解调器 107，软解调器 107 基于最大似然准则针对每个发射的比特生成比特度量，且以这种方式生成的软比特度量值被供给到迭代 LDPC 解码器 108。LDPC 解码器 108 应用和积消息传递 (sum-product message-passing) 算法来以第二比特流 110 的形式生成接收数据。

[0058] 下面更详细地解释第一比特流 105 的 LDPC 编码以及 LDPC 编码比特流 109 到星座符号的映射。

[0059] LDPC 编码器 103 将第一比特流 105 的每 K 个比特映射到长度为 N 的码字 $\underline{c}$。这意味着 LDPC 编码器 103 实施 (N,K) LDPC 码。该 LDPC 码由其奇偶校验矩阵 $H = [\underline{H}_1 \ \underline{H}_2]$ 给出。生成矩阵由下式给出：

[0060]

$$\underline{G} = [\underline{I} \ \underline{H}_1^T \underline{H}_2^{-T}] \tag{5}$$

[0061] 其中 $\underline{I}_{k\times k}$ 是生成码字的系统比特的单位矩阵。基于 $\underline{G}$ 的码字 $\underline{c}$ 由下式给出：

[0062]

$$\underline{c} = \underline{u}\underline{G} = [\underline{u} \ \underline{u}\underline{H}_1^T \underline{H}_2^{-T}] = [\underline{u} \ \underline{p}], \tag{6}$$

[0063] 其中 $\underline{u}$ 是被映射到码字 $\underline{c}$ 的 K 个比特的向量，也被称为系统比特向量。码字 $\underline{c}$ 由系统比特向量 $\underline{u}$ 和被称为奇偶校验比特向量的向量 $\underline{p}$ 构成。

[0064] 为 LDPC 编码比特流 109 的一部分的码字 c 被映射到 $\frac{N}{\log_2 M}$ 个星座符号。此映射对应于图 2 中示出的比特块形式的布置。

[0065] 图 2 示出了根据本发明实施例的调制所对应的比特块 200。

[0066] 块 200 的高度 201，即块 200 的行 202 的数目为 $s = \max(\frac{\log_2 M}{2}, 1)$ M 是调制阶，例如对于 8PSK，M = 8；对于 16QAM，M = 16；对于 64QAM，M = 64。

[0067] 块 200 的宽度 203，即块 200 的列 204 的数目为 N/s。这意味着块 200 包括 N 比特。

因此，块 200 可以恰好容纳码字 $\underline{c}$ 的比特。

[0068] 块 200 的每个奇数列，即第一列、第三列等对应于 I（同相）分量。块 200 的每个偶数列，即第二列、第四列等对应于 Q（正交）分量。块 200 的每两列被映射到一个星座符号。例如，块 200 的第一列和第二列被映射到一个星座符号，其中第一列指定 I 分量而第二列指定 Q 分量。

[0069] 在另一个实施例中，块 200 的前一半的列中的每一列，即第一列、第二列等直到第 $\left(\frac{1}{2}\cdot\frac{N}{s}\right)$ 列对应于 I（同相）分量。块 200 的后一半的列中的每一列，即第 $\left(\frac{1}{2}\cdot\frac{N}{s}+1\right)$ 列、第 $\left(\frac{1}{2}\cdot\frac{N}{s}+2\right)$ 列等直到第 $\frac{N}{s}$ 列对应于 Q（正交）分量。例如，块 200 的第一列和第 $\left(\frac{1}{2}\cdot\frac{N}{s}+1\right)$ 列被映射到一个星座符号，其中第一列指定 I 分量而第 $\left(\frac{1}{2}\cdot\frac{N}{s}+1\right)$ 列指定 Q 分量。

[0070] 所述列以如下方式被映射到星座符号，即块 200 的上部行对应于 I 分量和 Q 分量的较高有效比特（MSB），而块 200 的较低行对应于 I 分量和 Q 分量的较低有效比特（LSB）。例如，在 16QAM 的情况下，第一列在第一行中容纳 I 分量的较高有效比特，而在第二行中容纳 I 分量的较低有效比特。

[0071] 码字 $\underline{c}$ 以行方向被写到块 200。这意味着码字 $\underline{c}$ 的第一比特对应于块 200 的第一列的第一行中的比特，码字 $\underline{c}$ 的第二比特对应于第二列的第一行中的比特，...，码字 $\underline{c}$ 的第 N/s+1 比特对应于第一列的第二行中的比特，等等。

[0072] 假设输入到 M-ary 调制器 104 的比特缺省地以列方向被写到块 200，即第一个 $\frac{\log_2 M}{2}$ 从上到下写入第一列，第二个 $\frac{\log_2 M}{2}$ 从上到下写入第二列，等等，以行方向（而不是列方向）写码字 $\underline{c}$ 对应于由下式给出的码字 $\underline{c}$ 到码字 $\underline{\tilde{c}}$ 的置换：

[0073]

$$\underline{\tilde{c}} = \underline{u}\tilde{G} \tag{7}$$

[0074] 其中经修改的生成矩阵 $\tilde{\underline{G}}$ 由下式给出：

[0075]

$$\tilde{\underline{G}} = \underline{G}\underline{P}. \tag{8}$$

[0076] 矩阵 $\underline{P} = [p_{i,j}]$ 是根据码字 $\underline{c}$ 的置换的置换矩阵。它由下式给出：

[0077] 当 $i = \theta$ 且 $j = \mu$ 时，$P_{i,j} = 1$；否则，$P_{i,j} = 0$ (9)

[0078] 其中 $\theta = 0,1,\ldots,N-1$，并且 $\mu = N/s \times (\theta \bmod s) + \mathrm{floor}(\theta/s)$，$\theta = 0,1,\ldots,N-1$。

[0079] 这意味着码字 $\underline{\tilde{c}}$ 以列方向被写到块 200，这等价于以行方向写码字 $\underline{c}$。

[0080] 这意味着在 LDPC 码中，码字 $\underline{c}$ 的置换等价于将生成矩阵 G 修改为 $\tilde{\underline{G}} = \underline{G}\underline{P}$，其中 $\underline{P}$ 是置换矩阵。对应的奇偶校验矩阵是 $\tilde{\underline{H}} = \underline{H}\underline{P}^T$。$\tilde{\underline{G}}$ 和 $\tilde{\underline{H}}$ 可以分别被预设计并存储在发射器和接收器中。因此，置换设计（或等价的，矩阵 $\underline{P}$ 的优化）不引入任何等待时间并且不增大处理复杂性。

[0081] 码字 $\underline{c}$ 的系统比特被写到 MSB 并因此被高度保护。这导致了高性能。

[0082] 为了验证上述 UEP（不等错误保护）置换方案导致了高性能，在独立 Rayleigh 衰落信道上的 16QAM 和 64QAM 的情况下，将其性能与上述 LDPC 编码调制的两种现有技术方法

(方法 1 :使用根据 IEEE 802.11a 的比特交织器 ;方法 2 :LDPC 码字到 M-ary 符号的直接映射)的性能进行比较。

[0083] 一种有实践意义的可高效编码不规则 LDPC 码,即偏微分 (PD)LDPC 码被应用于仿真。使用了 1/2 速率 PD-LDPC(2304,1152)。在这种情况下,H_2 是低双对角矩阵 (low-dual-diagonal matrix),H_1 具有列权重 6,并用渐进边增长 (progressive edge-growth)(PEG) 方法来设计。

[0084] 利用 Gray 映射的 16QAM 和 64QAM 的星座遵循 IEEE 802.11a(见 [3])。在接收器处,基于最大似然准则生成的比特度量被传递到应用和积消息传递算法的迭代 LDPC 解码器中。迭代的最大数目被设置为 40。图 3 示出了结果。

[0085] 图 3 示出了说明本发明实施例的性能的图。

[0086] 第一个图 301 示出了针对 64QAM 的比特错误率 (BER) 和帧错误率 (FER) 性能。第二个图 302 示出了针对 16QAM 的比特错误率 (BER) 和帧错误率 (FER) 性能。

[0087] 在图 301、302 中,BER 和 FER 从下向上,即朝相应的 y 轴 303 的方向增长。码率与码字大小的比率从左向右,即朝相应的 x 轴 304 的方向增长,并且以 dB 给出。

[0088] 从表 1 中针对 16QAM 和 64QAM 的总结中可以看出所建议置换所实现的超过两种现有技术方法的性能增益。在 16QAM 中,方法 1 和方法 2 具有相似的性能,而在 64QAM 中,方法 2 性能优于方法 1。

[0089]

增益 (dB)		16QAM	64QAM
UEP 对方法 1	BER@10^{-4}	0.24	0.62
	FER@10^{-2}	0.26	0.45
UEP 对方法 2	BER@10^{-4}	0.24	0.48
	FER@10^{-2}	0.26	0.57

[0090] 表 1

[0091] 此外,针对为 HT WLAN(高吞吐量无线局域网)系统设计的 2×2 空分复用 (SM) MIMO-OFDM(多输入多输出正交频分复用)系统,给出了所建议置换与方法 2 的性能的比较。由于在上面可看出,在 Rayleigh 衰落信道上的 64QAM 中,方法 2 性能优于方法 1,所以所建议置换仅与方法 2 相比较。

[0092] 一种实际的 IEEE 802.11n 信道模型被用于检查,并且接收器考虑实际频率偏移估算、相位补偿和信道估算(包括 SNR 估算)。表 2 中给出了仿真参数,图 4 中示出了结果。

[0093]

LDPC 码	列权重为{6, 2, 1}的1/2速率PD-LDPC（2304, 1152）；用于多个空间流的单个编码器。
调制	利用Gray映射的64QAM
OFDM 大小	128
数据子载波数目	96
空间流数目	2
多个天线	Tx: 2 Rx: 2
IEEE 802.11n信道模型	信道B，信道E
包大小	1012比特

[0094] 表 2

[0095] 图 4 示出了说明本发明实施例的性能的图。

[0096] 第一个图 401 示出了针对信道 B 的比特错误率 (BER) 和帧错误率 (FER) 性能。第二个图 402 示出了针对信道 E 的比特错误率 (BER) 和帧错误率 (FER) 性能。

[0097] 与上面相似，在图 401、402 中，BER 和 FER 从下向上，即朝相应的 y 轴 403 的方向增长。码率与码字大小的比率从左向右，即朝相应的 x 轴 404 的方向增长，并且以 dB 给出。

[0098] 针对信道B和信道E，所建议置换超过方法2的BER增益在BER $= 10^{-4}$ 分别是 1.5dB 和 1.1dB，且针对信道 B 和信道 E，PER 增益在 PER $= 10^{-2}$ 分别是 0.46dB 和 0.67dB。

[0099] 仿真进一步示出，所建议 UEP 映射超过直接映射（方法 2）的 PER 性能增益随 LDPC 的列权重的增大而增大。然而，较高列权重并不总是导致较好的 PER 性能，这里存在一个折衷。增大列权重导致较大的自由距离。另一方面，其还增大短周 (short girth) 的可能性和数量。

[0100] 通过仿真，还可以看出，LDPC 码字到具有不相等 SNR 的信道中的不同分配导致不相同的性能。例如，在具有不同 SNR 的两个信道的情况下，有益的是，系统比特被高度保护，即将所有系统比特分配到具有较高 SNR 的信道中并将所有奇偶校验比特分配到具有较低 SNR 的信道中。

[0101] 在此文献中，引用了以下出版物：

[0102] [1]R. G. Gallager, Low-Density Parity-Check Codes. Cambridge, MA :MIT Press, 1963

[0103] [2]J. Hou, P. H. Siegel, L. B. Milstein, and H. D. Pfister, " Capacity-approaching bandwidth-efficient coded modulation schemes based on low-density parity-check codes, " IEEE Trans. Inform. Theory, vol. 49, no. 9, pp. 2141-2155, Sept. 2003.

[0104] [3]IEEE 802.11a wireless LAN medium access control and physical layer specifications :high-speed physical layer in the 5GHz band. IEEE std 802.11a-1999, Sep. 1999

[0105] [4]A. H. S. Mohammadi, A. K. Khandani, " On unequal error protection on turbo-encoder output bits," Electronics Lett., vol. 33, no. 4, pp. 273-274, Feb. 1997

[0106] [5]V. Kumar, O. Milenkovic," On unequal error protection LDPC codes based on Plotkin-type Construction" , IEEE GlobeCom 2004, Dallas, TX, Nov. 2004

[0107] [6]X. Tang, M. -S. Alouini, A. J. Goldsmith, " Effect of channel estimation error on M-QAM BER performance in Rayleigh fading" , IEEE Trans. Commun., vol. 47, no. 12, pp. 1856-1864, Dec. 1999

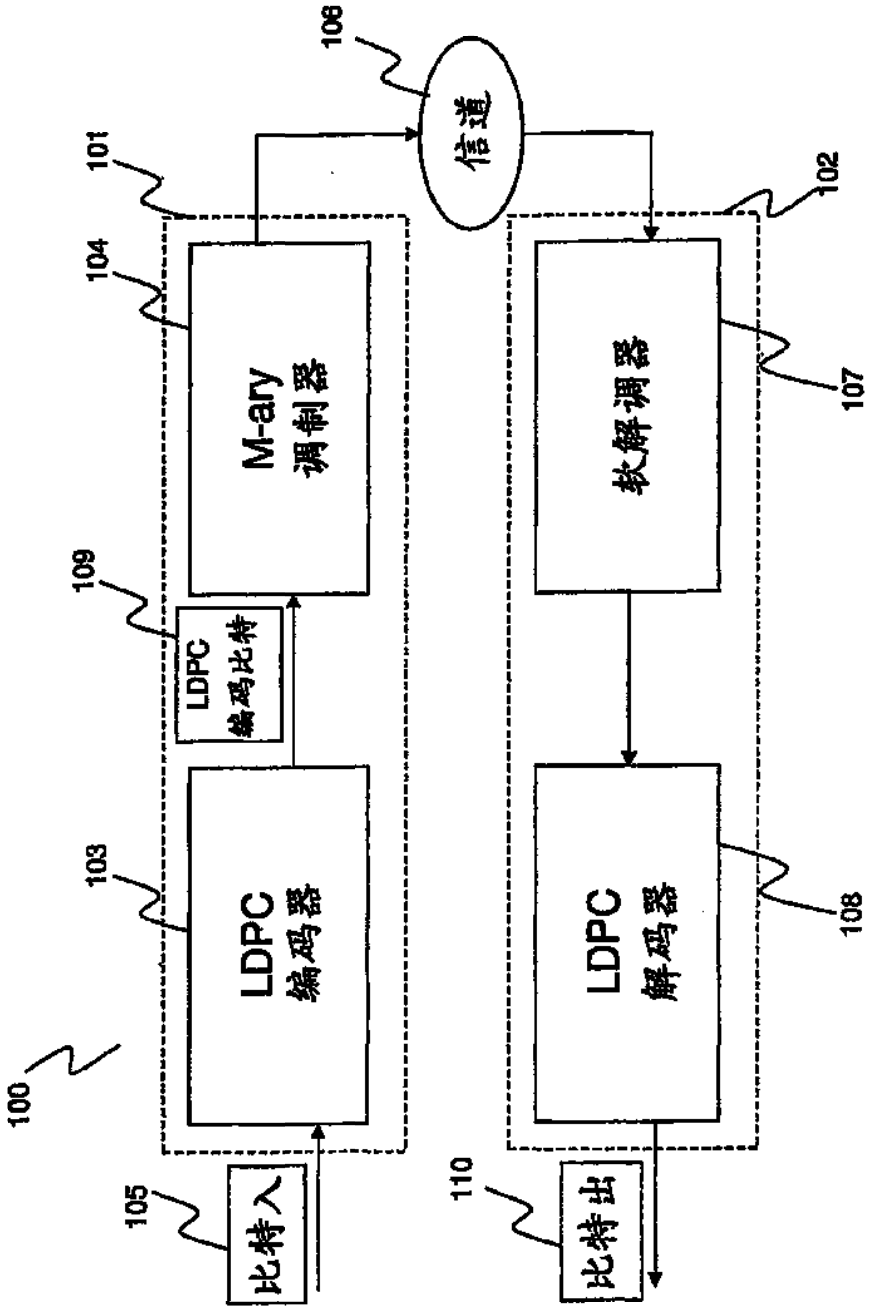

106
信道
101
102
104
M-ary
调制器
软解调器
107
109
LDPC
编码比特
103
LDPC
编码器
LDPC
解码器
108
100
105
比特入
110
比特出

图 1

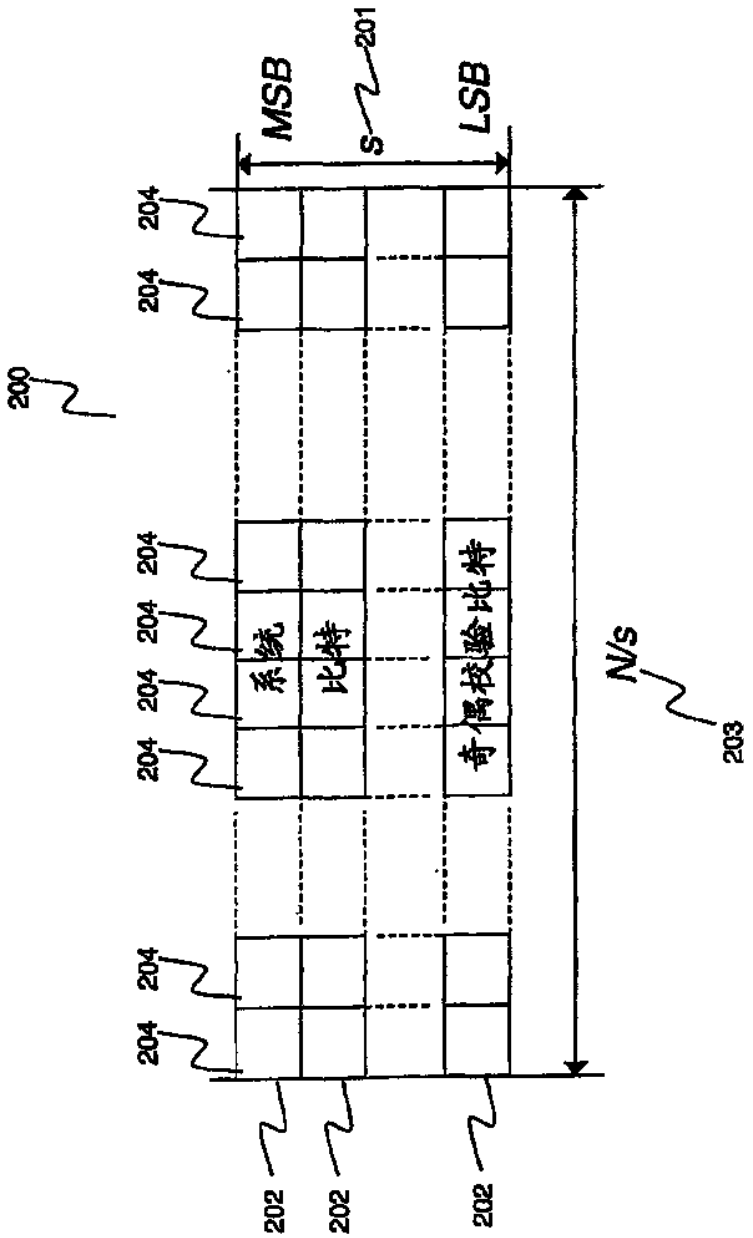

MSB
S
201
LSB
200
204
204
204
204
204
204
204
204
系统比特
奇偶校验比特
N/s
203
202
202
202

图 2

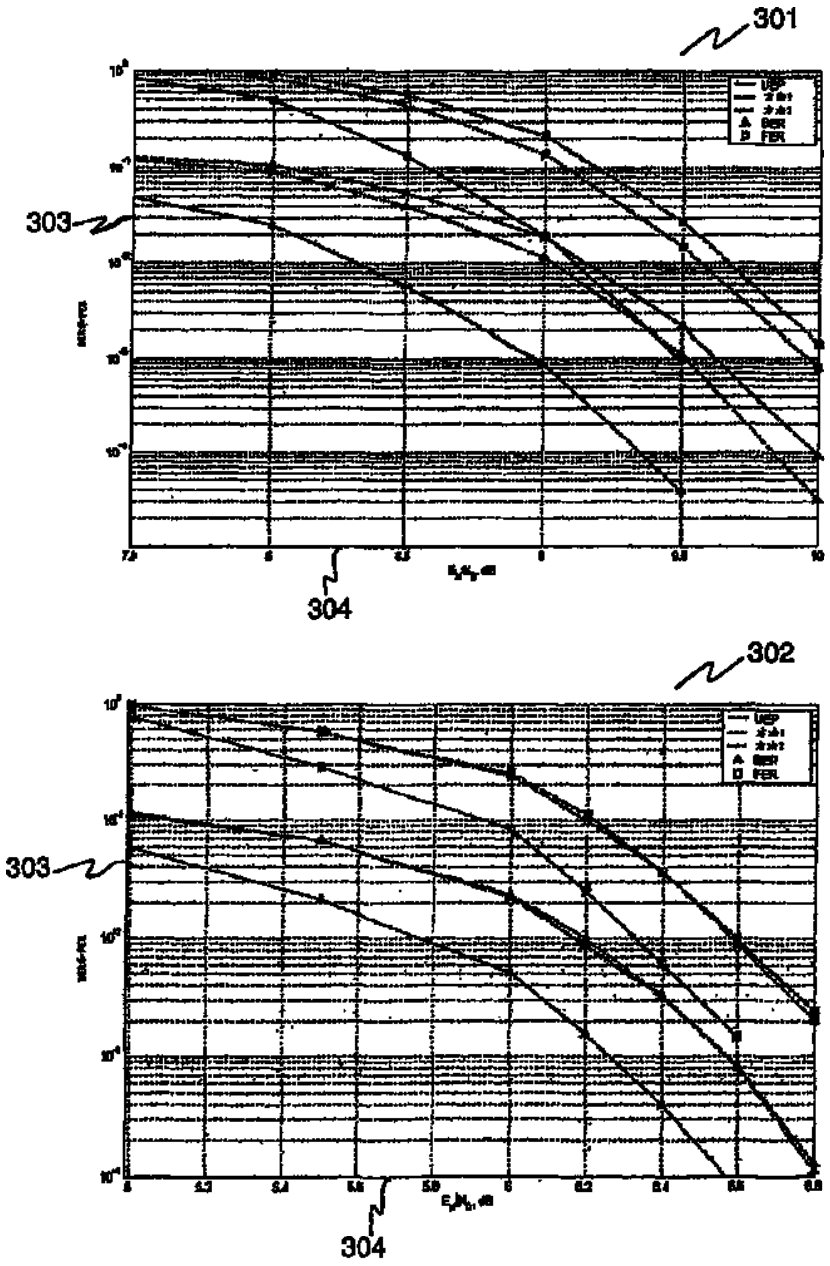


图 3

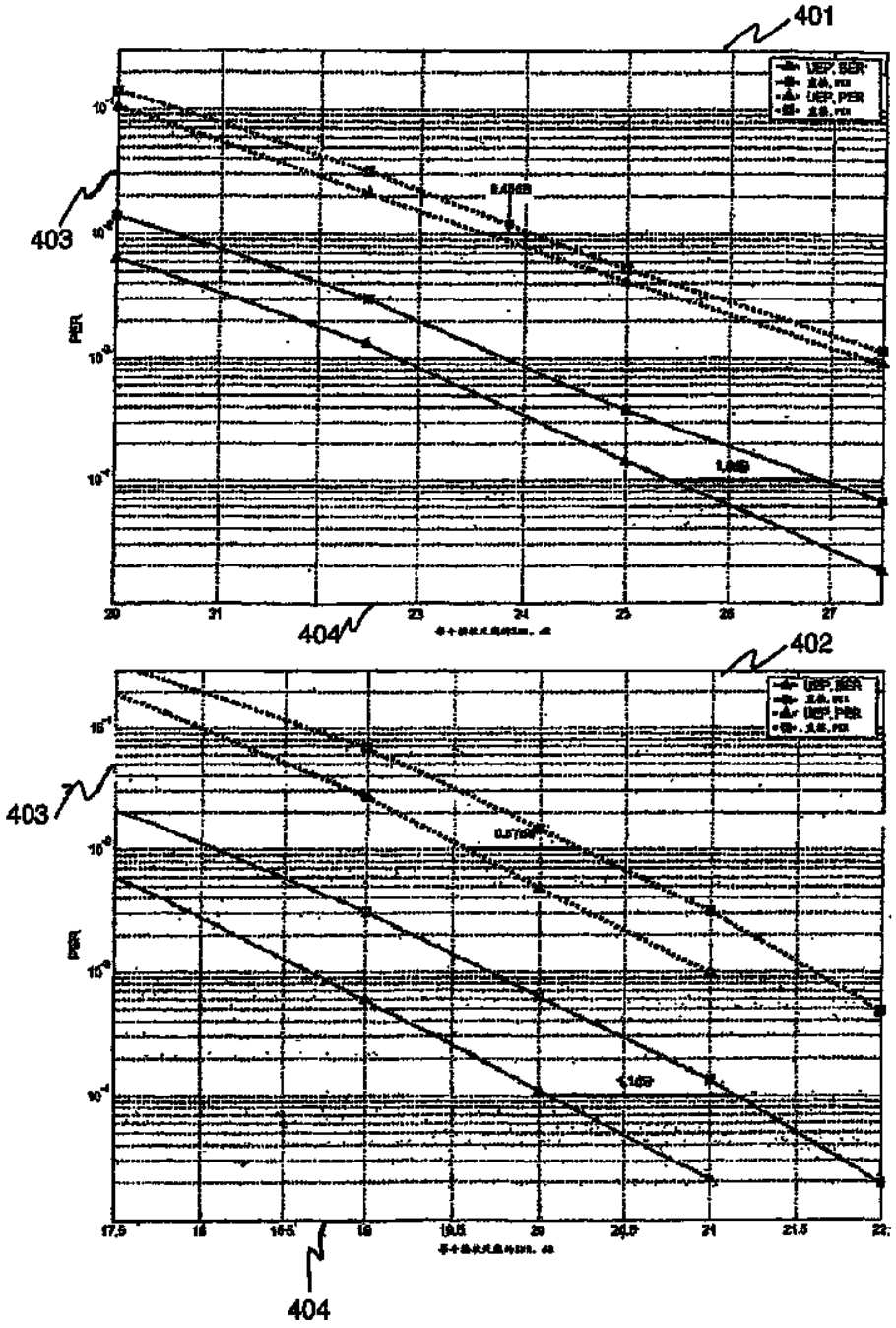


图 4

RU2428796C2

Methods and devices to increase efficiency and to make it possible to quickly decode transfers with multiple code units

Abstract

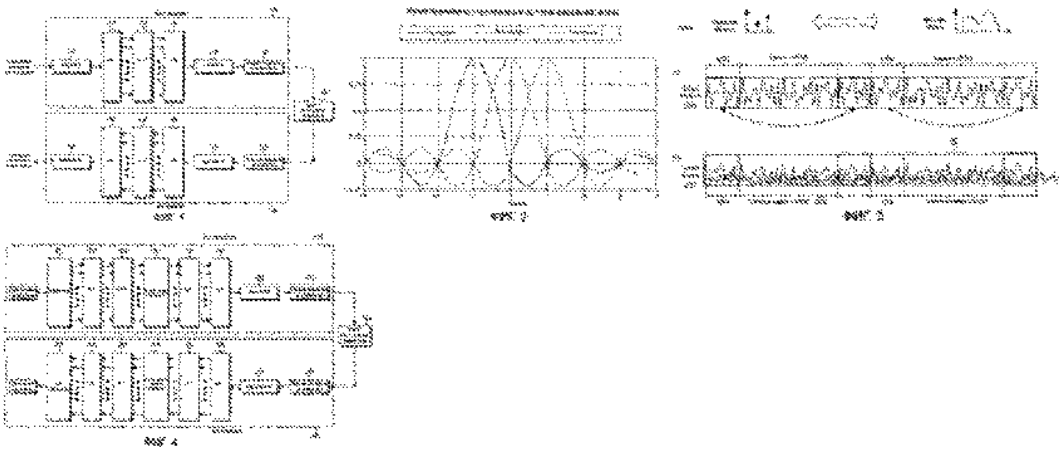
FIELD: information technologies.

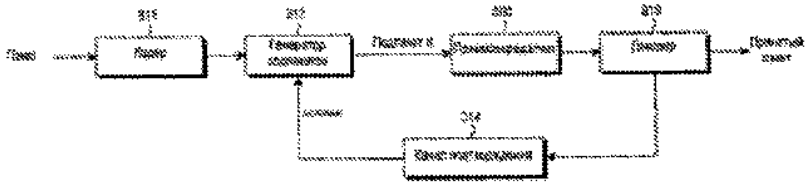
SUBSTANCE: method includes separation of resource elements from several code units into various groups into different groups and decoding of coded digits of resource elements in each group without waiting for complete reception of a transport unit to start its decoding. The method includes separation of coded digits from several code units into different groups and decoding of code units that include coded digits in each group. The first CRC joins the transport unit, and the second CRC joints at least one code unit from the transport unit. The method to configure an improved channel interleaver, including conversion from coded digits of various code units into modulation symbols and conversion from modulation symbols into time, frequency and space resources, to guarantee that each code unit receives approximately the same protective level.

EFFECT: improved noise immunity.

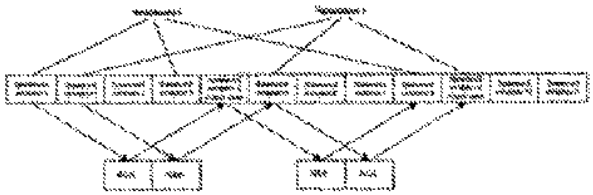
48 cl, 34 dwg

Images (33)

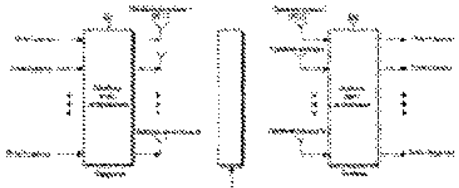




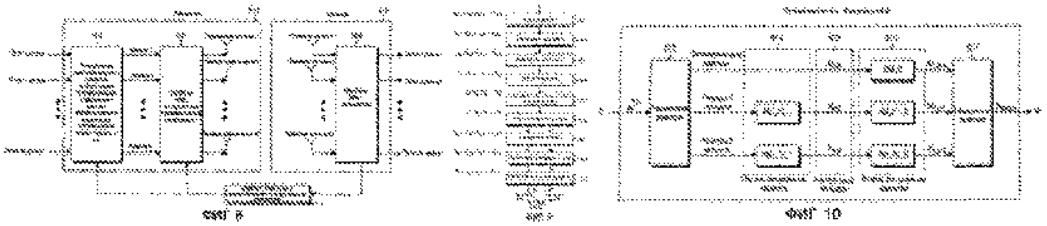
ФИГ. 5



ФИГ. 6



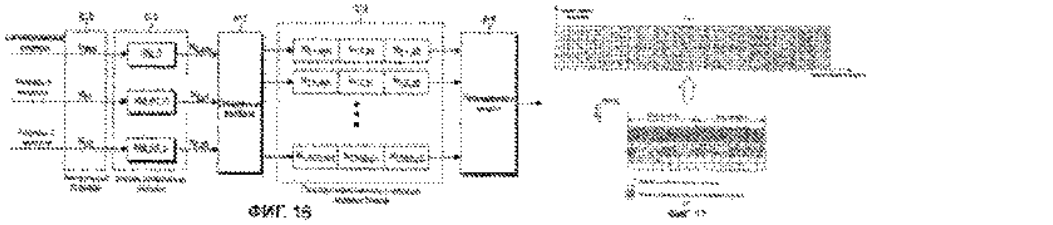
ФИГ. 7



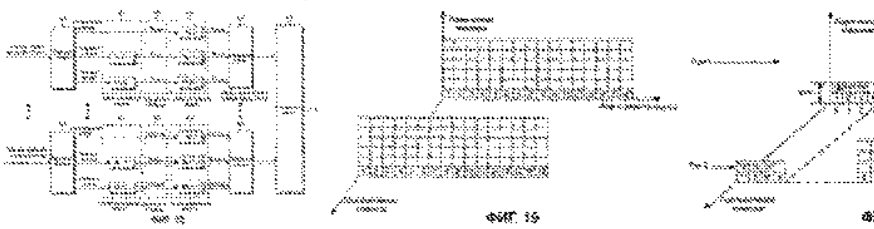
ФИГ. 8

ФИГ. 10

ФИГ. 11



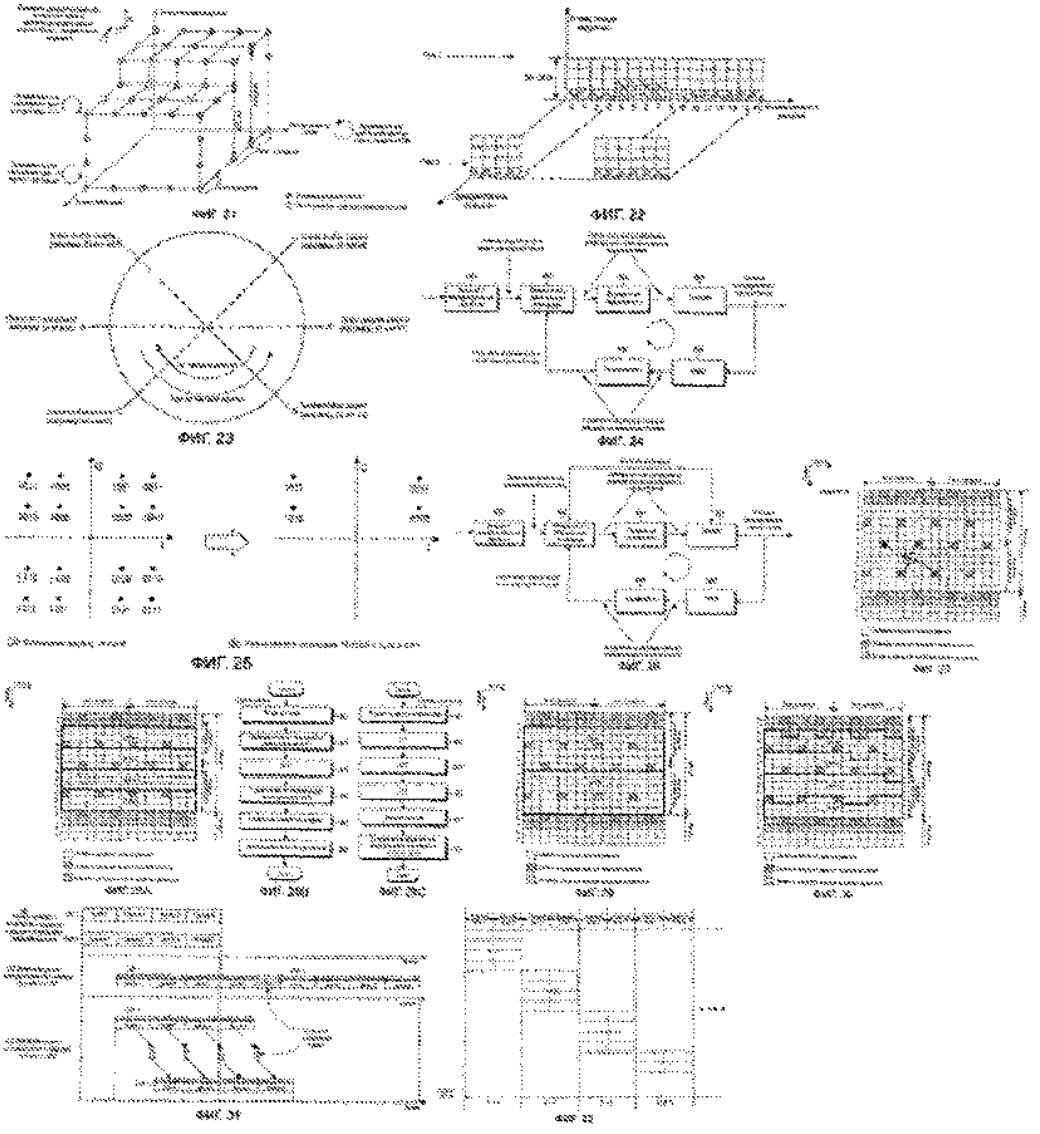
ФИГ. 14



ФИГ. 15

ФИГ. 19

ФИГ. 20



Classifications

H04L1/0041 Arrangements at the transmitter end

View 24 more classifications

Russia

Download PDF Find Prior Art Similar

Other languages

Russian

Inventor

Фарук КХАН (US)

Фарук КХАН

Чжоуюэ ПИ (US)

Чжоуюэ ПИ

Worldwide applications

2008 US ~~DE~~ ~~DE~~ EP EP ~~DE~~ ~~DE~~ JP CA ~~CN~~ CN WO RU KR AU TW 2012 JP US 2013 JP 2015 JP US

Application RU2009134548A events

2007-03-16

Priority to US91850307P

2008-03-13

Application filed by Самсунг Электроникс Ко., Лтд.

2011-03-20

Publication of RU2009134548A

2011-09-10

Application granted

2011-09-10

Publication of RU2428796C2

Show all events

Info

Patent citations (64)

Cited by (122)

Similar documents

Priority and Related Applications

External links

Espacenet

Global Dossier

Discuss

Description

BACKGROUND

FIELD OF THE INVENTION

The present invention relates to methods and devices for transmitting data in a communication system, and more particularly, to methods and devices for improving transmission performance with multiple code blocks and enabling fast decoding of transmissions with multiple code blocks in a communication system.

State of the art

Orthogonal Frequency Division Multiplexing (OFDM) is a technology for multiplexing data in the frequency domain. Modulation symbols are carried on frequency subcarriers, and the subcarriers overlap with each other in the frequency domain. However, orthogonality is maintained at the sampling frequency under the assumption that the transmitter and receiver have perfect frequency synchronization. In the case of frequency drift due to

imperfect frequency synchronization or due to high mobility, the orthogonality of the subcarriers at the sampling frequencies is violated, resulting in inter-carrier interference (ICI).

The cyclic prefix (CP) part in the received signal is often distorted by the previous orthogonal frequency division multiplexing (OFDM) symbol in multipath fading. When the cyclic prefix (CP) part is long enough, the received orthogonal frequency division multiplexing (OFDM) symbol without the cyclic prefix part (CP) should contain only its own signal that is minimized by the multipath fading channel. The main advantage of orthogonal frequency division multiplexing (OFDM) over other transmission schemes is that orthogonal frequency division multiplexing (OFDM) demonstrates reliability for compensating multipath fading.

Single-Carrier Frequency Division Multiple Access (SC-FDMA), which uses single-carrier modulation and frequency-domain correction, is a technique that has similar performance and complexity as the orthogonal frequency division multiplexed access system (OFDMA) . Single Carrier Frequency Division Multiple Access (SC-FDMA) is selected as the uplink shared access scheme in the Long Term Evolution (LTE) system of the Third Generation Partnership Project (3GPP). LTE 3GPP is a project in the Third Generation Partnership Project to improve the standard of mobile phones in the Universal Mobile Telecommunications System to cope with future requirements.

Hybrid Automatic Repeat reQuest (HARQ) is widely used in communication systems to combat decoding errors and improve reliability. N-channel synchronous hybrid automatic repeat request (HARQ) is often used in wireless communication systems because of the simplicity of the N-channel synchronous hybrid automatic repeat request (HARQ). Synchronous Hybrid Automatic Repeat reQuest (HARQ) is adopted as the HARQ scheme for the uplink in a long term evolution (LTE) system in 3GPP. On the downlink in LTE systems, an asynchronous adaptive HARQ is adopted as the HARQ scheme because of its flexibility and additional performance advantages over synchronous HARQ.

Multi-antenna communication systems, often referred to as multi-input multi-output (MIMO) systems, are widely used in wireless communications to enhance the performance of communications systems. In a MIMO system,

the transmitter has several antennas capable of transmitting independent signals, and the receiver is equipped with several receiving antennas. Many MIMO schemes are often used in an advanced wireless system.

When the channel is suitable, for example when the mobile communication speed is low, a closed loop with multiple inputs and multiple outputs (MIMO) can be used to improve system performance. In closed MIMO systems, receivers report back to the transmitter a channel condition and / or preferred MIMO transmission processing schemes. The transmitter uses this feedback information along with other considerations, such as scheduling priority, data and resource availability, to jointly optimize the transmission scheme. A popular closed-loop MIMO scheme is called MIMO precoding. Using precoding, the streams of transmitted data are pre-multiplied by a precoding matrix before being transmitted to several transmit antennas.

Another perspective of a multi-input multiple output (MIMO) system is whether multiple data streams are encoded for transmission separately or together. All layers for data transmission are encoded together in a single codeword (SCW) MIMO system, although all layers can be separately encoded in a multi-codeword MIMO system (MCW). Both single-user MIMO (SU-MIMO) and multi-user MIMO (MU-MIMO) are downlink adopted by Long Term Evolution (LTE). MU-MIMO is also adopted in the uplink of the Long Term Evolution (LTE) system, the adoption of SU-MIMO for the uplink of the Long Term Evolution (LTE) system, however, is still under discussion.

In a Long Term Evolution (LTE) system, when the transport block is large, the transport block is segmented into several code blocks so that several encoded packets can be generated. This partitioning of the transport block provides benefits such as enabling parallel processing or conveyor implementation and a flexible compromise between energy consumption and hardware complexity.

Different modulation schemes, such as quadrature phase shift keying (QPSK), on-off phase shift keying (BPSK), eight-position phase shift keying (8-PSK), 16-position quadrature amplitude modulation (16-QAM) or 64-position quadrature amplitude modulation (64-QAM), can be used for adaptive modulation and to increase the spectral efficiency of modulation. In the case of 16-QAM modulation, the four digits, b0b1b2b3, are converted to

complex-valued modulation symbols x = I + jQ. Different modulation positions, however, have different levels of protection.

When multiple code blocks are transmitted, transmission performance is dictated by the code block, which has the worst performance. A channel interleaver, including conversion from coded bits of different code blocks to modulation symbols and conversion from modulation symbols to temporal, frequency, and spatial resources, needs to be carefully designed to ensure that each code block receives approximately the same level of protection. When multiple code blocks are transmitted, it is useful to allow the receiver to start decoding some code blocks while the receiver is still demodulating modulation symbols for other code blocks. In a Long Term Evolution (LTE) system, this is a problem due to a poor effect on channel estimation performance if there are not enough reference signals during demodulation and decoding.

In order to maintain good channel estimation performance, reference signal interpolation is often used in selected resource elements located around a resource element that needs to be estimated in order to obtain a channel estimate for a resource element with improved performance. However, this means that the demodulation of the modulation symbol in the resource element to be evaluated must wait until all the resource elements selected for evaluating the resource element are received. In other words, if it becomes necessary to demodulate the resource element to be estimated before receiving the orthogonal frequency division multiplexing (OFDM) symbol, which contains some or all of the selected resource elements for estimating the resource element, channel estimation performance for resource elements can be severely affected.

SUMMARY OF THE INVENTION

The present invention is the creation of improved methods and devices for transmitting signals with multiple code blocks.

Another objective of the present invention is to provide an improved configuration of channel interleavers and advanced wireless receivers.

Another objective of the present invention is the creation of methods and devices that enable the rapid decoding of multiple code blocks, while maintaining good channel estimation performance.