송신기.

청구항 11

제7항에 있어서,

상기 전송 블록의 개수가 2개인 경우 상기 코드워드의 개수는 2개이고,

상기 2개의 코드워드 중 제 1 코드워드에 대응하는 심볼 스트림은 제 1 레이어 및 제 2 레이어에, 상기 2개의 코드워드 중 제 2 코드워드에 대응하는 심볼 스트림은 제 3 레이어 및 제 4 레이어에 각각 상기 소정 순서에 따라 분할되어 맵핑되는, 송신기.

청구항 12

제11항에 있어서,

상기 제 1 코드워드에 대응하는 심볼 스트림의 짝수번째 심볼들은 상기 제 1 레이어에, 상기 제 1 코드워드에 대응하는 심볼 스트림의 홀수번째 심볼들은 상기 제 2 레이어에 맵핑되고,

상기 제 2 코드워드에 대응하는 심볼 스트림의 짝수번째 심볼들은 상기 제 3 레이어에, 상기 제 2 코드워드에 대응하는 심볼 스트림의 홀수번째 심볼들은 상기 제 4 레이어에 맵핑하는, 송신기.

청구항 13

삭제

청구항 14

삭제

청구항 15

삭제

청구항 16

삭제

청구항 17

삭제

청구항 18

삭제

청구항 19

삭제

청구항 20

삭제

청구항 21

삭제

청구항 22

삭제

청구항 23

서지

청구항 24

명세

발명의 설명

발명의 상세한 설명

기 술 분 야

[0001] 본 발명은 무선 통신 시스템에 관한 것으로서, 보다 상세하게는, 무선 통신 시스템에서 송신측에서 다중 송신 안테나를 통한 신호 송신을 위한 데이터 처리 방법 및 송신기에 관한 것이다.

배 경 기 술

[0002] 최근에 무선 통신 시스템의 기반 기술로서 직교주파수분할다중화(OFDM: Orthogonal Frequency Division Multiplexing) 기법과 다중입력다중출력(MIMO: Multi-Input Multi-Output) 안테나 기술에 관한 연구가 활발하게 진행되고 있다.

[0003] OFDM의 기본 원리는 고속 전송률(high-rate)을 갖는 직렬 데이터 열(serial data stream)을 낮은 전송률(slow-rate)을 갖는 다수의 병렬 데이터 열로 분할하고, 다수의 반송파를 사용하여 병렬 데이터 열을 동시에 전송하는 것이다. 다시 말해서, 상기 고속 전송률을 갖는 데이터 열은, 직/병렬 변환부(Serial to Parallel converter)를 통해 다수의 낮은 전송률의 데이터 열로 변환되고, 상기 병렬로 변환된 다수의 데이터 열에 각각의 부 반송파가 곱해진 후 각각의 데이터 열이 합해져서 수신 단으로 전송된다. 상기 다수의 반송파 각각을 부반송파(sub-carrier)라 한다. OFDM 기법에서 다수의 반송파 사이에 직교성(orthogonality)이 존재하기 때문에, 반송파의 주파수 활용은 상호 중첩되어도 수신 단에서의 검출이 가능하다.

[0004] 직/병렬 변환부에 의해 생성된 다수의 병렬 데이터 열은, IDFT(Inverse Discrete Fourier Transform)에 의하여 다수의 부 반송파로 전송될 수 있으며, 상기 IDFT는 역 고속 푸리에 변환(IFFT: Inverse Fast Fourier Transform)을 사용하여 효율적으로 구현될 수 있다.

[0005] 낮은 전송률을 갖는 부 반송파의 심볼 구간(symbol duration)은 상대적으로 높은 다중경로 지연확산에 의해 발생하는 시간 상에서의 상대적인 신호 분산(dispersion)이 감소한다. OFDM 심볼 사이에 채널의 지연 확산보다 긴 보호구간(guard interval)을 삽입하여 심볼간 간섭(inter-symbol interference)을 줄일 수 있다. 또한, 보호구간에 OFDM 신호의 일부를 복사하여 심볼의 유효부분에 복사된 OFDM 심볼은 순환적으로 확장(cyclically extended)되어 심볼을 보호할 수 있다.

[0006] OFDMA(orthogonal frequency division multiple access) 방식은 OFDM을 보조 방식으로 사용하는 시스템에 있어 이용 가능한 부 반송파의 일부를 각 사용자에게 제공하여 다중 접속을 실현하는 다중 접속 방법을 말한다. OFDMA는 부 반송파라는 주파수 자원을 각 사용자에게 제공하며, 각각의 주파수 자원은 다수의 사용자에게 서로 독립적으로 제공되어 서로 중첩되지 않는 것이 일반적이다. 결국 주파수 자원은 상호 배타적으로 할당된다.

[0007] 도 1a 및 도 1b는 각각 종래기술에 따른 OFDMA 방식에 의한 송신측 및 수신측에서의 데이터 처리 절차를 설명하기 위한 블록도이다.

[0008] 도 1a를 참조하면, 송신측에서는 각 사용자를 위한 비트 스트림(bit stream)에 대하여 QPSK(quadrature phase shift keying), 16 QAM(quadrature amplitude modulation) 또는 64 QAM 등의 변조 기법에 따라 심볼 매핑(constellation mapping)을 수행한다[S11]. 심볼 매핑에 의해 생성된 복소 차원의 변조심볼은 하나의 심볼에 매핑된다.

[0009] 상기 비트 스트림은 데이터 심볼로 매핑되고, 상기 데이터 심볼은 직/병렬 변환부를 통해 병렬의 데이터 심볼로 변환된다[S12]. 상기 직/병렬 변환 과정에 의하여 상기 데이터 심볼은 각 사용자(n)에게 할당된 부반송파의 개수(Nsub(n))만큼의 병렬 심볼로 변환된다. 도 1a에 도시된 바와 같이, 사용자 1의 데이터 심볼은 상기 사용자 1에게 할당된 반송파의 개수인 Nsub(1)개 만큼의 병렬 심볼로 변환된다. 각 사용자(n)에게 할당되는 부반송파는 서로 중첩되거나 서로 다를 수 있기 때문에, 상기 각 사용자의 데이터 심볼은 서로 동일하거나 서로 다른 개수

$Nu(n)$개의 병렬 심볼로 변환될 수 있다.

[0010] 상기 병렬로 변환된 특정 사용자를 위한 데이터 심볼들은 전체 Nc개의 반송파 중에서 특정 n번째 사용자에 할당된 $Nu(n)$개의 부반송파에 매핑(mapping)되고, 나머지 $Nc-(Nu(n))$개의 반송파는 다른 사용자의 데이터 심볼들이 매핑된다(S13). 심볼을 반송파로 매핑시키는 매핑(symbol to subcarrier mapping) 모듈에 의하여, 사용자가 할당되지 않은 부반송파에는 0이 채워진다(zero padding). 상기 매핑(symbol to subcarrier Mapping) 모듈에 의한 Nc-포인트 IFFT(Inverse Fast Fourier Transform) 모듈에 의해 시간 영역 신호(time domain signals)로 변환된다(S14).

[0011] 상기 IFFT 모듈로부터 출력된 OFDM 심볼은 ISI(inter-symbol interference)를 줄이기 위해서 순환 전치(CP: cyclic prefix)가 삽입된다(S15). CP가 삽입된 OFDM 심볼들은 병/직렬 변환(parallel to serial conver)이 수행된 후(S16), 수신측으로 전송된다.

[0012] 도 1b를 참조하면, 종래기술에 따른 OFDMA 방식에 따른 수신측에서의 데이터 처리 과정은 도 1a에 도시된 바와 같은 송신측에서의 데이터 처리 과정의 역으로 진행된다. 수신된 데이터 심볼은 직/병렬 변환과 Nc-포인트 FFT를 거친 후 심볼에 매핑(subcarrier to symbol mapping)된다. 병렬의 심볼들은 직렬로 변환되어 심볼 디매핑 과정을 거쳐 비트열이 출력된다.

[0013] MIMO 기법은 이제까지 한 개의 송신 안테나와 한 개의 수신 안테나를 사용했던 것에서 탈피하여 다중 송신 안테나와 다중 수신 안테나를 채택해 데이터 송수신 효율을 향상시킬 수 있는 방법이다. 즉, 무선 통신 시스템의 송신단 및/또는 수신단에서 다중 안테나를 사용하여 용량 증대 혹은 성능 개선을 목적하는 기술이다. 이하에서 MIMO 기법은 다중 안테나 기법 또는 기술로 불릴 수 있다.

[0014] 다중안테나 기술은 하나의 전체 메시지를 수신하기 위해 단일 안테나 경로에 의존하지 않고 다수의 안테나에서 수신된 단편적인 데이터 조각을 한데 모아 완성하는 기술을 응용한 것이다. 특정 범위에서 데이터 전송 속도를 향상시키거나 특정 데이터 전송 속도에 대해 시스템 범위를 증가시킬 수 있다. 다중안테나 기술은 이동통신 단말과 중계기 등에 폭넓게 사용될 수 있는 차세대 이동통신 기술로서, 데이터 통신 확대 등으로 인해 한계 상황에 다다른 이동통신의 전송량 한계를 극복할 수 있는 차세대 기술로 활발한 연구로 있다. 차세대 이동통신은 기존 이동통신에 비해 훨씬 많은 데이터 전송량을 요구하므로 효율적인 다중안테나 기술이 반드시 필요할 것으로 예상된다.

[0015] 현재 연구되고 있는 다양한 전송효율 향상 기술 중 송/수신단 모두에 다수의 안테나를 사용하는 다중 안테나(MIMO) 기술은 추가적인 주파수 할당이나 전력증가 없이도 통신 용량 및 송수신 성능을 획기적으로 향상시킬 수 있는 방법으로서 현재 가장 큰 주목을 받고 있다.

[0016] 도 2는 종래기술에 따른 일반적인 다중 안테나 시스템의 개략 구성도이다. 도 2에 도시된 바와 같이, 송수신 단에서 안테나의 수를 같이 늘리게 되면 송신기나 수신기에서만 다수의 안테나를 사용하게 되는 경우와 달리 안테나 수에 비례하여 이론적인 채널 전송 용량이 증가하므로, 주파수 효율을 획기적으로 향상시킬 수 있다. 10 년대 중반 다중 안테나 시스템의 이론적 용량 증가가 증명된 이후 이를 실질적인 데이터 전송률 향상으로 이끌어내기 위한 다양한 기술들이 현재까지 활발히 연구되고 있으며, 이들 중 몇 개의 기술들은 이미 3세대 이동 통신과 차세대 무선랜 등의 다양한 무선 통신의 표준(standard)에 반영되고 있다.

[0017] 현재까지의 다중 안테나 관련 연구 동향을 살펴보면 다양한 채널 환경 및 다중 접속 환경에서의 다중 안테나 통신 용량 계산 등과 관련된 정보 이론 측면 연구, 다중 안테나 시스템의 무선 채널 측정 및 모형 도출 연구, 그리고 전송 신뢰도 향상 및 전송률 향상을 위한 시공간 신호 처리 기술 연구 등 다양한 관점에서 활발한 연구가 진행되고 있다.

[0018] 다중 안테나의 기술은 다양한 채널 경로를 통과한 심볼들을 이용하여 수신 신뢰도를 높이는 공간 다이버시티(spatial diversity) 방식과, 다수의 송신 안테나를 이용하여 다수의 데이터 심볼을 동시에 송신하여 전송률을 향상시키는 공간 멀티플렉싱(spatial multiplexing) 방식이 있다. 또한, 이러한 두 가지 방식을 적절히 결합하여 각각의 장점을 적절히 얻고자 하는 방식에 대한 연구도 최근 많이 이루어지고 있다.

[0019] 공간 다이버시티 방식의 경우에는 시공간 블록 부호 계열과 다이버시티 이득과 부호화 이득을 동시에 이용하는 시공간 트렐리스(trellis) 부호 계열 방식이 있다. 일반적으로 비트 오류율 개선 성능과 부호 생성 자유도는 트렐리스 부호 방식이 우수하지만, 연산 복잡도는 시공간 블록 부호가 간단하며, 공간 다이버시티 이득은 송신 안테나 수와 수신 안테나 수의 곱에 해당하는 양을 얻을 수 있다.

[0020] 공간 멀티플렉싱 기법은 각 송신 안테나에서 서로 다른 데이터 열을 송신하는 방법인데, 이때 수신기에서는 송신기로부터 동시에 전송된 데이터 사이에 상호 간섭이 발생하게 된다. 수신기에서는 이 간섭을 적절한 신호처리 기법을 이용하여 제거한 후 수신한다. 여기에 사용되는 잡음 제거 방식은 최대 우도(maximum likelihood) 수신기, ZF 수신기, MMSE 수신기, D-BLAST, V-BLAST 등이 있으며, 특히 송신단에서 채널 정보를 알 수 있는 경우에는 SVD(singular value decomposition) 방식 등을 사용할 수 있다.

[0021] 마지막으로, 공간 다이버시티와 공간 멀티플렉싱의 결합된 기법을 들 수 있다. 공간 다이버시티 이득만을 얻을 경우 다이버시티 차수의 증가에 따른 성능개선 이득이 점차 포화하며, 공간 멀티플렉싱 이득만을 취하면 무선 채널에서의 송신 신뢰도가 떨어진다. 이를 해결하면서 두 가지 이득을 모두 얻는 방식들에 대한 연구들이 있으며, 이 중 사용되는 것들 두오(Double-STTD), 사용된 BICM(STTD+BICM) 등의 방식이 있다.

[0022] 일반적인 통신 시스템에서는 채널에서 생긴 오류(error)를 수신단에서 정정해주기 위해서 송신단에서 보내는 정보를 오류정정부호(forward error correction code)를 사용하여 채널 부호화(channel encoding)를 수행한 후 전송하게 된다. 수신단에서는 수신신호를 복조(demodulation)한 후, 오류정정부호의 복호(decoding)와 과정을 거친 후 전송 정보를 복원하게 된다. 이러한 복호화 과정을 통해 채널에 의해서 생긴 수신 신호 상의 오류를 정정하게 된다.

[0023] 모든 오류정정부호에는 채널 오류 정정 시에 최대 정정 가능한 한계가 있다. 즉, 수신 신호가 해당 오류정정 부호가 갖는 한계를 넘는 오류를 갖고 있다면, 수신단에서는 오류가 없는 정보로 복호할 수 없게 된다. 따라서, 수신단에서는 복호한 정보에 오류가 있는지 없는지 판단할 필요가 생기게 된다. 이렇게, 오류정정부호와 함께 쓰여 오류 검출(error detection)을 위해서 특별한 형태의 부호화 과정이 필요하다. 이런 오류 검출 부호 중 일반적으로 CRC 코드(Cyclic Redundancy Check code)가 널리 쓰인다.

[0024] CRC는 오류 정정이 아니라 오류 검출을 위해 사용하는 부호화(coding) 방법의 하나이다. 일반적으로는 전송 정보를 CRC를 사용하여 부호화한 후, CRC 부호화된 정보에 오류정정부호를 사용하는 방식으로 이용된다.

[0025] 이하에서는 다양한 채널 코딩 기법들 중에서 터보 코드(turbo code)에 대해 설명하도록 한다. 터보 부호기(turbo encoder)는 두 개의 구성 부호화기들(recursive systematic convolution encoder)과 인터리버(interleaver)로 구성된다. 상기 인터리버는 터보 코드의 실제 구현시 병렬 복호화를 용이하게 하기 위해 사용되는데, 일종의 QPP(quadratic polynomial permutation)이다. 이러한 QPP 인터리버는 특정의 데이터 블록 크기에 대해서만 정의되어 있다. 터보 코드의 성능은 데이터 블록 크기가 증가할수록 좋은 것으로 알려져 있다. 그러나, 실제 통신 시스템에서는 실제 구현의 편리성을 위하여 일정 크기 이상의 데이터 블록(예를 들어, 전송 블록)의 경우 여러 개의 작은 데이터 블록으로 나누어 부호화를 수행하게 된다. 나누어진 작은 데이터 블록을 코드 블록(code block)이라 부른다. 즉, 하나의 코드워드는 그 길이가 길 경우 여러 개의 코드 블록으로 나뉘게 된다. 통상적으로 CRC와 오류 정정 부호가 적용되어 부호화된 하나의 덩어리를 '코드워드(codeword)'라고 하지만, 본 문서에서 코드워드의 의미는 하나의 데이터 블록, 예를 들어, 전송 블록(transport block)에 CRC가 부가된 후 채널 부호화가 수행된 데이터 덩어리를 '코드워드'라 한다. 따라서, 하나의 전송 블록의 크기가 일정 기준값을 초과하여 둘 이상의 코드 블록으로 분할되는 경우 모든 코드 블록들이 채널 부호화된 결과가 하나의 codeword가 된다.

[0026] 코드 블록은 일반적으로 같은 크기를 갖게 되지만, QPP 인터리버의 크기 제한 때문에 여러 개의 코드 블록들 중 하나의 코드 블록은 다른 크기를 가질 수도 있다. 코드 블록 단위로 오류 정정 부호화 과정을 거친 후 무선 채널을 통한 전송 과정에서의 버스트 에러(burst error)의 영향을 줄이기 위해 인터리빙(interleaving)을 수행한다. 그 다음에, 실제 무선 자원에 매핑되어 전송된다. 실제 전송시 사용되는 무선 자원의 양이 일정하기 때문에 이에 맞추기 위해서는 부호화된 코드 블록에 대하여 레이트 매칭(rate matching)을 수행하여야 한다. 일반적으로 레이트 매칭은 펑처링(puncturing)이나 반복(repetition) 과정에 의해 수행된다. 레이트 매칭은 3GPP의 ECDMA 시스템 예시와 같이 부호화된 코드 블록 단위로 수행할 수도 있다. 또는, 인터리빙된 코드 블록의 정보어 부분(systematic part)과 패리티 부분(parity part)을 분리하여 레이트 매칭을 수행할 수도 있다. 도 3은 부호화된 코드 블록의 정보어 부분과 패리티 부분을 분리하여 레이트 매칭을 수행하는 과정을 설명하기 위한 도면이다. 도 3에서 순환 버퍼(circular buffer)에서 전송 시작 위치와 전송할 데이터 크기에 따라 레이트 매칭을 수행할 수 있다. 도 3에서, 코드 레이트(code rate)는 1/3을 가정하였다.

[0027] 공간 멀티플렉싱을 이용하는 MIMO 시스템의 경우, 하나의 코드워드가 다수의 전송 데이터 스트림으로 전송되는 방법을 단일 코드워드(SCW: single codeword)라 하고, 하나 이상의 코드워드가 전송되는 방법을 멀티 코드워드(MCW: multiple codeword)라고 한다. 도 4a 및 도 4b는 SCW 및 MCW의 개념을 설명하기 위한 도면이며, SCW와

BCE의 하이브리드(hybrid) 형태의 구성도 가능하다. 예를 들면, 4 개의 송신 안테나와 4 개의 수신 안테나를 가정할 경우, 코드워드를 두 개만 사용할 수 있다. 이 경우, 2 개의 데이터 스트림을 전송하는 SCW가 두 개 연접되어 BCE 시스템을 구성하게 된다.

[0028] 도 5는 본 발명의 실시예에 따른 BCBL 시스템의 BS-BSU에 사용되는 코딩 체인(coding chain)을 도시한 것이다. 공간 멀티플ексing이 사용되는 경우 최대 2 개의 데이터 스트림 전송이 가능하고, 스트림은 BCE로 전송이 되기 때문에 1개의 스트림이 전송될 경우와 2개의 스트림이 전송될 경우, 모두 동일한 코딩 체인의 사용이 가능하다.

발명의 내용

해결 하고자하는 과제

[0029] 본 발명의 목적은 무선 통신 시스템에서 다중 송신 안테나를 통해 데이터를 전송하는 경우 하나의 코드워드를 둘 이상의 레이어(layer)로 효율적으로 맵핑하기 위한 데이터 처리 방법을 제공하는 것이다.

[0030] 본 발명의 다른 목적은 무선 통신 시스템에서 페이딩을 감소시키기 위해 다중 송신 안테나를 통해 데이터를 전송할 수 있도록 데이터를 처리하는 방법을 제공하는 것이다.

과제 해결수단

[0031] 본 발명에 따라 무선 통신 시스템에서 하나의 코드워드가 공간적으로 분배되는 방법이 제시된다. 페이딩(fading)은 무선 통신 시스템의 성능 저하를 가져오는 주요 원인들 중의 하나이다. 시간, 주파수, 공간에 따라 채널 이득값이 변하고 채널 이득값이 낮을수록 성능저하가 심각해진다. 페이딩을 극복할 수 있는 방법 중 하나인 다이버시티(diversity)는 여러 개의 독립적인 채널들이 모두 낮은 이득값을 가질 확률이 매우 낮다는 사실을 이용한다. 일반적으로 시간, 주파수, 공간상에서 두 지점간의 채널 이득값의 상관관계는 시간적, 주파수, 공간의 거리가 클수록 독립적이 된다. 따라서, 페이딩을 극복하려면 하나의 코드 블록(code block)으로 부호화된 비트들은 시간, 주파수, 공간에 걸쳐서 크게 퍼질 수 있도록 배치되도록 함으로써 다이버시티(diversity)에 의한 이득을 더 얻을 수 있다.

[0032] 본 발명의 일 양상으로서, 무선 통신 시스템에서 다중 안테나를 이용하여 송신 신호를 전송하는 경우 심볼 매핑(symbol mapping) 이후에 하나의 코드워드가 공간적으로 분할되는 방법이 제시된다. 하나의 코드워드가 둘 이상의 코드 블록들로 분할되어 채널 코딩되는 경우 공간 다이버시티 효과를 얻기 위해서 공간 분할을 인터리빙 이전에 수행할 때에는 공간 분할 기능에 심볼들을 넣어 레이어들에 분산시키는 것이 바람직하다. 공간 분할을 인터리빙 이후에 수행하는 경우에는 공간 분할 기능에 심볼들을 넣어 레이어들에 분산시킬 수도 있으며, 다른처럼 둘 이상의 심볼 그룹들로 분할하는 것도 가능하다.

[0033] 본 발명의 다른 양상으로서, 본 발명에 따른 데이터 처리 방법은, 무선 통신 시스템의 송신측에서 다중 송신 안테나를 통한 신호 송신을 위한 데이터 처리 방법에 있어서, 하나의 데이터 블록에 대해 채널 부호화를 수행하는 단계와, 상기 데이터 블록이 채널 부호화되어 출력된 비트 스트림에 대해 심볼 매핑을 수행하여 심볼 스트림을 출력하는 단계와, 상기 심볼 스트림에 포함된 심볼들의 순서를 재배열하는 단계와, 상기 재배열된 심볼 스트림을 적어도 둘 이상의 심볼 그룹으로 분할하여 출력하는 단계를 포함하여 구성될 수 있다.

[0034] 본 발명의 또 다른 양상으로서, 본 발명에 따른 데이터 처리 방법은, 무선 통신 시스템의 송신측에서 다중 송신 안테나를 통한 신호 송신을 위한 데이터 처리 방법에 있어서, 하나의 데이터 블록에 대해 채널 부호화를 수행하는 단계와, 상기 데이터 블록이 채널 부호화되어 출력된 비트 스트림에 대해 심볼 매핑을 수행하여 심볼 스트림을 출력하는 단계와, 상기 심볼 스트림을 적어도 둘 이상의 심볼 그룹으로 그룹별하여 각 심볼 그룹을 출력하는 단계를 포함하여 구성됨을 특징으로 한다.

[0035] 본 발명의 또 다른 양상으로서, 본 발명에 따른 송신기는, 무선 통신 시스템에서 다중 송신 안테나를 통한 신호 송신을 위한 송신기에 있어서, 하나의 데이터 블록에 대해 채널 부호화를 수행하는 채널 부호화기와, 상기 데이터 블록이 채널 부호화되어 출력된 비트 스트림에 대해 심볼 매핑을 수행하여 심볼 스트림을 출력하는 심볼 매핑 모듈과, 상기 심볼 스트림에 포함된 심볼들의 순서를 재배열하여 출력하는 인터리버와, 상기 재배열된 심볼 스트림을 적어도 둘 이상의 심볼 그룹으로 분할하여 출력하는 공간 분할 모듈을 포함하여 구성됨을 특징으로 한다.

[0036] 본 발명의 또 다른 양상으로서, 본 발명에 따른 송신기는, 무선 통신 시스템에서 다중 송신 안테나를 통한 신호 송신을 위한 송신기에 있어서, 하나의 데이터 블록에 대해 각 코드 블록 단위로 채널 부호화를 수행하는 채널 부호화기와, 상기 데이터 블록의 채널 부호화되어 출력된 비트 스트림에 대해 심볼 매핑을 수행하여 심볼 스트림을 출력하는 심볼 매핑 모듈과, 상기 심볼 스트림을 적어도 둘 이상의 심볼 그룹으로 그룹핑하여 각 심볼 그룹을 출력하는 층간 분배 모듈을 포함하여 구성될 수 있다.

[0037] 본 발명의 또 다른 양상으로서, 본 발명에 따른 데이터 처리 방법은, 무선 통신 시스템의 송신측에서 다중 송신 안테나를 통한 신호 송신을 위한 데이터 처리 방법에 있어서, 하나의 데이터 블록에 대해 채널 부호화를 수행하는 단계와, 상기 데이터 블록이 채널 부호화되어 출력된 비트 스트림에 대해 비트 그룹 단위로 인터리빙을 수행하는 단계와, 상기 인터리빙된 비트 스트림을 적어도 둘 이상의 비트 그룹으로 그룹핑하여 각 비트 그룹을 출력하는 단계와, 상기 각 비트 그룹에 대해 심볼 매핑을 수행하는 단계를 포함하여 구성될 수 있다.

[0038] 본 발명의 또 다른 양상으로서, 본 발명에 따른 송신기는, 무선 통신 시스템에서 다중 송신 안테나를 통한 신호 송신을 위한 송신기에 있어서, 하나의 데이터 블록에 대해 채널 부호화를 수행하는 채널 부호화기와, 상기 데이터 블록이 채널 부호화되어 출력된 비트 스트림에 대해 비트 그룹 단위로 인터리빙을 수행하는 인터리버와, 상기 인터리빙된 비트 스트림을 적어도 둘 이상의 비트 그룹으로 그룹핑하여 각 비트 그룹을 출력하는 층간 분배 모듈과, 상기 각 비트 그룹에 대해 심볼 매핑을 수행하는 심볼 매핑 모듈을 포함하여 구성될 수 있다.

[0039] 이상의 본 발명의 실시 양상들에 있어서, 채널 부호화는 하나 또는 둘 이상의 코드 블록들로 이루어지는 상기 데이터 블록에 대해 각 코드 블록 단위로 수행될 수 있다.

효 과

[0040] 본 발명에 따르면 하나의 데이터 블록이 다수의 코드 블록으로 분할되어 채널 부호화되는 경우에, 전송 채널에 부과된 기능의 추가에 의해 각 코드 블록에 동일한 종류 다이버시티를 줄 수 있는 효과가 있다.

발명의 실시를 위한 구체적인 내용

[0041] 이하에서 첨부된 도면을 참조하여 설명되는 본 발명의 실시예들에 의해 본 발명의 구성, 작용 및 다른 특징들이 용이하게 이해될 수 있을 것이다. 이하의 실시예들은 본 발명의 기술적 특징들이 E-UMTS(Evolved Universal Mobile Telecommunications System)에 적용된 예들이다. E-UMTS는 LTE(Long Term Evolution) 시스템에라 불리기도 한다. UMTS 및 E-UMTS의 기술 규격(technical specification)의 상세한 내용은 각각 "3rd Generation Partnership Project; Technical Specification Group Radio Access Network"의 Release 7 과 Release 8 을 참조할 수 있다.

[0042] 도 6은 LTE 시스템의 하향링크 FDD(Frequency Division Duplex) 서브프레임(subframe)의 구조를 도시한 것이다. 하나의 서브프레임의 길이가 1 ms로 짧아서 시간축으로는 채널의 변화가 적은 반면, 주파수축은 최대 20 MHz까지 사용할 수 있어 변화가 크며, 또 공간축으로의 변화는 채널이 원칙적 독립적일 수 있기 때문에, 주파수축과 시간축으로 고루 분산시켜 다이버시티 이득을 얻는 것이 필요하다. 본 발명에 따른 실시예들은 FDD 뿐만 아니라, FDD 시스템과는 다른 서브프레임 구조를 갖는 TDD(Time Division Duplex) 시스템에서도 적용 가능하다.

[0043] 도 7a 및 도 7b는 각각 본 발명의 일 실시예에 따른 전송 채널 구조를 도시한 것으로서, 도 7a는 레이어(layer)가 세 개인 랭크 3(Rank 3)의 전송 구조이고, 도 7b는 레이어의 개수가 4인 랭크 4(Rank 4)의 전송 구조이다.

[0044] 도 7a의 첫 번째 전송 블록(TB1)은 전송 채널에 따른 데이터 처리 과정을 거쳐 하나의 스트림(stream), 즉 하나의 레이어(layer)로 전달된다. 다시 말해서, 하나의 전송 블록에 CRC 부가 알고리즘에 따른 CRC가 부가된 후, 채널 부호화 과정을 거쳐 하나의 레이어에 할당된다. 전송 블록은 IMTS 시스템 또는 E-UMTS 시스템에서 사용되는 용어로서, 전송 채널(transport channel)을 통해 교환되는 기본적인 데이터 단위를 의미한다. 채널 부호화는 콘볼루션 코드(convolution code), 터보 코드(turbo code) 또는 LDPC(Low Density Parity Check) 코드 등에 의해 수행될 수 있다.

[0045] 도 7a의 두 번째 전송 블록(TB2) 및 도 7b의 첫 번째 및 두 번째 전송 블록(TB1, TB2)의 경우는 하나의 전송 블록이 두 개의 스트림, 즉 두 개의 레이어에 맵핑된다. 하나의 전송 블록의 크기가 미리 설정된 일정한 값을 초과하는 경우에는 다수의 코드 블록들(CB: code blocks)로 분할되어 처리된다. 여기, CRC는 전송 블록 단위로 부가되거나 코드 블록 단위로 부가될 수 있다. 또한, CRC가 부가된 전송 블록이 다수의 코드 블록들로 분할된 후

각 코드 블록에 다시 CRC를 부가하는 것도 가능하다. 채널 부호화된 코드 블록 다위로 수행된다. 채널 부호화가 수행된 코드 블록들을 각 해이어에 할당하는 과정에서 공간 다이버시티를 고려한 레이터 처리(data processing) 과정이 요구된다. 각 레이어에 할당될 코드 블록들은 다중 안테나 송신을 위한 프리코딩(precoding) 과정을 거쳐 다수의 송신 안테나들을 통해 수신측으로 송신된다.

[0046] 도 7a 및 도 7b에서 채널 부호화 과정 이후 각 레이어 별로 이질의 데이터 처리 과정과 본원된 다양한 실시예들을 이하에서 설명하도록 한다.

[0047] 도 8은 본 발명의 일 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다.

[0048] 도 8을 참조하면, 채널 부호화된 각 코드 블록(CB1, CB2)은 정보어 부분(systematic part)과 패리티 부분(parity part)으로 이루어진다. 레이트 매칭 모듈(81)은 채널 부호화된 코드 블록들에 대해 레이트 매칭을 수행한다. 레이트 매칭은 채널 부호화된 코드 블록들의 크기를 각 설정된 값에 맞춰주기는 과정을 의미한다. 예를 들어, 도 8의 순환 버퍼를 이용하여 전송 시에 위치를 제어함으로써 전송할 코드 블록의 크기를 조절할 수 있다. 레이트 매칭은 채널 부호화된 각 코드 블록별로 수행되거나 각 코드 블록의 연결된 전체에 대해 수행될 수 있다.

[0049] 송각 분할 모듈(82)은 레이트 매칭이 이루어진 비트 스트림을 두 개의 비트 스트림으로 분할하여 출력한다. 이때, 비트 스트림의 각 비트들의 순서는 변경되지 않는다. 분할되는 비트 스트림의 개수는 레이어의 개수에 대응한다. 도 8은 레이어의 개수가 2인 경우이다.

[0050] 분할된 각 비트 스트림은 각 심볼 매핑 모듈(83a, 83b)에 입력된다. 각 심볼 매핑 모듈(83a, 83b)은 심볼 매핑을 수행하여 심볼 시퀀스를 출력한다. 심볼 매핑 방법으로 QPSK(quadrature phase shift keying), 16 QAM(quadrature amplitude modulation) 또는 64 QAM 등의 변조 기법이 이용될 수 있으나 이에 한정되는 것은 아니다.

[0051] 각 심볼 매핑 모듈(83a, 83b)로부터 출력된 심볼 스트림은 각 인터리버(84a, 84b)에 입력된다. 각 인터리버(84a, 84b)는 각 심볼 스트림에 대해 인터리빙을 수행하여 심볼의 순서를 재배열한다. 상기 인터리빙은 OFDM 심볼 기반(OFDM symbol based) 인터리빙인 것이 바람직하다. OFDM 심볼 기반 인터리빙은 하나의 OFDM 심볼 내부에서 부반송파에 할당되는 심볼을 인터리빙하는 방법을 의미한다. OFDM 심볼 기반 인터리빙에 의해 부반송파에 할당되는 심볼의 순서가 재배열된다. 각 인터리버(84a, 84b)에 의해 인터리빙된 각 심볼 스트림은 각 레이어에 할당된다.

[0052] 도 9a는 본 발명의 다른 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다.

[0053] 도 9a를 참조하면, 레이트 매칭 모듈(91)은 채널 부호화된 코드 블록들(CB1, CB2)에 대해 레이트 매칭을 수행한다. 심볼 매핑 모듈(92)은 상기 레이트 매칭 모듈(91)로부터 출력된 비트 스트림에 대해 심볼 매핑을 수행하여 심볼 스트림을 출력한다. 상기 레이트 매칭 및 심볼 매핑에 대한 구체적 설명은 도 8의 설명 부분을 참조할 수 있다.

[0054] 인터리버(93)는 상기 심볼 매핑 모듈(92)로부터 출력된 심볼 스트림에 대해 인터리빙을 수행하여 심볼들의 순서를 재배열한다. 이때, 상기 인터리버(93)는 각 코드 블록(CB1, CB2)에 대응하는 심볼들이 서로 고루 섞여도록 인터리빙을 수행하는 것이 바람직하며, 다시 말해서, 채널 부호화된 첫 번째 코드 블록(CB1)이 매핑된 심볼 스트림과 두 번째 코드 블록(CB2)이 매핑된 심볼 스트림이, 도 9a의 (b)에 도시된 바와 같이, 서로 골고루 조합될 수 있도록 인터리빙을 수행한다. 인터리빙에 의한 심볼들의 재배열 순서는 소정의 알고리즘에 의해 미리 설정되는 것이 바람직하고, 도 8에서처럼, OFDM 심볼 기반 인터리빙을 수행하는 것이 바람직하다.

[0055] 공간 분할 모듈(94)은 상기 인터리버(93)로부터 출력된 심볼 스트림을 레이어의 개수에 따라 분할하여 출력한다. 분할된 각 심볼 스트림은 각 레이어에 할당된다. 도 9a에서, 상기 인터리버(93) 및 공간 분할 모듈(94)은 물리적으로 독립된 구성요소로 도시되었으나, 실제로 구현시에는 일체형으로 구성될 수 있다. 즉, 상기 인터리버(93)가 인터리빙을 수행한 후 다수의 심볼 스트림으로 분할하여 출력함으로써 각 심볼 스트림을 각 레이어에 할당하는 것도 가능하다.

[0056] 도 9b는 본 발명의 또 다른 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다. 도 9b에서, 레이트 매칭 모듈(95)에 의해 레이트 매칭이 수행되어 출력된 레이어 비트열은 비트 레벨 인터리버(96)로 입력된다. 상기 비트 레벨 인터리버(96)는 입력된 레이어 비트열에 대해 인터리빙을 수행한다. 이때, 인터리빙은 하나 혹 이상의 비트를 포함하는 비트 그룹 단위로 수행되는 것이 바람직하다. 각 비트 그룹에 포함되는 비트들의 개수는 채

및 제2 심볼 매핑 모듈(98a, 98b)에서 하나의 심볼에 매상되는 비트들의 개수와 동일하다. 예를 들어, 상기 제1 및 제2 심볼 매핑 모듈(98a, 98b)에서 심볼 매핑 기법으로 BPSK를 사용하는 경우 각 비트 그룹은 하나의 비트로 구성되고, QPSK를 사용하는 경우 두 개의 비트들로 구성되고, 16QAM을 사용하는 경우 네 개의 비트들로 구성되며, 64QAM을 사용하는 경우 여섯 개의 비트들로 구성된다. 도 9b에서는 심볼 매핑 기법으로 QPSK가 사용된 경우로서, 상기 비트 레벨 인터리버(96)에 행(row) 방향으로 쓰여진(write) 상기 데이터 비트열은 두 개씩의 비트들로 이루어지는 비트 그룹 단위로 열 방향으로 읽히(read) 출력된다.

[0057] 상기 비트 레벨 인터리버(96)로부터 출력된 데이터 비트열은 공간 분할 모듈(97)에 의해 레이어 개수에 따라 분할된다. 상기 제1 및 제2 심볼 매핑 모듈들(98a, 98b)은 각각 상기 공간 분할 모듈(97)에 의해 분할된 데이터 비트열에 대해 심볼 매핑을 수행한다. 도 9b는 심볼 매핑 기법으로 QPSK를 사용하는 경우이므로 두 개의 비트들이 하나의 심볼에 매상된다. 도 9b에서, 상기 공간 분할 모듈(97)에 의한 공간 분할 및 상기 제1 및 제2 심볼 매핑 모듈들(98a, 98b)에 의한 심볼 매핑의 순서는 바뀔 수 있다. 즉, 상기 비트 레벨 인터리버(96)로부터 출력된 데이터열에 대해 심볼 매핑을 수행한 후에, 레이어 개수에 따라 심볼열을 분할할 수 있다.

[0058] 도 9b에서 상기 제1 및 제2 심볼 매핑 모듈들(98a, 98b)로부터 출력된 각 심볼열은 도 9a의 공간 분할 모듈(94)로부터 출력된 각 심볼열과 동일함을 확인할 수 있다. 도 9a의 IS-기반 인터리버(92)는 심볼 레벨에서 인터리빙을 수행하고, 도 9b의 비트 레벨 인터리버(96)는 비트 레벨에서 인터리빙을 수행하지만, 상기 비트 레벨 인터리버(96)가 심볼 매핑 기법에 대응하는 개수의 비트들로 구성되는 비트 그룹 단위로 인터리빙을 수행하기 때문에 심볼 레벨에서 인터리빙을 한 것과 등가의(equivalent) 결과를 낳게 되는 것이다. 또한, 도 9b에서의 공간 분할 모듈(97)은 9a와 같이 하나의 심볼 매핑 모듈 뒤에 위치하여도 위와 등가의 결과를 낳게 된다.

[0059] 도 10a는 본 발명의 또 다른 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다. 도 10a의 실시예의 구조는 도 9a의 실시예와 유사하다. 도 9a의 공간 분할 모듈(94)이 복잡 분해 모듈(104)로 대체된 차이점이 있다. 상기 공간 분해 모듈(104)은 인터리버(102)로부터 출력된 심볼 스트림을 레이어의 개수에 따라 단순히 분할하는 데 그치지 않고, 분할 과정에서 심볼들의 순서를 다시 한 번 재배열하는 과정을 수행한다. 즉, 도 10a의 (b)에서, 상기 인터리버(102)에서 소정의 인터리빙 알고리즘에 따라 각 코드 블록에 대응하는 심볼들의 순서가 재배열된 후, 상기 공간 분해 모듈(104)에 의해 다수의 심볼 스트림으로 분할되면서 심볼들의 순서가 소정 방식에 의해 재조정된다. 이때, 상기 공간 분해 모듈(104)은 공간적인 인터리빙 기능을 수행한다고 볼 수 있다. 예를 들어, 각 코드 블록에 대응하는 심볼 스트림의 개수 만큼 심볼들에 의해 하나의 심볼 스트림을 구성하여 출력하고, 동일 방에 심볼들에 의해 다른 심볼 스트림을 구성하여 출력할 수 있다. 상기 공간 분해 모듈(104)에서의 심볼들의 순서를 재조정하는 방식은 공간 다이버시티 효과를 극대화하는 관점에서 자유롭게 결정될 수 있다. 도 10a에서, 실제 구현 측면에서는 상기 인터리버(103) 및 공간 분해 모듈(104)이 일체화되어 구성될 수 있을 것이다. 도 10의 레이어 매핑 모듈(101), 심볼 매핑 모듈(102), 인터리버(103)에 대한 구체적 설명은 도 9의 설명 부분을 참조할 수 있다.

[0060] 도 10b는 본 발명의 또 다른 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다. 도 10b의 실시예는, 도 10a의 실시예에서와 마찬가지로, 비트 레벨 인터리버(106)가 제1 및 제2 심볼 매핑 모듈들(108a, 108b)에서 사용하는 심볼 매핑 기법에 따라 적어도 하나 이상의 비트로 이루어지는 비트 그룹 단위로 인터리빙을 수행한다는 점에서 도 10a와 같다. 또한, 공간 분해 모듈(107)에서 레이어의 개수에 따른 데이터 비트열을 공간 분해하는 경우에 상기 비트 레벨 인터리버(106)에서 사용된 비트 그룹 단위로 분해가 이루어진다는 점도, 실제 구현 측면에는, 상기 비트 레벨 인터리버(106)와 상기 공간 분해 모듈(107)이 일체화되어 구현될 수 있음을 보여다. 또한, 도 10b에서, 상기 공간 분해 모듈(107)에 의한 공간 분해 및 상기 제1 및 제2 심볼 매핑 모듈들(108a, 108b)에 의한 심볼 매핑의 순서는 바뀔 수 있다. 즉, 도 10b에서의 공간 분해 모듈(107)은 10a와 같이 하나의 심볼 매핑 모듈 뒤에 위치하여도 위와 등가의 결과를 낳게 된다.

[0061] 도 11은 본 발명의 또 다른 실시예에 따른 전송 채널 구조를 설명하기 위한 도면이다. 도 10a의 실시예와 비교했을 때, 도 11의 실시예는 공간 분해 모듈(113)과 인터리버(114a, 114b)의 위치가 바뀌어진 구조를 갖는다. 즉, 심볼 매핑 모듈(112)로부터 출력된 심볼 스트림을 상기 공간 분해 모듈(113)에 의해 두 개의 심볼 스트림으로 분할된다. 이때, 상기 공간 분해 모듈(113)은 분할된 각 심볼 스트림이 각 코드 블록(CB1, CB2)에 대응하는 심볼들을 골고루 포함하도록 심볼들의 순서를 재조정하면서 상기 심볼 매핑 모듈(112)로부터 출력된 심볼 스트림을 분할한다. 상기 각 인터리버(114a, 114b)는 상기 공간 분해 모듈(113)로부터 출력된 각 심볼 스트림에 대해 인터리빙을 수행하여 심볼들을 재배열한다. 상기 각 인터리버(114a, 114b)로부터 출력된 심볼 스트림은 각 레이어에 할당된다. 도 11의 실시예, 상기 공간 분해 모듈(113) 및 상기 각 인터리버(114a, 114b)는 실제 구현 측면

에서 생략화되어 구성될 수 있을 것이다.

[0062] 도 8 및 도 11의 실시예들에 있어서도, 도 9b 또는 도 10b의 예와 같이, 채널 부호화된 코드 블록들에 대해서 심볼 매핑 이전 단계에서 비트 단위 인터리빙이 수행될 수 있다. 즉, 심볼 매핑 이전 단계에서 채널 부호화된 코드 블록들에 대해 비트 그룹 단위로 인터리빙을 수행하여 코드 블록들에 포함되는 비트들의 순서를 재배열할 수 있다.

[0063] 도 8 및 도 11의 실시예들에 걸쳐서 CS-기반 인터리빙이 사용되지 않는 경우를 고려해 볼 수 있는데, 이 경우에도 구 개의 코드블록이 하나의 심볼에 포함되지 않을 허용하여 하는 경우를 고려하면, 심볼 단위의 프로세싱이 비트 단위의 프로 세싱 보다 간단하다. 예를 들면, 예시된 CB1의 길이가 16 비트, CB2의 길이가 10 비트 이고 16QAM을 사용한다면, 4비트씩 1심볼이 구성되므로, CB1의 마지막 2비트와 CB2의 처음 2비트가 하나의 16QAM 심볼에 포함되게 된다. 이러한 것을 허용하지 않으면, CB1과 CB2의 길이를 모듈레이션 차수의 정수배로 제한 시켜야 한다. 이 경우는 앞에서 CS-기반 인터리빙을 항상 인터리빙(Identity Interleaving)으로 사용한 것과 동일하다. 따라서 CS-기반 인터리빙이 사용되지 않는 경우도 본 명세에서 제안하는 심볼 매핑 모듈과 층 맵핑(또는 분배) 모듈의 동작이 동일하게 적용된다.

[0064] 본 발명은 본 명세의 기술 및 필수적 특징을 벗어나지 않는 범위에서 다른 특정한 형태로 구체화될 수 있음은 당업자에게 자명하다. 따라서, 상기의 상세한 설명은 모든 면에서 제한적으로 해석되어서는 아니되고 예시적인 것으로 고려되어야 한다. 본 발명의 범위는 첨부된 청구항의 합리적 해석에 의해 결정되어야 하고, 본 발명의 등가적 범위 내에서의 모든 변경은 본 발명의 범위에 포함된다.

도면의 간단한 설명

[0065] 도 1a 및 도 1b는 각각 종래기술에 따른 eMBB 방식에 의한 송신측 및 수신측에서의 데이터 처리 절차를 설명하기 위한 블록도이다.

[0066] 도 2는 종래기술에 따른 일반적인 무선 인터넷 시스템의 개략 구성도이다.

[0067] 도 3은 부호화된 코드 블록의 정보어 부분과 패리티 부분을 분리하여 레이트 매칭을 수행하는 과정을 설명하기 위한 도면이다.

[0068] 도 4a 및 도 4b는 SDF 및 SDM의 개념을 설명하기 위한 도면이다.

[0069] 도 5는 종래기술에 따른 3GPP 시스템의 HS-SCCH에 사용되는 코딩 체인(coding chain)을 도시한 것이다.

[0070] 도 6은 LTE 시스템의 하향링크 서브프레임의 구조를 도시한 것이다.

[0071] 도 7a 및 도 7b는 각각 본 발명의 일 실시예에 따른 LTE 시스템의 전송 체인 구조를 도시한 것이다.

[0072] 도 8은 본 발명의 일 실시예에 따른 전송 체인 구조를 설명하기 위한 도면이다.

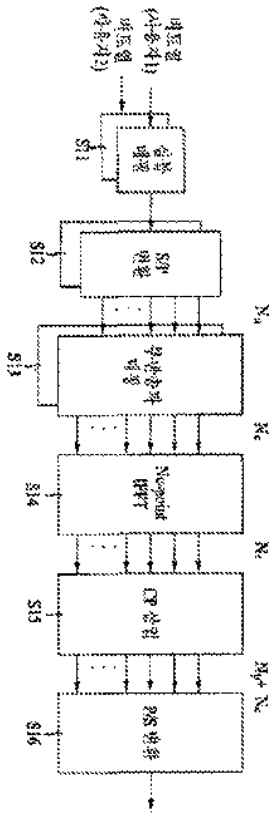
[0073] 도 9a 및 도 9b는 본 발명의 다른 실시예에 따른 전송 체인 구조를 설명하기 위한 도면이다.

[0074] 도 10a 및 도 10b는 본 발명의 또 다른 실시예에 따른 전송 체인 구조를 설명하기 위한 도면이다.

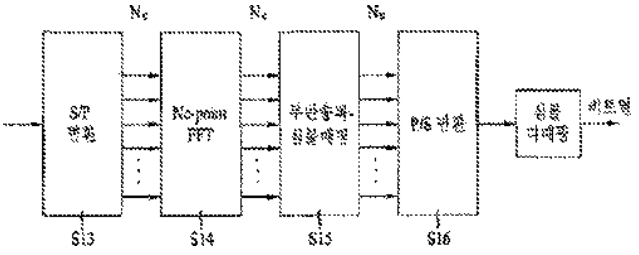
[0075] 도 11 본 발명의 또 다른 실시예에 따른 전송 체인 구조를 설명하기 위한 도면이다.

도면

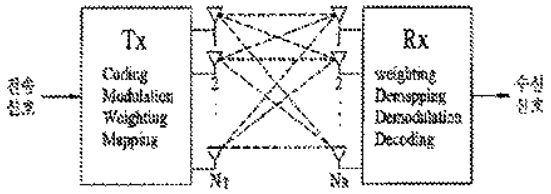
도면1a



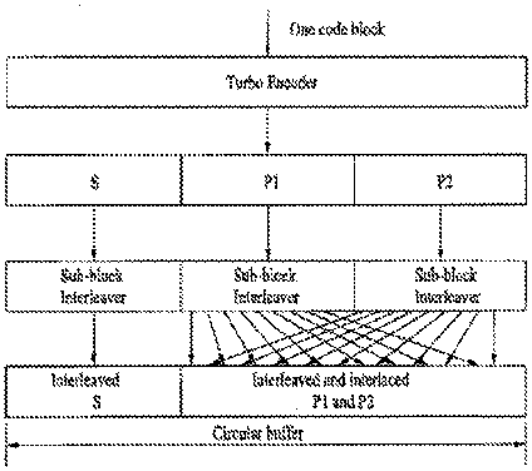
도면1b



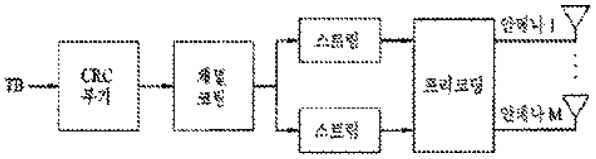
도면2



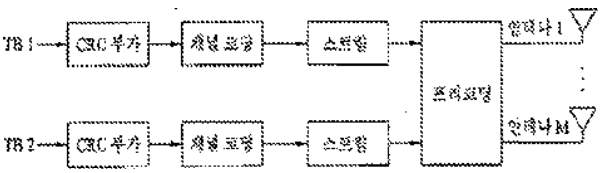
도면3



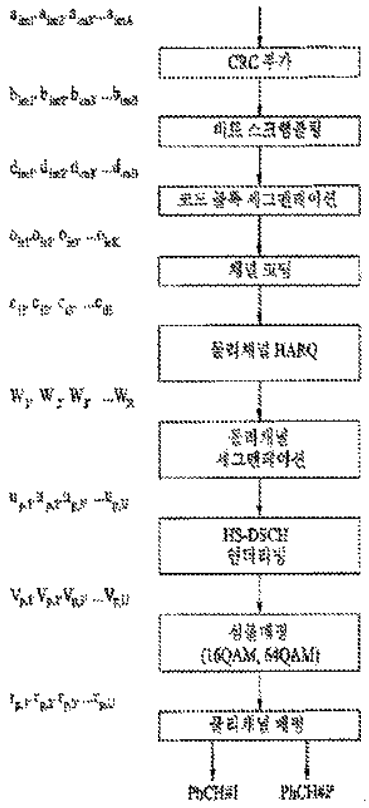
도면3a



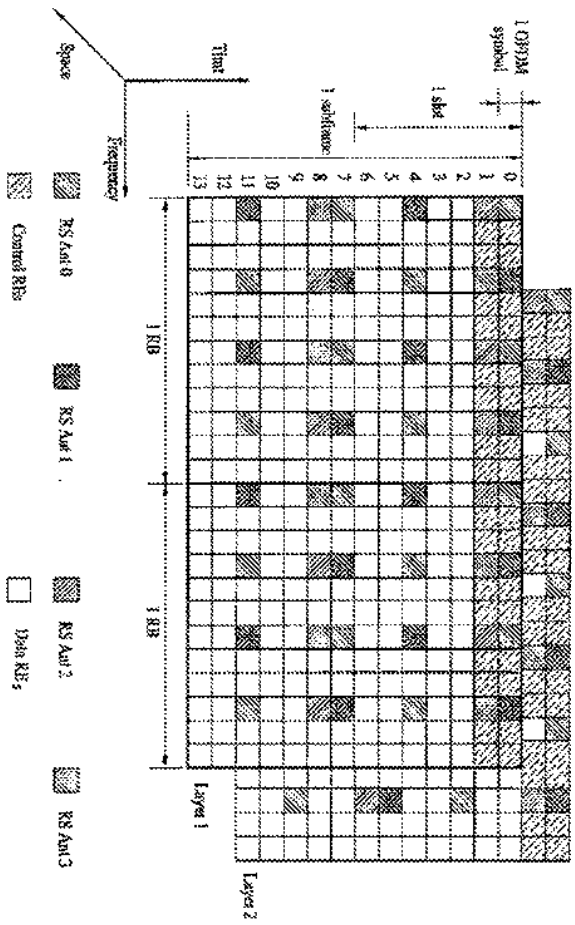
도면4



도면3


CRC 추가
터보 스크램블링
코드 블록 세그먼테이션
채널 코딩
물리채널 HARQ
프레임
세그멘테이션
HS-DSCH
인터리빙
성좌매핑
(16QAM, 64QAM)
물리채널 매핑
PhCH#1
PhCH#2

도면5


Space
Time
Frequency
1 OFDM symbol
1 slot
1 subframe
1 RB
1 RB
Layer 1
Layer 2
RS Ant 0
RS Ant 1
RS Ant 2
RS Ant 3
Control RBs
Data RBs

TB1
TB2
CRC 부가
채널 부호화
스크램블
CRC 부가
채널 부호화
채널별 프로세싱
스크램블 (Layer 1)
스크램블 (Layer 2)
프리코딩
안테나 1
안테나 2
안테나 3
안테나 4

도면 13

도면7

TB 4
TB 3
CRC 부가
CRC 부가
부호화
부호화
레이트 매칭
레이트 매칭
(2-layer) 심볼
(1-layer) 심볼
(2-layer) 심볼
(1-layer) 심볼
프리코딩
안테나 4
안테나 3
안테나 2
안테나 1

도면26

채널 부호화
91
93
94
Layer 1
Layer 2
CB1
CB2
S
P
Rate matched bits
1 OFDM Symbol

(a)

(b)

도면 15

CB1
CB2
S
P
S
P
1 OFDM Symbol
Layer 1
Layer 2
(a)
(b)

도면12

도면10b

도면 8F

(a)
Layer 2
Layer 1
OS 기반 응용 프로그램
네트워크 응용 프로그램
응용 계층 프로그램
통신 계층 프로그램
네트워크 계층 프로그램
(b)
1 OFDM Symbol
CH1
CH2

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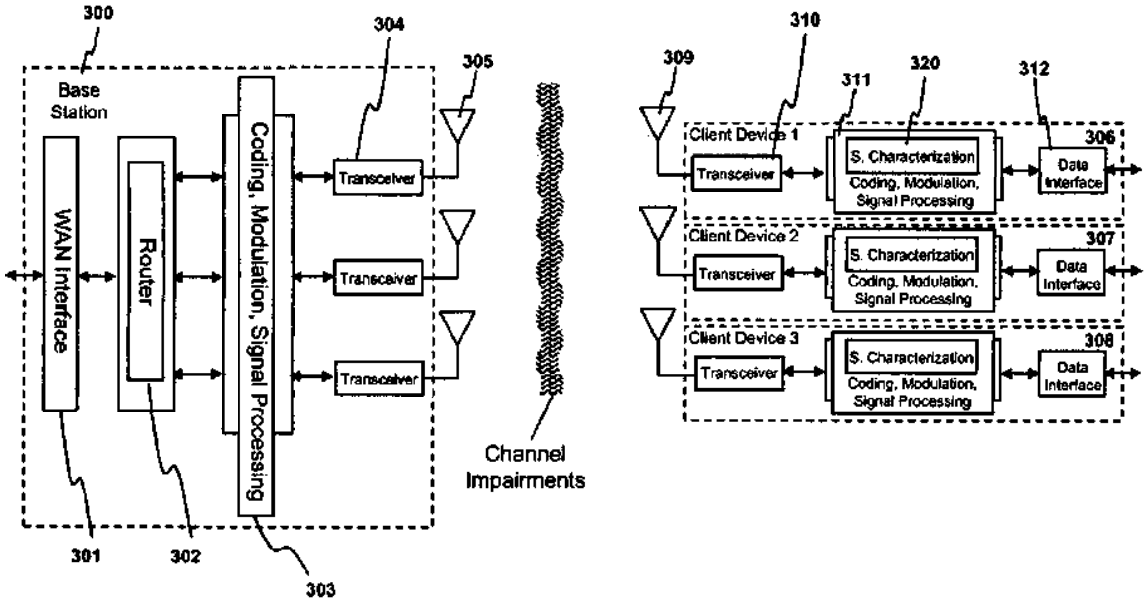
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3-Antenna Base Station with 3 Single-Antenna Client Devices

(57) **Abrégé/Abstract:**
A system and method are described for compensating for frequency and phase offsets in a multiple antenna system (MAS) with multiuser (MU) transmissions ("MU-MAS"). The method may include transmitting a training signal from each antenna of a base station to one of a plurality of wireless client devices, generating frequency offset compensation data, receiving the frequency offset compensation data at the base station, computing MU-MAS precoder weights based on the frequency offset compensation data to precancel the frequency offset at the transmitter, and computing a plurality of MU-MAS precoder weights based on channel characterization data.

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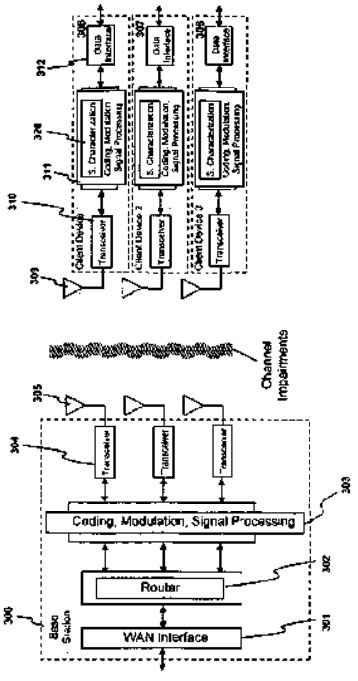


Fig. 3

3-Antenna Base Station with 3 Single-Antenna Client Devices

(57) Abstract: A system and method are described for compensating for frequency and phase offsets in a multiple antenna system (MAS) with multiuser (MU) transmissions ("MU-MAS"). The method may include transmitting a training signal from each antenna of a base station to one of a plurality of wireless client devices, generating frequency offset compensation data, receiving the frequency offset compensation data at the base station, computing MU-MAS precoder weights based on the frequency offset compensation data to precancel the frequency offset at the transmitter, and computing a plurality of MU-MAS precoder weights based on channel characterization data.

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SYSTEM AND METHOD FOR DISTRIBUTED INPUT DISTRIBUTED OUTPUT WIRELESS COMMUNICATIONS

BACKGROUND OF THE INVENTION

Claim To Priority

[0001] This application is a continuation in part of Application No. 10/902,978 filed July 30, 2004.

Field of the Invention

[0002] This invention relates generally to the field of communication systems. More particularly, the invention relates to a system and method for distributed input-distributed output wireless communications using space-time coding techniques.

Description of the Related Art

Space-Time Coding of Communication Signals

[0003] A relatively new development in wireless technology is known as spatial multiplexing and space-time coding. One particular type of space-time coding is called MIMO for "Multiple Input Multiple Output" because several antennas are used on each end. By using multiple antennas to send and receive, multiple independent radio waves may be transmitted at the same time within the same frequency range. The following articles provide an overview of MIMO:

[0004] *IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, VOL. 21, NO. 3, APRIL 2003* : "From Theory to Practice: An Overview of MIMO Space–Time Coded Wireless Systems", by David Gesbert, Member, IEEE, Mansoor Shafi, Fellow, IEEE, Da-shan Shiu, Member, IEEE, Peter J. Smith, Member, IEEE, and Ayman Naguib, Senior Member, IEEE.

[0005] IEEE TRANSACTIONS ON COMMUNICATIONS, VOL. 50, NO. 12, DECEMBER 2002: "Outdoor MIMO Wireless Channels: Models and Performance Prediction", David Gesbert, Member, IEEE, Helmut Bölcskei,

Member, IEEE, Dhananjay A. Gore, and Arogyaswami J. Paulraj, Fellow, IEEE.

[0006] Fundamentally, MIMO technology is based on the use of spatially distributed antennas for creating parallel spatial data streams within a common frequency band. The radio waves are transmitted in such a way that the individual signals can be separated at the receiver and demodulated, even though they are transmitted within the same frequency band, which can result in multiple statistically independent (i.e. effectively separate) communications channels. Thus, in contrast to standard wireless communication systems which attempt to inhibit multi-path signals (i.e., multiple signals at the same frequency delayed in time, and modified in amplitude and phase), MIMO can rely on uncorrelated or weakly-correlated multi-path signals to achieve a higher throughput and improved signal-to-noise ratio within a given frequency band. By way of example, MIMO technology achieves much higher throughput in comparable power and signal-to-noise ratio (SNR) conditions where a conventional non-MIMO system can achieve only lower throughput. This capability is described on Qualcomm Incorporated's (Qualcomm is one of the largest providers of wireless technology) website on a page entitled "What MIMO Delivers" at http://www.cdmatech.com/products/what_mimo_delivers.jsp: "MIMO is the only multiple antenna technique that increases spectral capacity by delivering two or more times the peak data rate of a system per channel or per MHz of spectrum. To be more specific, for wireless LAN or Wi-Fi® applications QUALCOMM's fourth generation MIMO technology delivers speeds of 315Mbps in 36MHz of spectrum or 8.8 Mbps/MHz. Compare this to the peak capacity of 802.11a/g (even with beam-forming or diversity techniques) which delivers only 54Mbps in 17MHz of spectrum or 3.18 Mbps/MHz."

[0007] MIMO systems typically face a practical limitation of fewer than 10 antennas per device (and therefore less than 10X throughput improvement in the network) for several reasons:

1. Physical limitations: MIMO antennas on a given device must have sufficient separation between them so that each receives a statistically independent signal. Although MIMO throughput improvements can be seen with antenna spacing of even fractions of the wavelength,

the efficiency rapidly deteriorates as the antennas get closer, resulting in lower MIMO throughput multipliers.

[0008] See, for example, the following references:

[1] D.-S. Shiu, G. J. Foschini, M. J. Gans, and J. M. Kahn, "Fading correlation and its effect on the capacity of multielement antenna systems," *IEEE Trans. Comm.*, vol. 48, no. 3, pp. 502–513, Mar. 2000.

[2] V. Pohl, V. Jungnickel, T. Haustein, and C. von Helmolt, "Antenna spacing in MIMO indoor channels," *Proc. IEEE Veh. Technol. Conf.*, vol. 2, pp. 749 – 753, May 2002.

[3] M. Stoytchev, H. Safar, A. L. Moustakas, and S. Simon, "Compact antenna arrays for MIMO applications," *Proc. IEEE Antennas and Prop. Symp.*, vol. 3, pp. 708 – 711, July 2001.

[4] A. Forenza and R. W. Heath Jr., "Impact of antenna geometry on MIMO communication in indoor clustered channels," *Proc. IEEE Antennas and Prop. Symp.*, vol. 2, pp. 1700 – 1703, June 2004.

[0009] Also, for small antenna spacing, mutual coupling effects may degrade the performance of MIMO systems.

[0010] See, for example, the following references:

[5] M.J. Fakhereddin and K.R. Dandekar, "Combined effect of polarization diversity and mutual coupling on MIMO capacity," Proc. IEEE Antennas and Prop. Symp., vol. 2, pp. 495–498, June 2003.

[7] P. N. Fletcher, M. Dean, and A. R. Nix, "Mutual coupling in multi-element array antennas and its influence on MIMO channel capacity," IEEE Electronics Letters, vol. 39, pp. 342–344, Feb. 2003.

[8] V. Jungnickel, V. Pohl, and C. Von Helmolt, "Capacity of MIMO systems with closely spaced antennas," IEEE Comm. Lett., vol. 7, pp. 361–363, Aug. 2003.

[10] J. W. Wallace and M. A. Jensen, "Termination-dependent diversity performance of coupled antennas: Network theory analysis," IEEE Trans. Antennas Propagat., vol. 52, pp. 98–105, Jan. 2004.

[13] C. Waldschmidt, S. Schulteis, and W. Wiesbeck, "Complete RF system model for analysis of compact MIMO arrays," IEEE Trans. on Veh. Technol., vol. 53, pp. 579–586, May 2004.

[14] M. L. Morris and M. A. Jensen, "Network model for MIMO systems with coupled antennas and noisy amplifiers," IEEE Trans. Antennas Propagat., vol. 53, pp. 545–552, Jan. 2005.

[0011] Moreover, as the antennas are crowded together, the antennas typically must be made smaller, which can impact the antenna efficiency as well.

[0012] See, for example, the following reference

[15] H. A. Wheeler, "Small antennas," IEEE Trans. Antennas Propagat., vol. AP-23, n.4, pp. 462-469, July 1975.

[16] J. S. McLean, "A re-examination of the fundamental limits on the radiation Q of electrically small antennas," IEEE Trans. Antennas Propagat., vol. 44, n.5, pp. 672-676, May 1996.

[0013] Finally, with lower frequencies and longer wavelengths, the physical size of a single MIMO device can become unmanageable. An extreme example is in the HF band, where MIMO device antennas may have to be separated from each other by 10 meters or more.

2. Noise limitations. Each MIMO receiver/transmitter subsystem produces a certain level of noise. As more and more of these subsystems are placed in close proximity to each other, the noise floor increases. Meanwhile, as increasingly more distinct signals need to be distinguished from each other in a many-antenna MIMO system, an increasingly lower noise floor is required.

3. Cost and power limitations. Although there are MIMO applications where cost and power consumption are not an issue, in a typical wireless product, both cost and power consumption are critical constraints in developing a successful product. A separate RF subsystem is required for each MIMO antenna, including separate Analog-to-Digital (A/D) and Digital-to-Analog (D/A) converters. Unlike many aspects of digital systems which scale with Moore's Law (an empirical observation, made by Intel co-founder Gordon Moore, that the number of transistors on an integrated circuit for minimum component cost doubles about every 24 months; source: http://www.intel.com/technology/mooreslaw/), such analog-intensive subsystems typically have certain physical structural size and power requirements, and scale in cost and power linearly. So, a many-antenna MIMO device would become prohibitively expensive and power consumptive compared to a single-antenna device.

[0014] As a result of the above, most MIMO systems contemplated today are on the order of 2-to-4 antennas, resulting in a 2-to-4X increase in throughput, and some increase in SNR due to the diversity benefits of a multi-antenna system. Up to 10 antenna MIMO systems have been contemplated

(particularly at higher microwave frequencies due to shorter wavelengths and closer antenna spacing), but much beyond that is impractical except for very specialized and cost-insensitive applications.

Virtual Antenna Arrays

[0015] One particular application of MIMO-type technology is a virtual antenna array. Such a system is proposed in a research paper presented at European Cooperation in the field of Scientific and Technical Research, EURO-COST, Barcelona, Spain, Jan 15-17, 2003: Center for Telecommunications Research, King's College London, UK: "A step towards MIMO: Virtual Antenna Arrays", Mischa Dohler & Hamid Aghvami.

[0016] Virtual antenna arrays, as presented in this paper, are systems of cooperative wireless devices (such as cell phones), which communicate amongst each other (if and when they are near enough to each other) on a separate communications channel than their primary communications channel to the their base station so as to operate cooperatively (e.g. if they are GSM cellular phones in the UHF band, this might be a 5 GHz Industrial Scientific and Medical (ISM) wireless band). This allows single antenna devices, for example, to potentially achieve MIMO-like increases in throughput by relaying information among several devices in range of each other (in addition to being in range of the base station) to operate as if they are physically one device with multiple antennas.

[0017] In practice, however, such a system is extremely difficult to implement and of limited utility. For one thing, there are now a minimum of two distinct communications paths per device that must be maintained to achieve improved throughput, with the second relaying link often of uncertain availability. Also, the devices are more expensive, physically larger, and consume more power since they have at a minimum a second

communications subsystem and greater computational needs. In addition, the system is reliant on very sophisticated real-time of coordination of all devices, potentially through a variety of communications links. Finally, as the simultaneous channel utilization (e.g. the simultaneous phone call transmissions utilizing MIMO techniques) grows, the computational burden for each device grows (potentially exponentially as channel utilization increases linearly), which may very well be impractical for portable devices with tight power and size constraints.

SUMMARY OF THE INVENTION

[0018] A system and method are described for compensating for frequency and phase offsets in a multiple antenna system (MAS) with multi-user (MU) transmissions ("MU-MAS"). For example, a method according to one embodiment of the invention comprises: transmitting a training signal from each antenna of a base station to one or each of a plurality of wireless client devices, one or each of the client devices analyzing each training signal to generate frequency offset compensation data, and receiving the frequency offset compensation data at the base station; computing MU-MAS precoder weights based on the frequency offset compensation data to pre-cancel the frequency offset at the transmitter; precoding training signal using the MU-MAS precoder weights to generate precoded training signals for each antenna of the base station; transmitting the precoded training signal from each antenna of a base station to each of a plurality of wireless client devices, each of the client devices analyzing each training signal to generate channel characterization data, and receiving the channel characterization data at the base station; computing a plurality of MU-MAS precoder weights based on the channel characterization data, the MU-MAS precoder weights calculated to pre-cancel frequency and phase offset and/or inter-user interference; precoding data using the MU-MAS precoder weights to generate precoded data signals for each antenna of the base station; and transmitting the

precoded data signals through each antenna of the base station to each respective client device.

BRIEF DESCRIPTION OF THE DRAWINGS

[0019] A better understanding of the present invention can be obtained from the following detailed description in conjunction with the drawings, in which:

[0020] **FIG. 1** illustrates a prior art MIMO system.

[0021] **FIG. 2** illustrates an *N*-antenna Base Station communicating with a plurality of Single-antenna Client Devices.

[0022] **FIG. 3** illustrates a three Antenna Base Station communicating with three Single-Antenna Client Devices

[0023] **FIG. 4** illustrates training signal techniques employed in one embodiment of the invention.

[0024] **FIG. 5** illustrates channel characterization data transmitted from a client device to a base station according to one embodiment of the invention.

[0025] **FIG. 6** illustrates a Multiple-Input Distributed-Output ("MIDO") downstream transmission according to one embodiment of the invention.

[0026] **FIG. 7** illustrates a Multiple-Input Multiple Output ("MIMO") upstream transmission according to one embodiment of the invention.

[0027] **FIG. 8** illustrates a base station cycling through different client groups to allocate throughput according to one embodiment of the invention.

[0028] **FIG. 9** illustrates a grouping of clients based on proximity according to one embodiment of the invention.

[0029] **FIG. 10** illustrates an embodiment of the invention employed within an NVIS system.

[0030] **FIG. 11** illustrates an embodiment of the DIDO transmitter with I/Q compensation functional units.

[0031] **FIG. 12** a DIDO receiver with I/Q compensation functional units.

[0032] **FIG. 13** illustrates one embodiment of DIDO-OFDM systems with I/Q compensation.

[0033] **FIG. 14** illustrates one embodiment of DIDO 2 x 2 performance with and without I/Q compensation.

[0034] **FIG. 15** illustrates one embodiment of DIDO 2 x 2 performance with and without I/Q compensation.

[0035] **FIG. 16** illustrates one embodiment of the SER (Symbol Error Rate) with and without I/Q compensation for different QAM constellations.

[0036] **FIG. 17** illustrates one embodiment of DIDO 2 x 2 performances with and without compensation in different user device locations.

[0037] **FIG. 18** illustrates one embodiment of the SER with and without I/Q compensation in ideal (i.i.d. (independent and identically-distributed)) channels.

[0038] **FIG. 19** illustrates one embodiment of a transmitter framework of adaptive DIDO systems.

[0039] **FIG. 20** illustrates one embodiment of a receiver framework of adaptive DIDO systems.

[0040] **FIG. 21** illustrates one embodiment of a method of adaptive DIDO-OFDM.

[0041] **FIG. 22** illustrates one embodiment of the antenna layout for DIDO measurements.

[0042] **FIG. 23** illustrates embodiments of array configurations for different order DIDO systems.

[0043] **FIG. 24** illustrates the performance of different order DIDO systems.

[0044] **FIG. 25** illustrates one embodiment of the antenna layout for DIDO measurements.

[0045] **FIG. 26** illustrates one embodiment of the DIDO 2 x 2 performance with 4-QAM and FEC rate ½ as function of the user device location.

[0046] **FIG. 27** illustrates one embodiment of the antenna layout for DIDO measurements.

[0047] **FIG. 28** illustrates how, in one embodiment, DIDO 8 x 8 yields larger SE than DIDO 2 x 2 for lower TX power requirement.

[0048] **FIG. 29** illustrates one embodiment of DIDO 2 x 2 performance with antenna selection.

[0049] **FIG. 30** illustrates average bit error rate (BER) performance of different DIDO precoding schemes in i.i.d. channels.

[0050] **FIG. 31** illustrates the signal to noise ratio (SNR) gain of ASel as a function of the number of extra transmit antennas in i.i.d. channels.

[0051] **FIG. 32** illustrates the SNR thresholds as a function of the number of users (*M*) for block diagnalization (BD) and ASel with 1 and 2 extra antennas in i.i.d. channels.

[0052] **FIG. 33** illustrates the BER versus per-user average SNR for two users located at the same angular direction with different values of Angle Spread (AS).

[0053] **FIG. 34** illustrates similar results as FIG. 33, but with higher angular separation between the users.

[0054] **FIG. 35** plots the SNR thresholds as a function of the AS for different values of the mean angles of arrival (AOAs) of the users.

[0055] **FIG. 36** illustrates the SNR threshold for an exemplary case of five users.

[0056] **FIG. 37** provides a comparison of the SNR threshold of BD and ASel, with 1 and 2 extra antennas, for two user case.

[0057] **FIG. 38** illustrates similar results as FIG. 37, but for a five user case.

[0058] **FIG. 39** illustrates the SNR thresholds for a BD scheme with different values of AS.

[0059] **FIG. 40** illustrates the SNR thresholds in spatially correlated channels with AS= 0.1° for BD and ASel with 1 and 2 extra antennas.

[0060] **FIG. 41** illustrates the computation of the SNR thresholds for two more channel scenarios with AS= 5°.

[0061] **FIG. 42** illustrates the computation of the SNR thresholds for two more channel scenarios with AS= 10°.

[0062] **FIGS. 43-44** illustrate the SNR thresholds as a function of the number of users (*M*) and angle spread (AS) for BD and ASel schemes, with 1 and 2 extra antennas, respectively.

[0063] **FIG 45** illustrates a receiver equipped with frequency offset estimator/compensator.

[0064] **FIG. 46** illustrates DIDO 2 × 2 system model according to one embodiment of the invention.

[0065] **FIG. 47** illustrates a method according to one embodiment of the invention.

[0066] **FIG. 48** illustrates SER results of DIDO 2 × 2 systems with and without frequency offset.

[0067] **FIG. 49** compares the performance of different DIDO schemes in terms of SNR thresholds.

[0068] **FIG. 50** compares the amount of overhead required for different embodiments of methods.

[0069] **FIG. 51** illustrates a simulation with a small frequency offset of f_{max} = 2Hz and no integer offset correction.

[0070] **FIG. 52** illustrates results when turning off the integer offset estimator.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

[0071] In the following description, for the purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. It will be apparent, however, to one skilled in the art that the present invention may be practiced without some of these specific details. In other instances, well-known structures and devices are shown in block diagram form to avoid obscuring the underlying principles of the invention.

[0072] **Figure 1** shows a prior art MIMO system with transmit antennas 104 and receive antennas 105. Such a system can achieve up to 3X the throughput that would normally be achievable in the available channel. There are a number of different approaches in which to implement the details of such a MIMO system which are described in published literature on the subject, and the following explanation describes one such approach.

[0073] Before data is transmitted in the MIMO system of **Figure 1**, the channel is "characterized." This is accomplished by initially transmitting a "training signal" from each of the transmit antennas 104 to each of the receivers 105. The training signal is generated by the coding and modulation subsystem 102, converted to analog by a D/A converter (not shown), and then converted from baseband to RF by each transmitter 103, in succession. Each receive antenna 105 coupled to its RF Receiver 106 receives each training signal and converts it to baseband. The baseband signal is converted to digital by an A/D converter (not shown), and the signal processing subsystem 107 characterizes the training signal. Each signal's characterization may include many factors including, for example, phase and amplitude relative to a

reference internal to the receiver, an absolute reference, a relative reference, characteristic noise, or other factors. Each signal's characterization is typically defined as a vector that characterizes phase and amplitude changes of several aspects of the signal when it is transmitted across the channel. For example, in a quadrature amplitude modulation ("QAM")-modulated signal the characterization might be a vector of the phase and amplitude offsets of several multipath images of the signal. As another example, in an orthogonal frequency division multiplexing ("OFDM")-modulated signal, it might be a vector of the phase and amplitude offsets of several or all of the individual sub-signals in the OFDM spectrum.

[0074] The signal processing subsystem 107 stores the channel characterization received by each receiving antenna 105 and corresponding receiver 106. After all three transmit antennas 104 have completed their training signal transmissions, then the signal processing subsystem 107 will have stored three channel characterizations for each of three receiving antennas 105, resulting in a 3x3 matrix 108, designated as the channel characterization matrix, "H." Each individual matrix element $H_{i,j}$ is the channel characterization (which is typically a vector, as described above) of the training signal transmission of transmit antenna 104 i as received by the receive antenna 105 j.

[0075] At this point, the signal processing subsystem 107 inverts the matrix H 108, to produce H^{-1}, and awaits transmission of actual data from transmit antennas 104. Note that various prior art MIMO techniques described in available literature, can be utilized to ensure that the H matrix 108 can be inverted.

[0076] In operation, a payload of data to be transmitted is presented to the data Input subsystem 100. It is then divided up into three parts by splitter 101 prior to being presented to coding and modulation subsystem 102. For

example, if the payload is the ASCII bits for "abcdef," it might be divided up into three sub-payloads of ASCII bits for "ad," "be," and "cf" by Splitter 101. Then, each of these sub-payloads is presented individually to the coding and modulation subsystem 102.

[0077] Each of the sub-payloads is individually coded by using a coding system suitable for both statistical independence of each signal and error correction capability. These include, but are not limited to Reed-Solomon coding, Viterbi coding, and Turbo Codes. Finally, each of the three coded sub-payloads is modulated using an appropriate modulation scheme for the channel. Examples of modulation schemes are differential phase shift key ("DPSK") modulation, 64-QAM modulation and OFDM. It should be noted here that the diversity gains provided by MIMO allow for higher-order modulation constellations that would otherwise be feasible in a SISO (Single Input-Single Output) system utilizing the same channel. Each coded and modulated signal is then transmitted through its own antenna 104 following D/A conversion by a D/A conversion unit (not shown) and RF generation by each transmitter 103.

[0078] Assuming that adequate spatial diversity exists amongst the transmit and receive antennas, each of the receiving antennas 105 will receive a different combination of the three transmitted signals from antennas 104. Each signal is received and converted down to baseband by each RF receiver 106, and digitized by an A/D converter (not shown). If y_n is the signal received by the nth receive antenna 105, and x_n is the signal transmitted by nth transmit antenna 104, and N is noise, this can be described by the following three equations:

$$y_1 = x_1 H_{11} + x_2 H_{12} + x_3 H_{13} + N$$
$$y_2 = x_1 H_{21} + x_2 H_{22} + x_3 H_{23} + N$$
$$y_3 = x_1 H_{31} + x_2 H_{32} + x_3 H_{33} + N$$

15

[0079] Given that this is a system of three equations with three unknowns, it is a matter of linear algebra for the signal processing subsystem 107 to derive x_1, x_2, and x_3 (assuming that N is at a low enough level to permit decoding of the signals):

$$x_1 = y_1H^{-1}{}_{11} + y_2H^{-1}{}_{12} + y_3H^{-1}{}_{13}$$
$$x_2 = y_1H^{-1}{}_{21} + y_2H^{-1}{}_{22} + y_3H^{-1}{}_{23}$$
$$x_3 = y_1H^{-1}{}_{31} + y_2H^{-1}{}_{32} + y_3H^{-1}{}_{33}$$

[0080] Once the three transmitted signals x_n are thus derived, they are then demodulated, decoded, and error-corrected by signal processing subsystem 107 to recover the three bit streams that were originally separated out by splitter 101. These bit streams are combined in combiner unit 108, and output as a single data stream from the data output 109. Assuming the robustness of the system is able to overcome the noise impairments, the data output 109 will produce the same bit stream that was introduced to the data Input 100.

[0081] Although the prior art system just described is generally practical up to four antennas, and perhaps up to as many as 10, for the reasons described in the Background section of this disclosure, it becomes impractical with large numbers of antennas (e.g. 25, 100, or 1000).

[0082] Typically, such a prior art system is two-way, and the return path is implemented exactly the same way, but in reverse, with each side of the communications channels having both transmit and receive subsystems.

[0083] **Figure 2** illustrates one embodiment of the invention in which a Base Station (BS) 200 is configured with a Wide Area Network (WAN) interface (e.g. to the Internet through a T1 or other high speed connection) 201 and is provisioned with a number (*N*) of antennas 202. For the time being, we use the term "Base Station" to refer to any wireless station that communicates

wirelessly with a set of clients from a fixed location. Examples of Base Stations are access points in wireless local area networks (WLANs) or WAN antenna tower or antenna array. There are a number of Client Devices 203-207, each with a single antenna, which are served wirelessly from the Base Station 200. Although for the purposes of this example it is easiest to think about such a Base Station as being located in an office environment where it is serving Client Devices 203-207 that are wireless-network equipped personal computers, this architecture will apply to a large number of applications, both indoor and outdoor, where a Base Station is serving wireless clients. For example, the Base Station could be based at a cellular phone tower, or on a television broadcast tower. In one embodiment, the Base Station 200 is positioned on the ground and is configured to transmit upward at HF frequencies (e.g., frequencies up to 24MHz) to bounce signals off the ionosphere as described in co-pending application entitled SYSTEM AND METHOD FOR ENHANCING NEAR VERTICAL INCIDENCE SKYWAVE ("NVIS") COMMUNICATION USING SPACE-TIME CODING, Serial No. 10/817,731, Filed April 20, 2004, which is assigned to the assignee of the present application.

[0084] Certain details associated with the Base Station 200 and Client Devices 203-207 set forth above are for the purpose of illustration only and are not required for complying with the underlying principles of the invention. For example, the Base Station may be connected to a variety of different types of wide area networks via WAN interface 201 including application-specific wide area networks such as those used for digital video distribution. Similarly, the Client Devices may be any variety of wireless data processing and/or communication devices including, but not limited to cellular phones, personal digital assistants ("PDAs"), receivers, and wireless cameras.

[0085] In one embodiment, the Base Station's n Antennas 202 are separated spatially such that each is transmitting and receiving signals which

are not spatially correlated, just as if the Base Station was a prior art MIMO transceiver. As described in the Background, experiments have been done where antennas placed within λ/6 (i.e. 1/6 wavelength) apart successfully achieve an increase in throughput from MIMO, but generally speaking, the further apart these Base Station antennas are placed, the better the system performance, and λ/2 is a desirable minimum. Of course, the underlying principles of the invention are not limited to any particular separation between antennas.

[0086] Note that a single Base Station 200 may very well have its antennas located very far apart. For example, in the HF spectrum, the antennas may be 10 meters apart or more (e.g., in an NVIS implementation mentioned above). If 100 such antennas are used, the Base Station's antenna array could well occupy several square kilometers.

[0087] In addition to spatial diversity techniques, one embodiment of the invention polarizes the signal in order to increase the effective throughput of the system. Increasing channel capacity through polarization is a well known technique which has been employed by satellite television providers for years. Using polarization, it is possible to have multiple (e.g., three) Base Station or users' antennas very close to each other, and still be not spatially correlated. Although conventional RF systems usually will only benefit from the diversity of two dimensions (e.g. *x* and *y*) of polarization, the architecture described herein may further benefit from the diversity of three dimensions of polarization (*x*, *y* and *z*).

[0088] In addition to space and polarization diversity, one embodiment of the invention employs antennas with near-orthogonal radiation patterns to improve link performance via pattern diversity. Pattern diversity can improve the capacity and error-rate performance of MIMO systems and its benefits

over other antenna diversity techniques have been shown in the following papers:

[17] L. Dong, H. Ling, and R. W. Heath Jr., "Multiple-input multiple-output wireless communication systems using antenna pattern diversity," *Proc. IEEE Glob. Telecom. Conf.*, vol. 1, pp. 997 – 1001, Nov. 2002.

[18] R. Vaughan, "Switched parasitic elements for antenna diversity," *IEEE Trans. Antennas Propagat.*, vol. 47, pp. 399 – 405, Feb. 1999.

[19] P. Mattheijssen, M. H. A. J. Herben, G. Dolmans, and L. Leyten, "Antenna-pattern diversity versus space diversity for use at handhelds," *IEEE Trans. on Veh. Technol.*, vol. 53, pp. 1035 – 1042, July 2004.

[20] C. B. Dietrich Jr, K. Dietze, J. R. Nealy, and W. L. Stutzman, "Spatial, polarization, and pattern diversity for wireless handheld terminals," *Proc. IEEE Antennas and Prop. Symp.*, vol. 49, pp. 1271 – 1281, Sep. 2001.

[21] A. Forenza and R. W. Heath, Jr., ``Benefit of Pattern Diversity Via 2-element Array of Circular Patch Antennas in Indoor Clustered MIMO Channels", *IEEE Trans. on Communications*, vol. 54, no. 5, pp. 943-954, May 2006.

Using pattern diversity, it is possible to have multiple Base Station or users' antennas very close to each other, and still be not spatially correlated.

[0089] **Figure 3** provides additional detail of one embodiment of the Base Station 200 and Client Devices 203-207 shown in **Figure 2**. For the purposes of simplicity, the Base Station 300 is shown with only three antennas 305 and only three Client Devices 306-308. It will be noted, however, that the embodiments of the invention described herein may be implemented with a virtually unlimited number of antennas 305 (i.e., limited only by available space and noise) and Client Devices 306-308.

[0090] **Figure 3** is similar to the prior art MIMO architecture shown in **Figure 1** in that both have three antennas on each sides of a communication channel. A notable difference is that in the prior art MIMO system the three antennas 105 on the right side of **Figure 1** are all a fixed distance from one another (e.g., integrated on a single device), and the received signals from each of the antennas 105 are processed together in the Signal Processing subsystem 107. By contrast, in **Figure 3**, the three antennas 309 on the right side of the diagram are each coupled to a different Client Device 306-308, each of which may be distributed anywhere within range of the Base Station 305. As such, the signal that each Client Device receives is processed independently from the other two received signals in its Coding, Modulation, Signal Processing subsystem 311. Thus, in contrast to a Multiple-Input (i.e. antennas 105) Multiple-Output (i.e. antennas 104) "MIMO" system, **Figure 3** illustrates a Multiple Input (i.e. antennas 305) Distributed Output (i.e. antennas 305) system, referred to hereinafter as a "MIDO" system.

[0091] Note that this application uses different terminology than previous applications, so as to better conform with academic and industry practices. In previously cited co-pending application, SYSTEM AND METHOD FOR ENHANCING NEAR VERTICAL INCIDENCE SKYWAVE ("NVIS") COMMUNICATION USING SPACE-TIME CODING, Serial No. 10/817,731, Filed April 20, 2004, and Application No. 10/902,978 filed July 30, 2004 for which this is application is a continuation-in-part, the meaning of "Input" and "Output" (in the context of SIMO, MISO, DIMO and MIDO) is reversed from how the terms are used in this application. In the prior applications, "Input" referred to the wireless signals as they are input to the receiving antennas (e.g. antennas 309 in Figure 3), and "Output" referred to the wireless signals as they are output by the transmitting antennas (e.g. antennas 305). In academia and the wireless industry, the reverse meaning of "Input" and "Output" is commonly used, in which "Input" refers to the wireless signals as

they are input to the channel (i.e. the transmitted wireless signals from antennas 305) and "Output" refers to the wireless signals as they are output from the channel (i.e. wireless signals received by antennas 309). This application adopts this terminology, which is the reverse of the applications cited previously in this paragraph. Thus, the following terminology equivalences shall be drawn between applications:

10/817,731 and 10/902,978		Current Application
SIMO	=	MISO
MISO	=	SIMO
DIMO	=	MIDO
MIDO	=	DIMO

[0092] The MIDO architecture shown in **Figure 3** achieves a similar capacity increase as MIMO over a SISO system for a given number of transmitting antennas. However, one difference between MIMO and the particular MIDO embodiment illustrated in **Figure 3** is that, to achieve the capacity increase provided by multiple base station antennas, each MIDO Client Device 306-308 requires only a *single* receiving antenna, whereas with MIMO, each Client Device requires *as least as many* receiving antennas as the capacity multiple that is hoped to be achieved. Given that there is usually a practical limit to how many antennas can be placed on a Client Device (as explained in the Background), this typically limits MIMO systems to between four to ten antennas (and 4X to 10X capacity multiple). Since the Base Station 300 is typically serving many Client Devices from a fixed and powered location, is it practical to expand it to far more antennas than ten, and to separate the antennas by a suitable distance to achieve spatial diversity. As illustrated, each antenna is equipped with a transceiver 304 and a portion of the processing power of a Coding, Modulation, and Signal Processing section 303. Significantly, in this embodiment, no matter how much Base Station 300

is expanded, each Client Device 306-308 only will require one antenna 309, so the cost for an individual user Client Device 306-308 will be low, and the cost of Base Station 300 can be shared among a large base of users.

[0093] An example of how a MIDO transmission from the Base Station 300 to the Client Devices 306-308 can be accomplished is illustrated in **Figures 4** through **6**.

[0094] In one embodiment of the invention, before a MIDO transmission begins, the channel is characterized. As with a MIMO system, a training signal is transmitted (in the embodiment herein described), one-by-one, by each of the antennas 405. **Figure 4** illustrates only the first training signal transmission, but with three antennas 405 there are three separate transmissions in total. Each training signal is generated by the Coding, Modulation, and Signal Processing subsystem 403, converted to analog through a D/A converter, and transmitted as RF through each RF Transceiver 404. Various different coding, modulation and signal processing techniques may be employed including, but not limited to, those described above (e.g., Reed Solomon, Viterbi coding; QAM, DPSK, QPSK modulation, . . . etc).

[0095] Each Client Device 406-408 receives a training signal through its antenna 409 and converts the training signal to baseband by Transceiver 410. An A/D converter (not shown) converts the signal to digital where is it processed by each Coding, Modulation, and Signal Processing subsystem 411. Signal characterization logic 320 then characterizes the resulting signal (e.g., identifying phase and amplitude distortions as described above) and stores the characterization in memory. This characterization process is similar to that of prior art MIMO systems, with a notable difference being that the each client device only computes the characterization vector for its one antenna, rather than for n antennas. For example, the Coding Modulation and Signal Processing subsystem 420 of client device 406 is initialized with a

known pattern of the training signal (either at the time of manufacturing, by receiving it in a transmitted message, or through another initialization process). When antenna 405 transmits the training signal with this known pattern, Coding Modulation and Signal Processing subsystem 420 uses correlation methods to find the strongest received pattern of the training signal, it stores the phase and amplitude offset, then it subtracts this pattern from the received signal. Next, it finds then second strongest received pattern that correlates to the training signal, it stores the phase and amplitude offset, then it subtracts this second strongest pattern from the received signal. This process continues until either some fixed number of phase and amplitude offsets are stored (e.g. eight), or a detectable training signal pattern drops below a given noise floor. This vector of phase/amplitude offsets becomes element H_{11} of the vector 413. Simultaneously, Coding Modulation and Signal Processing subsystems for Client Devices 407 and 408 implement the same processing to produce their vector elements H_{21} and H_{31}.

[0096] The memory in which the characterization is stored may be a non-volatile memory such as a Flash memory or a hard drive and/or a volatile memory such as a random access memory (e.g., SDRAM, RDAM). Moreover, different Client Devices may concurrently employ different types of memories to store the characterization information (e.g., PDA's may use Flash memory whereas notebook computers may use a hard drive). The underlying principles of the invention are not limited to any particular type of storage mechanism on the various Client Devices or the Base Station.

[0097] As mentioned above, depending on the scheme employed, since each Client Device 406-408 has only one antenna, each only stores a 1x3 row 413-415 of the H matrix. **Figure 4** illustrates the stage after the first training signal transmission where the first column of 1x3 rows 413-415 has been stored with channel characterization information for the first of the three Base Station antennas 405. The remaining two columns are stored following the

channel characterization of the next two training signal transmissions from the remaining two base station antennas. Note that for the sake of illustration the three training signals are transmitted at separate times. If the three training signal patterns are chosen such as not to be correlated to one another, they may be transmitted simultaneously, thereby reducing training time.

[0098] As indicated in **Figure 5**, after all three pilot transmissions are complete, each Client Device 506-508 transmits back to the Base Station 500 the 1x3 row 513-515 of matrix H that it has stored. To the sake of simplicity, only one Client Device 506 is illustrated transmitting its characterization information in **Figure 5**. An appropriate modulation scheme (e.g. DPSK, 64QAM, OFDM) for the channel combined with adequate error correction coding (e.g. Reed Solomon, Viterbi, and/or Turbo codes) may be employed to make sure that the Base Station 500 receives the data in the rows 513-515 accurately.

[0099] Although all three antennas 505 are shown receiving the signal in **Figure 5**, it is sufficient for a single antenna and transceiver of the Base Station 500 to receive each 1x3 row 513-515 transmission. However, utilizing many or all of antennas 505 and Transceivers 504 to receive each transmission (i.e., utilizing prior art Single-Input Multiple-Output ("SIMO") processing techniques in the Coding, Modulation and Signal Processing subsystem 503) may yield a better signal-to-noise ratio ("SNR") than utilizing a single antenna 505 and Transceiver 504 under certain conditions.

[0100] As the Coding, Modulation and Signal Processing subsystem 503 of Base Station 500 receives the 1x3 row 513-515, from each Client Device 507-508, it stores it in a 3x3 H matrix 516. As with the Client Devices, the Base Station may employ various different storage technologies including, but not limited to non-volatile mass storage memories (e.g., hard drives) and/or volatile memories (e.g., SDRAM) to store the matrix 516. **Figure 5** illustrates

a stage at which the Base Station 500 has received and stored the 1x3 row 513 from Client Device 509. The 1x3 rows 514 and 515 may be transmitted and stored in H matrix 516 as they are received from the remaining Client Devices, until the entire H matrix 516 is stored.

[0101] One embodiment of a MIDO transmission from a Base Station 600 to Client Devices 606-608 will now be described with reference to **Figure 6**. Because each Client Device 606-608 is an independent device, typically each device is receiving a different data transmission. As such, one embodiment of a Base Station 600 includes a Router 602 communicatively positioned between the WAN Interface 601 and the Coding, Modulation and Signal Processing subsystem 603 that sources multiple data streams (formatted into bit streams) from the WAN interface 601 and routes them as separate bit streams u_1- u_3 intended for each Client Device 606-608, respectively. Various well known routing techniques may be employed by the router 602 for this purpose.

[0102] The three bit streams, u_1- u_3, shown in **Figure 6** are then routed into the Coding, Modulation and Signal Processing subsystem 603 and coded into statistically distinct, error correcting streams (e.g. using Reed Solomon, Viterbi, or Turbo Codes) and modulated using an appropriate modulation scheme for the channel (such as DPSK, 64QAM or OFDM). In addition, the embodiment illustrated in **Figure 6** includes signal precoding logic 630 for uniquely coding the signals transmitted from each of the antennas 605 based on the signal characterization matrix 616. More specifically, rather than routing each of the three coded and modulated bit streams to a separate antenna (as is done in **Figure 1**), in one embodiment, the precoding logic 630 multiplies the three bit streams u_1- u_3 in **Figure 6** by the inverse of the H matrix 616, producing three new bit streams, u'_1- u'_3. The three precoded bit streams are then converted to analog by D/A converters (not shown) and transmitted as RF by Transceivers 604 and antennas 605.

[0103] Before explaining how the bit streams are received by the Client Devices 606-608, the operations performed by the precoding module 630 will be described. Similar to the MIMO example from **Figure 1** above, the coded and modulated signal for each of the three source bit streams will be designated with u_n. In the embodiment illustrated in **Figure 6**, each u_i contains the data from one of the three bit streams routed by the Router 602, and each such bit stream is intended for one of the three Client Devices 606-608.

[0104] However, unlike the MIMO example of **Figure 1**, where each x_i is *transmitted* by each antenna 104, in the embodiment of the invention illustrated in **Figure 6**, each u_i is *received* at each Client Device antenna 609 (plus whatever noise N there is in the channel). To achieve this result, the output of each of the three antennas 605 (each of which we will designate as v_i) is a function of u_i and the H matrix that characterizes the channel for each Client Device. In one embodiment, each v_i is calculated by the precoding logic 630 within the Coding, Modulation and Signal Processing subsystem 603 by implementing the following formulas:

$$v_1 = u_1H^{-1}{}_{11} + u_2H^{-1}{}_{12} + u_3H^{-1}{}_{13}$$
$$v_2 = u_1H^{-1}{}_{21} + u_2H^{-1}{}_{22} + u_3H^{-1}{}_{23}$$
$$v_3 = u_1H^{-1}{}_{31} + u_2H^{-1}{}_{32} + u_3H^{-1}{}_{33}$$

[0105] Thus, unlike MIMO, where each x_i is calculated at the receiver after the signals have been transformed by the channel, the embodiments of the invention described herein solve for each v_i at the transmitter *before* the signals have been transformed by the channel. Each antenna 609 receives u_i *already separated* from the other u_{n-1} bit streams intended for the other antennas 609. Each Transceiver 610 converts each received signal to baseband, where it is digitized by an A/D converter (now shown), and each Coding, Modulation and Signal Processing subsystem 611, demodulates and

decodes the x_i bit stream intended for it, and sends its bit stream to a Data Interface 612 to be used by the Client Device (e.g., by an application on the client device).

[0106] The embodiments of the invention described herein may be implemented using a variety of different coding and modulation schemes. For example, in an OFDM implementation, where the frequency spectrum is separated into a plurality of sub-bands, the techniques described herein may be employed to characterize each individual sub-band. As mentioned above, however, the underlying principles of the invention are not limited to any particular modulation scheme.

[0107] If the Client Devices 606-608 are portable data processing devices such as PDAs, notebook computers, and/or wireless telephones the channel characterization may change frequently as the Client Devices may move from one location to another. As such, in one embodiment of the invention, the channel characterization matrix 616 at the Base Station is continually updated. In one embodiment, the Base Station 600 periodically (e.g., every 250 milliseconds) sends out a new training signal to each Client Device, and each Client Device continually transmits its channel characterization vector back to the Base Station 600 to ensure that the channel characterization remains accurate (e.g. if the environment changes so as to affect the channel or if a Client Device moves). In one embodiment, the training signal is interleaved within the actual data signal sent to each client device. Typically, the training signals are much lower throughput than the data signals, so this would have little impact on the overall throughput of the system. Accordingly, in this embodiment, the channel characterization matrix 616 may be updated continuously as the Base Station actively communicates with each Client Device, thereby maintaining an accurate channel characterization as the Client Devices move from one location to the next or if the environment changes so as to affect the channel.

[0108] One embodiment of the invention illustrated in **Figure 7** employs MIMO techniques to improve the upstream communication channel (i.e., the channel from the Client Devices 706-708 to the Base Station 700). In this embodiment, the channel from each of the Client Devices is continually analyzed and characterized by upstream channel characterization logic 741 within the Base Station. More specifically, each of the Client Devices 706-708 transmits a training signal to the Base Station 700 which the channel characterization logic 741 analyzes (e.g., as in a typical MIMO system) to generate an *N* x *M* channel characterization matrix 741, where *N* is the number of Client Devices and *M* is the number of antennas employed by the Base Station. The embodiment illustrated in **Figure 7** employs three antennas 705 at the Base Station and three Client Devices 706-608, resulting in a 3x3 channel characterization matrix 741 stored at the Base Station 700. The MIMO upstream transmission illustrated in **Figure 7** may be used by the Client Devices both for transmitting data back to the Base Station 700, and for transmitting channel characterization vectors back to the Base Station 700 as illustrated in **Figure 5**. But unlike the embodiment illustrated in **Figure 5** in which each Client Device's channel characterization vector is transmitted at a separate time, the method shown in **Figure 7** allows for the simultaneous transmission of channel characterization vectors from multiple Client Devices back to the Base Station 700, thereby dramatically reducing the channel characterization vectors' impact on return channel throughput.

[0109] As mentioned above, each signal's characterization may include many factors including, for example, phase and amplitude relative to a reference internal to the receiver, an absolute reference, a relative reference, characteristic noise, or other factors. For example, in a quadrature amplitude modulation ("QAM")-modulated signal the characterization might be a vector of the phase and amplitude offsets of several multipath images of the signal. As another example, in an orthogonal frequency division multiplexing

("OFDM")-modulated signal, it might be a vector of the phase and amplitude offsets of several or all of the individual sub-signals in the OFDM spectrum. The training signal may be generated by each Client Device's coding and modulation subsystem 711, converted to analog by a D/A converter (not shown), and then converted from baseband to RF by each Client Device's transmitter 709. In one embodiment, in order to ensure that the training signals are synchronized, Client Devices only transmit training signals when requested by the Base Station (e.g., in a round robin manner). In addition, training signals may be interleaved within or transmitted concurrently with the actual data signal sent from each client device. Thus, even if the Client Devices 706-708 are mobile, the training signals may be continuously transmitted and analyzed by the upstream channel characterization logic 741, thereby ensuring that the channel characterization matrix 741 remains up-to-date.

[0110] The total channel capacity supported by the foregoing embodiments of the invention may be defined as *min (N, M)* where *M* is the number of Client Devices and *N* is the number of Base Station antennas. That is, the capacity is limited by the number of antennas on either the Base Station side or the Client side. As such, one embodiment of the invention employs synchronization techniques to ensure that no more than *min (N, M)* antennas are transmitting/ receiving at a given time.

[0111] In a typical scenario, the number of antennas 705 on the Base Station 700 will be less than the number of Client Devices 706-708. An exemplary scenario is illustrated in **Figure 8** which shows five Client Devices 804-808 communicating with a base station having three antennas 802. In this embodiment, after determining the total number of Client Devices 804-808, and collecting the necessary channel characterization information (e.g., as described above), the Base Station 800 chooses a first group of three clients 810 with which to communicate (three clients in the example because

min (N, M) = 3). After communicating with the first group of clients 810 for a designated period of time, the Base Station then selects another group of three clients 811 with which to communicate. To distribute the communication channel evenly, the Base Station 800 selects the two Client Devices 807, 808 which were not included in the first group. In addition, because an extra antenna is available, the Base Station 800 selects an additional client device 806 included in the first group. In one embodiment, the Base Station 800 cycles between groups of clients in this manner such that each client is effectively allocated the same amount of throughput over time. For example, to allocate throughput evenly, the Base Station may subsequently select any combination of three Client Devices which excludes Client Device 806 (i.e., because Client Device 806 was engaged in communication with the Base Station for the first two cycles).

[0112] In one embodiment, in addition to standard data communications, the Base Station may employ the foregoing techniques to transmit training signals to each of the Client Devices and receive training signals and signal characterization data from each of the Client Devices.

[0113] In one embodiment, certain Client Devices or groups of client devices may be allocated different levels of throughput. For example, Client Devices may be prioritized such that relatively higher priority Client Devices may be guaranteed more communication cycles (i.e., more throughput) than relatively lower priority client devices. The "priority" of a Client Device may be selected based on a number of variables including, for example, the designated level of a user's subscription to the wireless service (e.g., user's may be willing to pay more for additional throughput) and/or the type of data being communicated to/from the Client Device (e.g., real-time communication such as telephony audio and video may take priority over non-real time communication such as email).

[0114] In one embodiment of the Base Station dynamically allocates throughput based on the Current Load required by each Client Device. For example, if Client Device 804 is streaming live video and the other devices 805-808 are performing non-real time functions such as email, then the Base Station 800 may allocate relatively more throughput to this client 804. It should be noted, however, that the underlying principles of the invention are not limited to any particular throughput allocation technique.

[0115] As illustrated in **Figure 9**, two Client Devices 907, 908 may be so close in proximity, that the channel characterization for the clients is effectively the same. As a result, the Base Station will receive and store effectively equivalent channel characterization vectors for the two Client Devices 907, 908 and therefore will not be able to create unique, spatially distributed signals for each Client Device. Accordingly, in one embodiment, the Base Station will ensure that any two or more Client Devices which are in close proximity to one another are allocated to different groups. In **Figure 9**, for example, the Base Station 900 first communicates with a first group 910 of Client Devices 904, 905 and 908; and then with a second group 911 of Client Devices 905, 906, 907, ensuring that Client Devices 907 and 908 are in different groups.

[0116] Alternatively, in one embodiment, the Base Station 900 communicates with both Client Devices 907 and 908 concurrently, but multiplexes the communication channel using known channel multiplexing techniques. For example, the Base Station may employ time division multiplexing ("TDM"), frequency division multiplexing ("FDM") or code division multiple access ("CDMA") techniques to divide the single, spatially-correlated signal between Client Devices 907 and 908.

[0117] Although each Client Device described above is equipped with a single antenna, the underlying principles of the invention may be employed

using Client Devices with multiple antennas to increase throughput. For example, when used on the wireless systems described above, a client with 2 antennas will realize a 2x increase in throughput, a client with 3 antennas will realize a 3x increase in throughput, and so on (i.e., assuming that the spatial and angular separation between the antennas is sufficient). The Base Station may apply the same general rules when cycling through Client Devices with multiple antennas. For example, it may treat each antenna as a separate client and allocate throughput to that "client" as it would any other client (e.g., ensuring that each client is provided with an adequate or equivalent period of communication).

[0118] As mentioned above, one embodiment of the invention employs the MIDO and/or MIMO signal transmission techniques described above to increase the signal-to-noise ratio and throughput within a Near Vertical Incidence Skywave ("NVIS") system. Referring to **Figure 10**, in one embodiment of the invention, a first NVIS station 1001 equipped with a matrix of *N* antennas 1002 is configured to communicate with *M* client devices 1004. The NVIS antennas 1002 and antennas of the various client devices 1004 transmit signals upward to within about 15 degrees of vertical in order to achieve the desired NVIS and minimize ground wave interference effects. In one embodiment, the antennas 1002 and client devices 1004, support multiple independent data streams 1006 using the various MIDO and MIMO techniques described above at a designated frequency within the NVIS spectrum (e.g., at a carrier frequency at or below 23 MHz, but typically below 10 MHz), thereby significantly increasing the throughput at the designated frequency (i.e., by a factor proportional to the number of statistically independent data streams).

[0119] The NVIS antennas serving a given station may be physically very far apart from each other. Given the long wavelengths below 10 MHz and the long distance traveled for the signals (as much as 300 miles round trip),

physical separation of the antennas by 100s of yards, and even miles, can provide advantages in diversity. In such situations, the individual antenna signals may be brought back to a centralized location to be processed using conventional wired or wireless communications systems. Alternatively, each antenna can have a local facility to process its signals, then use conventional wired or wireless communications systems to communicate the data back to a centralized location. In one embodiment of the invention, NVIS Station 1001 has a broadband link 1015 to the Internet 1010 (or other wide area network), thereby providing the client devices 1003 with remote, high speed, wireless network access.

[0120] In one embodiment, the Base Station and/or users may exploit polarization/pattern diversity techniques described above to reduce the array size and/or users' distance while providing diversity and increased throughput. As an example, in MIDO systems with HF transmissions, the users may be in the same location and yet their signals be uncorrelated because of polarization/pattern diversity. In particular, by using pattern diversity, one user may be communicating to the Base Station via groundwave whereas the other user via NVIS.

ADDITIONAL EMBODIMENTS OF THE INVENTION

I. DIDO-OFDM Precoding with I/Q Imbalance

[0121] One embodiment of the invention employs a system and method to compensate for in-phase and quadrature (I/Q) imbalance in distributed-input distributed-output (DIDO) systems with orthogonal frequency division multiplexing (OFDM). Briefly, according to this embodiment, user devices estimate the channel and feedback this information to the Base Station; the Base Station computes the precoding matrix to cancel inter-carrier and inter-user interference caused by I/Q imbalance; and parallel data streams are transmitted to multiple user devices via DIDO precoding; the user devices

demodulate data via zero-forcing (ZF), minimum mean-square error (MMSE) or maximum likelihood (ML) receiver to suppress residual interference.

[0122] As described in detail below, some of the significant features of this embodiment of the invention include, but are not limited to:

[0123] Precoding to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch) in OFDM systems;

[0124] Precoding to cancel inter-user interference and ICI (due to I/Q mismatch) in DIDO-OFDM systems;

[0125] Techniques to cancel ICI (due to I/Q mismatch) via ZF receiver in DIDO-OFDM systems employing block diagonalization (BD) precoder;

[0126] Techniques to cancel inter-user interference and ICI (due to I/Q mismatch) via precoding (at the transmitter) and a ZF or MMSE filter (at the receiver) in DIDO-OFDM systems;

[0127] Techniques to cancel inter-user interference and ICI (due to I/Q mismatch) via pre-coding (at the transmitter) and a nonlinear detector like a maximum likelihood (ML) detector (at the receiver) in DIDO-OFDM systems;

[0128] The use of pre-coding based on channel state information to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch) in OFDM systems;

[0129] The use of pre-coding based on channel state information to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch) in DIDO-OFDM systems;

[0130] The use of an I/Q mismatch aware DIDO precoder at the station and an IQ-aware DIDO receiver at the user terminal;

[0131] The use of an I/Q mismatch aware DIDO precoder at the station, an I/Q aware DIDO receiver at the user terminal, and an I/Q aware channel estimator;

[0132] The use of an I/Q mismatch aware DIDO precoder at the station, an I/Q aware DIDO receiver at the user terminal, an I/Q aware channel estimator, and an I/Q aware DIDO feedback generator that sends channel state information from the user terminal to the station;

[0133] The use of an I/Q mismatch-aware DIDO precoder at the station and an I/Q aware DIDO configurator that uses I/Q channel information to perform functions including user selection, adaptive coding and modulation, space-time-frequency mapping, or precoder selection;

[0134] The use of an I/Q aware DIDO receiver that cancels ICI (due to I/Q mismatch) via ZF receiver in DIDO-OFDM systems employing block diagonalization (BD) precoder;

[0135] The use of an I/Q aware DIDO receiver that cancels ICI (due to I/Q mismatch) via pre-coding (at the transmitter) and a nonlinear detector like a maximum likelihood detector (at the receiver) in DIDO-OFDM systems; and

[0136] The use of an I/Q aware DIDO receiver that cancels ICI (due to I/Q mismatch) via ZF or MMSE filter in DIDO-OFDM systems.

a. Background

[0137] The transmit and receive signals of typical wireless communication systems consist of in-phase and quadrature (I/Q) components. In practical systems, the inphase and quadrature components may be distorted due to imperfections in the mixing and baseband operations. These distortions manifest as I/Q phase, gain and delay mismatch. Phase imbalance is caused by the sine and cosine in the modulator/demodulator not being perfectly orthogonal. Gain imbalance is caused by different amplifications between the inphase and quadrature components. There may be an additional distortion, called delay imbalance, due to difference in delays between the I-and Q-rails in the analog circuitry.

[0138] In orthogonal frequency division multiplexing (OFDM) systems, I/Q imbalance causes inter-carrier interference (ICI) from the mirror tones. This effect has been studied in the literature and methods to compensate for I/Q mismatch in single-input single-output SISO-OFDM systems have been proposed in M. D. Benedetto and P. Mandarini, "Analysis of the e□ect of the I/Q baseband filter mismatch in an OFDM modem," Wireless personal communications, pp. 175–186, 2000; S. Schuchert and R. Hasholzner, "A novel I/Q imbalance compensation scheme for the reception of OFDM signals," IEEE Transaction on Consumer Electronics, Aug. 2001; M. Valkama, M. Renfors, and V. Koivunen, "Advanced methods for I/Q imbalance compensation in communication receivers," IEEE Trans. Sig. Proc., Oct. 2001; R. Rao and B. Daneshrad, "Analysis of I/Q mismatch and a cancellation scheme for OFDM systems," IST Mobile Communication Summit, June 2004; A. Tarighat, R. Bagheri, and A. H. Sayed, "Compensation schemes and performance analysis of IQ imbalances in OFDM receivers," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 53, pp. 3257–3268, Aug. 2005.

[0139] An extension of this work to multiple-input multiple-output MIMO-OFDM systems was presented in R. Rao and B. Daneshrad, "I/Q mismatch cancellation for MIMO OFDM systems," in Personal, Indoor and Mobile Radio Communications, 2004; PIMRC 2004. 15th IEEE International Symposium on, vol. 4, 2004, pp. 2710–2714. R. M. Rao, W. Zhu, S. Lang, C. Oberli, D. Browne, J. Bhatia, J. F. Frigon, J. Wang, P; Gupta, H. Lee, D. N. Liu, S. G. Wong, M. Fitz, B. Daneshrad, and O. Takeshita, "Multiantenna testbeds for research and education in wireless communications," IEEE Communications Magazine, vol. 42, no. 12, pp. 72–81, Dec. 2004; S. Lang, M. R. Rao, and B. Daneshrad, "Design and development of a 5.25 GHz software defined wireless OFDM communication platform," IEEE Communications Magazine, vol. 42, no. 6, pp. 6–12, June 2004, for spatial multiplexing (SM) and in A. Tarighat and A. H. Sayed, "MIMO OFDM receivers for systems with IQ imbalances," IEEE Trans. Sig. Proc., vol. 53, pp. 3583–3596, Sep. 2005, for orthogonal space-time block codes (OSTBC).

[0140] Unfortunately, there is currently no literature on how to correct for I/Q gain and phase imbalance errors in a distributed-input distributed-output (DIDO) communication system. The embodiments of the invention described below provide a solution to these problems.

[0141] DIDO systems consist of one Base Station with distributed antennas that transmits parallel data streams (via pre-coding) to multiple users to enhance downlink throughput, while exploiting the same wireless resources (i.e., same slot duration and frequency band) as conventional SISO systems. A detailed description of DIDO systems was presented in S. G. Perlman and T. Cotter, "System and Method for Distributed Input-Distributed Output Wireless Communications," Serial No. 10/902,978, filed July 30, 2004 ("Prior Application"), which is assigned to the assignee of the present application.

[0142] There are many ways to implement DIDO precoders. One solution is block diagonalization (BD) described in Q. H. Spencer, A. L. Swindlehurst, and M. Haardt, "Zero forcing methods for downlink spatial multiplexing in multiuser MIMO channels," IEEE Trans. Sig. Proc., vol. 52, pp. 461–471, Feb. 2004. K. K. Wong, R. D. Murch, and K. B. Letaief, "A joint channel diagonalization for multiuser MIMO antenna systems," IEEE Trans. Wireless Comm., vol. 2, pp. 773–786, Jul 2003; L. U. Choi and R. D. Murch, "A transmit preprocessing technique for multiuser MIMO systems using a decomposition approach," IEEE Trans. Wireless Comm., vol. 3, pp. 20–24, Jan 2004; Z. Shen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization," accepted for publication in IEEE Trans. Sig. Proc., Sep. 2005; Z. Shen, R. Chen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Sum capacity of multiuser MIMO broadcast channels with block diagonalization," submitted to IEEE Trans. Wireless Comm., Oct. 2005; R. Chen, R. W. Heath, and J. G. Andrews, "Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers," accepted to IEEE Trans. on Signal Processing, 2005. The methods for I/Q compensation presented in this document assume BD precoder, but can be extended to any type of DIDO precoder.

[0143] In DIDO-OFDM systems, I/Q mismatch causes two effects: ICI and inter-user interference. The former is due to interference from the mirror tones as in SISO-OFDM systems. The latter is due to the fact that I/Q mismatch destroys the orthogonality of the DIDO precoder yielding interference across users. Both of these types of interference can be cancelled at the transmitter and receiver through the methods described herein. Three methods for I/Q compensation in DIDO-OFDM systems are described and their performance is compared against systems with and without I/Q mismatch. Results are

presented based both on simulations and practical measurements carried out with the DIDO-OFDM prototype.

[0144] The present embodiments are an extension of the Prior Application. In particular, these embodiments relate to the following features of the Prior Application:

[0145] The system as described in the prior application, where the I/Q rails are affected by gain and phase imbalance;

[0146] The training signals employed for channel estimation are used to calculate the DIDO precoder with I/Q compensation at the transmitter; and

[0147] The signal characterization data accounts for distortion due to I/Q imbalance and is used at the transmitter to compute the DIDO precoder according to the method proposed in this document.

b. Embodiments of the Invention

[0148] First, the mathematical model and framework of the invention will be described.

[0149] Before presenting the solution, it is useful to explain the core mathematical concept. We explain it assuming I/Q gain and phase imbalance (phase delay is not included in the description but is dealt with automatically in the DIDO-OFDM version of the algorithm). To explain the basic idea, suppose that we want to multiply two complex numbers $s = s_I + js_Q$ and $h = h_I + jh_Q$ and let $x = h * s$. We use the subscripts to denote inphase and quadrature components. Recall that

$$x_I = s_I h_I - s_Q h_Q$$

and

$$x_Q = s_I h_Q + s_Q h_I .$$

[0150] In matrix form this can be rewritten as

$$\begin{bmatrix} x_I \\ x_Q \end{bmatrix} = \begin{bmatrix} h_I & -h_Q \\ h_Q & h_I \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix}.$$

[0151] Note the unitary transformation by the channel matrix (**H**). Now suppose that s is the transmitted symbol and h is the channel. The presence of I/Q gain and phase imbalance can be modeled by creating a non-unitary transformation as follows

$$\begin{bmatrix} x_I \\ x_Q \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix}, \quad \text{(A)}$$

[0152] The trick is to recognize that it is possible to write

$$\begin{aligned} \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} &= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & h_{12}+h_{21} \\ h_{12}+h_{21} & h_{22}-h_{11} \end{bmatrix} \\ &= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix}, \end{aligned}$$

[0153] Now, rewriting (A)

$$\begin{aligned} \begin{bmatrix} x_I \\ x_Q \end{bmatrix} &= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix} \\ &= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix} \begin{bmatrix} s_I \\ -s_Q \end{bmatrix} \end{aligned} \quad (5)$$

[0154] Let us define

$$\mathcal{H}_e = \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix}$$

and

$$\mathcal{H}_c = \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix}.$$

[0155] Both of these matrices have a unitary structure thus can be equivalently represented by complex scalars as

$$h_e = h_{11}+h_{22}+j(h_{21}-h_{12})$$

and

$$h_c = h_{11}-h_{22}+j(h_{21}+h_{12}).$$

[0156] Using all of these observations, we can put the effective equation back in a scalar form with two channels: the equivalent channel h_e and the conjugate channel h_c. Then the effective transformation in (5) becomes

$$x = h_e s + h_c s^*.$$

[0157] We refer to the first channel as the equivalent channel and the second channel as the conjugate channel. The equivalent channel is the one you would observe if there were no I/Q gain and phase imbalance.

[0158] Using similar arguments, it can be shown that the input-output relationship of a discrete-time MIMO *NxM* system with I/Q gain and phase imbalance is (using the scalar equivalents to build their matrix counterparts)

$$\mathbf{x}[t] = \sum_{\ell=0}^{L} \mathbf{h}_e[\ell]\mathbf{s}[t-\ell] + \mathbf{h}_c[\ell]\mathbf{s}^*[t-\ell]$$

where t is the discrete time index, $\mathbf{h}_e, \mathbf{h}_c \in C^{MxN}, \mathbf{s} = [s_1, ..., s_N], \mathbf{x} = [x_1, ..., x_M]$ and L is the number of channel taps.

[0159] In DIDO-OFDM systems, the received signal in the frequency domain is represented. Recall from signals and systems that if $FFT_K\{s[t]\} = S[k]$ then $FFT_K\{s^*[t]\} = S^*[(-k)] = S^*[K-k]\ for\ k = 0, 1, ..., K-1.$

With OFDM, the equivalent input-output relationship for a MIMO-OFDM system for subcarrier k is

$$\bar{\mathbf{x}}[k] = \mathbf{H}_e[k]\bar{\mathbf{s}}[k] + \mathbf{H}_c[k]\bar{\mathbf{s}}^*[K-k] \quad (1)$$

where $k = 0, 1, \ldots, K-1$ is the OFDM subcarrier index, $\mathbf{H}_e$ and $\mathbf{H}_c$ denote the equivalent and conjugate channel matrices, respectively, defined as

$$\mathbf{H}_e[k] = \sum_{\ell=0}^{L} \mathbf{h}_e[\ell] e^{-j\frac{2\Pi k}{K}\ell}$$

and

$$\mathbf{H}_c[k] = \sum_{\ell=0}^{L} \mathbf{h}_c[\ell] e^{-j\frac{2\Pi k}{K}\ell}.$$

[0160] The second contribution in (1) is interference from the *mirror tone*. It can be dealt with by constructing the following stacked matrix system (note carefully the conjugates)

$$\begin{bmatrix} \bar{\mathbf{x}}[k] \\ \bar{\mathbf{x}}^*[K-k] \end{bmatrix} = \begin{bmatrix} \mathbf{H}_e[k] & \mathbf{H}_c[k] \\ \mathbf{H}_c^*[K-k] & \mathbf{H}_e^*[K-k] \end{bmatrix} \begin{bmatrix} \bar{\mathbf{s}}[k] \\ \bar{\mathbf{s}}^*[K-k] \end{bmatrix}$$

where $\bar{\mathbf{s}} = [\bar{s}_1, \bar{s}_2]^T$ and $\bar{\mathbf{x}} = [\bar{x}_1, \bar{x}_2]^T$ are the vectors of transmit and receive symbols in the frequency domain, respectively.

[0161] Using this approach, an effective matrix is built to use for DIDO operation. For example, with DIDO 2 x 2 the input-output relationship (assuming each user has a single receive antenna) the first user device sees (in the absence of noise)

$$\begin{bmatrix} \bar{x}_1[k] \\ \bar{x}_1^*[K-k] \end{bmatrix} = \begin{bmatrix} \mathbf{H}_e^{(1)}[k] & \mathbf{H}_c^{(1)}[k] \\ \mathbf{H}_c^{(1)*}[K-k] & \mathbf{H}_e^{(1)*}[K-k] \end{bmatrix} \mathbf{W} \begin{bmatrix} \bar{s}_1[k] \\ \bar{s}_1^*[K-k] \\ \bar{s}_2[k] \\ \bar{s}_2^*[K-k] \end{bmatrix} \quad (2)$$

while the second user observes

$$\begin{bmatrix} \bar{x}_2[k] \\ \bar{x}_2^*[K-k] \end{bmatrix} = \begin{bmatrix} \mathbf{H}_e^{(2)}[k] & \mathbf{H}_c^{(2)}[k] \\ \mathbf{H}_c^{(2)*}[K-k] & \mathbf{H}_e^{(2)*}[K-k] \end{bmatrix} \mathbf{W} \begin{bmatrix} \bar{s}_1[k] \\ \bar{s}_1^*[K-k] \\ \bar{s}_2[k] \\ \bar{s}_2^*[K-k] \end{bmatrix} \quad (3)$$

where $\mathbf{H}_e^{(m)}, \mathbf{H}_c^{(m)} \in C^{1x2}$ denote the *m-th* row of the matrices $\mathbf{H}_e$ and $\mathbf{H}_c$, respectively, and $\mathbf{W} \in C^{4x4}$ is the DIDO pre-coding matrix. From (2) and (3) it is observed that the received symbol $\bar{x}_m[k]$ of user *m* is affected by two sources of interference caused by I/Q imbalance: *inter-carrier interference* from the mirror tone (i.e., $\bar{s}_m^*[K-k]$) and *inter-user interference* (i.e., $\bar{s}_p[k]$ and $\bar{s}_p^*[K-k]$ with $p \neq m$). The DIDO precoding matrix **W** in (3) is designed to cancel these two interference terms.

[0162] There are several different embodiments of the DIDO precoder that can be used here depending on joint detection applied at the receiver. In one

embodiment, block diagonalization (BD) is employed (see, e.g., Q. H. Spencer, A. L. Swindlehurst, and M. Haardt, "Zeroforcing methods for downlink spatial multiplexing in multiuser MIMO channels," IEEE Trans. Sig. Proc., vol. 52, pp. 461–471, Feb. 2004. K. K. Wong, R. D. Murch, and K. B. Letaief, "A joint channel diagonalization for multiuser MIMO antenna systems," IEEE Trans. Wireless Comm., vol. 2, pp. 773–786, Jul 2003. L. U. Choi and R. D. Murch, "A transmit preprocessing technique for multiuser MIMO systems using a decomposition approach," IEEE Trans. Wireless Comm., vol. 3, pp. 20–24, Jan 2004. Z. Shen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization," accepted for publication in IEEE Trans. Sig. Proc., Sep. 2005. Z. Shen, R. Chen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Sum capacity of multiuser MIMO broadcast channels with block diagonalization," submitted to IEEE Trans. Wireless Comm., Oct. 2005, computed from the composite channel $\left[\mathbf{H}_e^{(m)}, \mathbf{H}_c^{(m)}\right]$ (rather than $\mathbf{H}_e^{(m)}$). So, the current DIDO system chooses the precoder such that

$$H_w \overset{\Delta}{=} \begin{bmatrix} \mathbf{H}_e^{(1)}[k] & \mathbf{H}_c^{(1)}[k] \\ \mathbf{H}_e^{(1)*}[K-k] & \mathbf{H}_e^{(1)*}[K-k] \\ \mathbf{H}_e^{(2)}[k] & \mathbf{H}_c^{(2)}[k] \\ \mathbf{H}_e^{(2)*}[K-k] & \mathbf{H}_e^{(2)*}[K-k] \end{bmatrix} \mathbf{W} = \begin{bmatrix} \alpha_{1,1} & 0 & 0 & 0 \\ 0 & \alpha_{1,2} & 0 & 0 \\ 0 & 0 & \alpha_{2,1} & 0 \\ 0 & 0 & 0 & \alpha_{2,2} \end{bmatrix} \Delta \begin{bmatrix} H_w^{(1,1)} & H_w^{(1,2)} \\ H_w^{(2,1)} & H_w^{(2,2)} \end{bmatrix} \tag{4}$$

where $\alpha_{i,j}$ are constants and $H_w^{(i,j)} \in C^{2\times 2}$. This method is beneficial because using this precoder, it is possible to keep other aspects of the DIDO precoder the same as before, since the effects of I/Q gain and phase imbalance are completely cancelled at the transmitter.

[0163] It is also possible to design DIDO precoders that pre-cancel inter-user interference, without pre-cancelling ICI due to IQ imbalance. With this approach, the receiver (instead of the transmitter) compensates for the IQ imbalance by employing one of the receive filters described below. Then, the pre-coding design criterion in (4) can be modified as

$$H_w \overset{\Delta}{=} \begin{bmatrix} \mathbf{H}_c^{(1)}[k] & \mathbf{H}_c^{(1)}[k] \\ \mathbf{H}_c^{(1)*}[K-k] & \mathbf{H}_c^{(1)*}[K-k] \\ \mathbf{H}_c^{(2)}[k] & \mathbf{H}_c^{(2)}[k] \\ \mathbf{H}_c^{(2)*}[K-k] & \mathbf{H}_c^{(2)*}[K-k] \end{bmatrix} W = \begin{bmatrix} \alpha_{1,1} & \alpha_{1,2} & 0 & 0 \\ \alpha_{2,1} & \alpha_{2,2} & 0 & 0 \\ 0 & 0 & \alpha_{3,3} & \alpha_{3,4} \\ 0 & 0 & \alpha_{4,3} & \alpha_{4,4} \end{bmatrix} \Delta \begin{bmatrix} H_w^{(1,1)} & H_w^{(1,2)} \\ H_w^{(2,1)} & H_w^{(2,2)} \end{bmatrix}$$

(5)

$$\bar{\mathbf{x}}_1[k] = \begin{bmatrix} H_w^{(1,1)} & H_w^{(1,2)} \end{bmatrix} \begin{bmatrix} \bar{\mathbf{s}}_1[k] \\ \bar{\mathbf{s}}_2[k] \end{bmatrix} \quad (6)$$

and

$$\bar{\mathbf{x}}_2[k] = \begin{bmatrix} H_w^{(2,1)} & H_w^{(2,2)} \end{bmatrix} \begin{bmatrix} \bar{\mathbf{s}}_1[k] \\ \bar{\mathbf{s}}_2[k] \end{bmatrix} \quad (7)$$

[0164] where $\bar{\mathbf{s}}_m[k] = [\bar{s}_m[k], \bar{s}_m^*[K-k]]^T$ for the *m*-th transmit symbol and $\bar{\mathbf{x}}_m[k] = [\bar{x}_m[k], \bar{x}_m^*[K-k]]^T$ is the receive symbol vector for user *m*.

[0165] At the receive side, to estimate the transmit symbol vector $\bar{s}_m[k]$, user *m* employs ZF filter and the estimated symbol vector is given by

$$\hat{\mathbf{s}}_m^{(ZF)}[k] = [(H_w^{(m,m)\dagger} H_w^{(m,m)})^{-1} H_w^{(m,m)\dagger}] \bar{\mathbf{x}}_m[k] \quad (8)$$

[0166] While the ZF filter is the easiest to understand, the receiver may apply any number of other filters known to those skilled in the art. One popular choice is the MMSE filter where

$$\hat{\mathbf{s}}_m^{(MMSE)}[k] = (H_w^{(m,m)\dagger} + \rho^{-1} I)^{-1} H_w^{(m,m)} H_w^{(m,m)\dagger} \bar{\mathbf{x}}_m[k] \quad (9)$$

and ρ is the signal-to-noise ratio. Alternatively, the receiver may perform a maximum likelihood symbol detection (or sphere decoder or iterative variation). For example, the first user might use the ML receiver and solve the following optimization

$$\hat{\mathbf{s}}_m^{(ML)}[k] = \arg \min_{s_1, s_2 \in S} \left\| \bar{\mathbf{y}}_1[k] - \begin{bmatrix} H_w^{(1,1)} & H_w^{(1,2)} \end{bmatrix} \begin{bmatrix} \mathbf{s}_1[k] \\ \mathbf{s}_2[k] \end{bmatrix} \right\| \quad (10)$$

where ***S*** is the set of all possible vectors **s** and depends on the constellation size. The ML receiver gives better performance at the expense of requiring more complexity at the receiver. A similar set of equations applies for the second user.

[0167] Note that $H_w^{(1,2)}$ and $H_w^{(2,1)}$ in (6) and (7) are assumed to have zero entries. This assumption holds only if the transmit precoder is able to cancel completely the inter-user interference as for the criterion in (4). Similarly, $H_w^{(1,1)}$ and $H_w^{(2,2)}$ are diagonal matrices only if the transmit precoder is able to cancel completely the inter-carrier interference (i.e., from the mirror tones).

[0168] **Figure 13** illustrates one embodiment of a framework for DIDO-OFDM systems with I/Q compensation including IQ-DIDO precoder 1302 within a Base Station (BS), a transmission channel 1304, channel estimation logic 1306 within a user device, and a ZF, MMSE or ML receiver 1308. The channel estimation logic 1306 estimates the channels $H_c^{(m)}$ and $H_c^{(m)}$ via training symbols and feedbacks these estimates to the precoder 1302 within the AP. The BS computes the DIDO precoder weights (matrix **W**) to pre-cancel the interference due to I/Q gain and phase imbalance as well as inter-

user interference and transmits the data to the users through the wireless channel 1304. User device *m* employs the ZF, MMSE or ML receiver 1308, by exploiting the channel estimates provided by the unit 1304, to cancel residual interference and demodulates the data.

[0169] The following three embodiments may be employed to implement this I/Q compensation algorithm:

[0170] **Method 1 - TX compensation**: In this embodiment, the transmitter calculates the pre-coding matrix according to the criterion in (4). At the receiver, the user devices employ a "simplified" ZF receiver, where $H_w^{(1,1)}$ and $H_w^{(2,2)}$ are assumed to be diagonal matrices. Hence, equation (8) simplifies as

$$\hat{\mathbf{s}}_m[k] = \begin{bmatrix} 1/\alpha_{m,1} & 0 \\ 0 & 1/\alpha_{m,2} \end{bmatrix} \bar{\mathbf{x}}_m[k]. \quad (10)$$

[0171] **Method 2 - RX compensation**: In this embodiment, the transmitter calculates the pre-coding matrix based on the conventional BD method described in R. Chen, R. W. Heath, and J. G. Andrews, "Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers," accepted to IEEE Trans. on Signal Processing, 2005, without canceling inter-carrier and inter-user interference as for the criterion in (4). With this method, the pre-coding matrix in (2) and (3) simplifies as

$$\mathbf{W} = \begin{bmatrix} w_{1,1}[k] & 0 & w_{1,2}[k] & 0 \\ 0 & w_{1,1}^*[K-k] & 0 & w_{1,2}^*[K-k] \\ w_{2,1}[k] & 0 & w_{2,2}[k] & 0 \\ 0 & w_{2,1}^*[K-k] & 0 & w_{2,2}^*[K-k] \end{bmatrix}. \quad (12)$$

[0172] At the receiver, the user devices employ a ZF filter as in (8). Note that this method does not pre-cancel the interference at the transmitter as in the method 1 above. Hence, it cancels the inter-carrier interference at the

receiver, but it is not able to cancel the inter-user interference. Moreover, in method 2 the users only need to feedback the vector $H_e^{(m)}$ for the transmitter to compute the DIDO precoder, as opposed to method 1 that requires feedback of both $H_e^{(ra)}$ and $H_c^{(m)}$. Therefore, method 2 is particularly suitable for DIDO systems with low rate feedback channels. On the other hand, method 2 requires slightly higher computational complexity at the user device to compute the ZF receiver in (8) rather than (11).

[0173] **Method 3 - TX-RX compensation**: In one embodiment, the two methods described above are combined. The transmitter calculates the pre-coding matrix as in (4) and the receivers estimate the transmit symbols according to (8).

[0174] I/Q imbalance, whether phase imbalance, gain imbalance, or delay imbalance, creates a deleterious degradation in signal quality in wireless communication systems. For this reason, circuit hardware in the past was designed to have very low imbalance. As described above, however, it is possible to correct this problem using digital signal processing in the form of transmit pre-coding and/or a special receiver. One embodiment of the invention comprises a system with several new functional units, each of which is important for the implementation of I/Q correction in an OFDM communication system or a DIDO-OFDM communication system.

[0175] One embodiment of the invention uses pre-coding based on channel state information to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch) in an OFDM system. As illustrated in **Figure 11**, a DIDO transmitter according to this embodiment includes a user selector unit 1102, a plurality of coding modulation units 1104, a corresponding plurality of mapping units 1106, a DIDO IQ-aware precoding unit 1108, a plurality of RF

transmitter units 1114, a user feedback unit 1112 and a DIDO configurator unit 1110.

[0176] The user selector unit 1102 selects data associated with a plurality of users U_1-U_M, based on the feedback information obtained by the feedback unit 1112, and provides this information each of the plurality of coding modulation units 1104. Each coding modulation unit 1104 encodes and modulates the information bits of each user and send them to the mapping unit 1106. The mapping unit 1106 maps the input bits to complex symbols and sends the results to the DIDO IQ-aware precoding unit 1108. The DIDO IQ-aware precoding unit 1108 exploits the channel state information obtained by the feedback unit 1112 from the users to compute the DIDO IQ-aware precoding weights and precoding the input symbols obtained from the mapping units 1106. Each of the precoded data streams is sent by the DIDO IQ-aware precoding unit 1108 to the OFDM unit 1115 that computes the IFFT and adds the cyclic prefix. This information is sent to the D/A unit 1116 that operates the digital to analog conversion and send it to the RF unit 1114. The RF unit 1114 upconverts the baseband signal to intermediate/radio frequency and send it to the transmit antenna.

[0177] The precoder operates on the regular and mirror tones together for the purpose of compensating for I/Q imbalance. Any number of precoder design criteria may be used including ZF, MMSE, or weighted MMSE design. In a preferred embodiment, the precoder completely removes the ICI due to I/Q mismatch thus resulting in the receiver not having to perform any additional compensation.

[0178] In one embodiment, the precoder uses a block diagonalization criterion to completely cancel inter-user interference while not completely canceling the I/Q effects for each user, requiring additional receiver processing. In another embodiment, the precoder uses a zero-forcing criterion

to completely cancel both inter-user interference and ICI due to I/Q imbalance. This embodiment can use a conventional DIDO-OFDM processor at the receiver.

[0179] One embodiment of the invention uses pre-coding based on channel state information to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch) in a DIDO-OFDM system and each user employs an IQ-aware DIDO receiver. As illustrated in **Figure 12**, in one embodiment of the invention, a system including the receiver 1202 includes a plurality of RF units 1208, a corresponding plurality of A/D units 1210, an IQ-aware channel estimator unit 1204 and a DIDO feedback generator unit 1206.

[0180] The RF units 1208 receive signals transmitted from the DIDO transmitter units 1114, downconverts the signals to baseband and provide the downconverted signals to the A/D units 1210. The A/D units 1210 then convert the signal from analog to digital and send it to the OFDM units 1213. The OFDM units 1213 remove the cyclic prefix and operates the FFT to report the signal to the frequency domain. During the training period the OFDM units 1213 send the output to the IQ-aware channel estimate unit 1204 that computes the channel estimates in the frequency domain. Alternatively, the channel estimates can be computed in the time domain. During the data period the OFDM units 1213 send the output to the IQ-aware receiver unit 1202. The IQ-aware receiver unit 1202 computes the IQ receiver and demodulates/decodes the signal to obtain the data 1214. The IQ-aware channel estimate unit 1204 sends the channel estimates to the DIDO feedback generator unit 1206 that may quantize the channel estimates and send it back to the transmitter via the feedback control channel 1112.

[0181] The receiver 1202 illustrated in **Figure 12** may operate under any number of criteria known to those skilled in the art including ZF, MMSE, maximum likelihood, or MAP receiver. In one preferred embodiment, the

receiver uses an MMSE filter to cancel the ICI caused by IQ imbalance on the mirror tones. In another preferred embodiment, the receiver uses a nonlinear detector like a maximum likelihood search to jointly detect the symbols on the mirror tones. This method has improved performance at the expense of higher complexity.

[0182] In one embodiment, an IQ-aware channel estimator 1204 is used to determine the receiver coefficients to remove ICI. Consequently we claim a DIDO-OFDM system that uses pre-coding based on channel state information to cancel inter-carrier interference (ICI) from mirror tones (due to I/Q mismatch), an IQ-aware DIDO receiver, and an IQ-aware channel estimator. The channel estimator may use a conventional training signal or may use specially constructed training signals sent on the inphase and quadrature signals. Any number of estimation algorithms may be implemented including least squares, MMSE, or maximum likelihood. The IQ-aware channel estimator provides an input for the IQ-aware receiver.

[0183] Channel state information can be provided to the station through channel reciprocity or through a feedback channel. One embodiment of the invention comprises a DIDO-OFDM system, with I/Q-aware precoder, with an I/Q-aware feedback channel for conveying channel state information from the user terminals to the station. The feedback channel may be a physical or logical control channel. It may be dedicated or shared, as in a random access channel. The feedback information may be generated using a DIDO feedback generator at the user terminal, which we also claim. The DIDO feedback generator takes as an input the output of the I/Q aware channel estimator. It may quantize the channel coefficients or may use any number of limited feedback algorithms known in the art.

[0184] The allocation of users, modulation and coding rate, mapping to space-time-frequency code slots may change depending on the results of the

DIDO feedback generator. Thus, one embodiment comprises an IQ-aware DIDO configurator that uses an IQ-aware channel estimate from one or more users to configure the DIDO IQ-aware precoder, choose the modulation rate, coding rate, subset of users allowed to transmit, and their mappings to space-time-frequency code slots.

[0185] To evaluate the performance of the proposed compensation methods, three DIDO 2 x 2 systems will be compared:

1. With I/Q mismatch: transmit over all the tones (except DC and edge tones), without compensation for I/Q mismatch;
2. With I/Q compensation: transmit over all the tones and compensate for I/Q mismatch by using the "method 1" described above;
3. Ideal: transmit data only over the odd tones to avoid inter-user and inter-carrier (i.e., from the mirror tones) interference caused to I/Q mismatch.

[0186] Hereafter, results obtained from measurements with the DIDO-OFDM prototype in real propagation scenarios are presented. **Figure 14** depicts the 64-QAM constellations obtained from the three systems described above. These constellations are obtained with the same users' locations and fixed average signal-to-noise ratio (~45 dB). The first constellation 1401 is very noisy due to interference from the mirror tones caused by I/Q imbalance. The second constellation 1402 shows some improvements due to I/Q compensations. Note that the second constellation 1402 is not as clean as the ideal case shown as constellation 1403 due to possible phase noise that yields inter-carrier interference (ICI).

[0187] **Figure 15** shows the average SER (Symbol Error Rate) 1501 and per-user goodput 1502 performance of DIDO 2×2 systems with 64-QAM and coding rate 3/4, with and without I/Q mismatch. The OFDM bandwidth is 250

KHz, with 64 tones and cyclic prefix length $L_{cp}=4$. Since in the ideal case we transmit data only over a subset of tones, SER and goodput performance is evaluated as a function of the *average per-tone transmit power* (rather than total transmit power) to guarantee a fair comparison across different cases. Moreover, in the following results, we use normalized values of transmit power (expressed in decibel), since our goal here is to compare the relative (rather than absolute) performance of different schemes. Figure 15 shows that in presence of I/Q imbalance the SER saturates, without reaching the target SER (~ 10^{-2}), consistently to the results reported in A. Tarighat and A. H. Sayed, "MIMO OFDM receivers for systems with IQ imbalances," IEEE Trans. Sig. Proc., vol. 53, pp. 3583–3596, Sep. 2005. This saturation effect is due to the fact that both signal and interference (from the mirror tones) power increase as the TX power increases. Through the proposed I/Q compensation method, however, it is possible to cancel the interference and obtain better SER performance. Note that the slight increase in SER at high SNR is due to amplitude saturation effects in the DAC, due to the larger transmit power required for 64-QAM modulations.

[0188] Moreover, observe that the SER performance with I/Q compensation is very close to the ideal case. The 2 dB gap in TX power between these two cases is due to possible phase noise that yields additional interference between adjacent OFDM tones. Finally, the goodput curves 1502 show that it is possible to transmit twice as much data when the I/Q method is applied compared to the ideal case, since we use all the data tones rather than only the odd tones (as for the ideal case).

[0189] **Figure 16** graphs the SER performance of different QAM constellations with and without I/Q compensation. We observe that, in this embodiment, the proposed method is particularly beneficial for 64-QAM constellations. For 4-QAM and 16-QAM the method for I/Q compensation

yields worse performance than the case with I/Q mismatch, possibly because the proposed method requires larger power to enable both data transmission and interference cancellation from the mirror tones. Moreover, 4-QAM and 16-QAM are not as affected by I/Q mismatch as 64-QAM due to the larger minimum distance between constellation points. See A. Tarighat, R. Bagheri, and A. H. Sayed, "Compensation schemes and performance analysis of IQ imbalances in OFDM receivers," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 53, pp. 3257–3268, Aug. 2005. This can be also observed in **Figure 16** by comparing the I/Q mismatch against the ideal case for 4-QAM and 16-QAM. Hence, the additional power required by the DIDO precoder with interference cancellation (from the mirror tones) does not justify the small benefit of the I/Q compensation for the cases of 4-QAM and 16-QAM. Note that this issue may be fixed by employing the methods 2 and 3 for I/Q compensation described above.

[0190] Finally, the relative SER performance of the three methods described above is measured in different propagation conditions. For reference, also described is the SER performance in presence of I/Q mismatch. **Figure 17** depicts the SER measured for a DIDO 2 x 2 system with 64-QAM at carrier frequency of 450.5 MHz and bandwidth of 250 KHz, at two different users' locations. In Location 1 the users are ~6λ from the BS in different rooms and NLOS (Non-Line of Sight)) conditions. In Location 2 the users are ~λ from the BS in LOS (Line of Sight).

[0191] **Figure 17** shows that all three compensation methods always outperform the case of no compensation. Moreover, it should be noted that method 3 outperforms the other two compensation methods in any channel scenario. The relative performance of method 1 and 2 depends on the propagation conditions. It is observed through practical measurement campaigns that method 1 generally outperforms method 2, since it pre-

cancels (at the transmitter) the inter-user interference caused by I/Q imbalance. When this inter-user interference is minimal, method 2 may outperform method 1 as illustrated in graph 1702 of **Figure 17**, since it does not suffer from power loss due to the I/Q compensation precoder.

[0192] So far, different methods have been compared by considering only a limited set of propagation scenarios as in **Figure 17**. Hereafter, the relative performance of these methods in ideal i.i.d.(independent and identically-distributed) channels is measured. DIDO-OFDM systems are simulated with I/Q phase and gain imbalance at the transmit and receive sides. **Figure 18** shows the performance of the proposed methods with only gain imbalance at the transmit side (i.e., with 0.8 gain on the I rail of the first transmit chain and gain 1 on the other rails). It is observed that method 3 outperforms all the other methods. Also, method 1 performs better than method 2 in i.i.d. channels, as opposed to the results obtained in Location 2 in graph 1702 of **Figure 17**.

[0193] Thus, given the three novel methods to compensate for I/Q imbalance in DIDO-OFDM systems described above, Method 3 outperforms the other proposed compensation methods. In systems with low rate feedback channels, method 2 can be used to reduce the amount of feedback required for the DIDO precoder, at the expense of worse SER performance.

II. Adaptive DIDO Transmission Scheme

[0194] Another embodiment of a system and method to enhance the performance of distributed-input distributed-output (DIDO) systems will now be described. This method dynamically allocates the wireless resources to different user devices, by tracking the changing channel conditions, to increase throughput while satisfying certain target error rate. The user devices estimate the channel quality and feedback it to the Base Station (BS);

the Base Station processes the channel quality obtained from the user devices to select the best set of user devices, DIDO scheme, modulation/coding scheme (MCS) and array configuration for the next transmission; the Base Station transmits parallel data to multiple user devices via pre-coding and the signals are demodulated at the receiver.

[0195] A system that efficiently allocates resources for a DIDO wireless link is also described. The system includes a DIDO Base Station with a DIDO configurator, which processes feedback received from the users to select the best set of users, DIDO scheme, modulation/coding scheme (MCS) and array configuration for the next transmission; a receiver in a DIDO system that measures the channel and other relevant parameters to generate a DIDO feedback signal; and a DIDO feedback control channel for conveying feedback information from users to the Base Station.

[0196] As described in detail below, some of the significant features of this embodiment of the invention include, but are not limited to:

[0197] Techniques to adaptively select number of users, DIDO transmission schemes (i.e., antenna selection or multiplexing), modulation/coding scheme (MCS) and array configurations based on the channel quality information, to minimize SER or maximize per-user or downlink spectral efficiency;

[0198] Techniques to define sets of DIDO *transmission modes* as combinations of DIDO schemes and MCSs;

[0199] Techniques to assign different DIDO modes to different time slots, OFDM tones and DIDO substreams, depending on the channel conditions;

[0200] Techniques to dynamically assign different DIDO modes to different users based on their channel quality;

[0201] Criterion to enable adaptive DIDO switching based on link quality metrics computed in the time, frequency and space domains;

[0202] Criterion to enable adaptive DIDO switching based on lookup tables.

[0203] A DIDO system with a DIDO configurator at the Base Station as in **Figure 19** to adaptively select the number of users, DIDO transmission schemes (i.e., antenna selection or multiplexing), modulation/coding scheme (MCS) and array configurations based on the channel quality information, to minimize SER or maximize per user or downlink spectral efficiency;

[0204] A DIDO system with a DIDO configurator at the Base Station and a DIDO feedback generator at each user device as in **Figure 20**, which uses the estimated channel state and/or other parameters like the estimated SNR at the receiver to generate a feedback message to be input into the DIDO configurator.

[0205] A DIDO system with a DIDO configurator at the Base Station, DIDO feedback generator, and a DIDO feedback control channel for conveying DIDO-specific configuration information from the users to the Base Station.

a. Background

[0206] In multiple-input multiple-output (MIMO) systems, diversity schemes such as orthogonal space-time block codes (OSTBC) (See V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Spacetime block codes from orthogonal designs," IEEE Trans. Info. Th., vol. 45, pp. 1456–467, Jul. 1999) or antenna selection (See R. W. Heath Jr., S. Sandhu, and A. J. Paulraj, "Antenna selection for spatial multiplexing systems with linear receivers," IEEE Trans. Comm., vol. 5, pp. 142–144, Apr. 2001) are conceived to combat channel fading, providing increased link robustness that translates in better coverage. On the other hand, spatial multiplexing (SM) enables transmission of multiple parallel data streams as a means to enhance systems throughput. See G. J.

Foschini, G.D. Golden, R. A. Valenzuela, and P. W. Wolniansky, "Simplified processing for high spectral efficiency wireless communication employing multielement arrays," IEEE Jour. Select. Areas in Comm., vol. 17, no. 11, pp. 1841 – 1852, Nov. 1999. These benefits can be simultaneously achieved in MIMO systems, according to the theoretical diversity/multiplexing tradeoffs derived in L. Zheng and D.N. C. Tse, "Diversity and multiplexing: a fundamental tradeoff in multiple antenna channels," IEEE Trans. Info. Th., vol. 49, no. 5, pp. 1073–1096, May 2003. One practical implementation is to adaptively switch between diversity and multiplexing transmission schemes, by tracking the changing channel conditions.

[0207] A number of adaptive MIMO transmission techniques have been proposed thus far. The diversity/multiplexing switching method in R. W. Heath and A. J. Paulraj, "Switching between diversity and multiplexing in MIMO systems," IEEE Trans. Comm., vol. 53, no. 6, pp. 962 – 968, Jun. 2005, was designed to improve BER (Bit Error Rate) for fixed rate transmission, based on instantaneous channel quality information. Alternatively, statistical channel information can be employed to enable adaptation as in S. Catreux, V. Erceg, D. Gesbert, and R. W. Heath. Jr., "Adaptive modulation and MIMO coding for broadband wireless data networks," IEEE Comm. Mag., vol. 2, pp. 108–115, June 2002 ("Catreux"), resulting in reduced feedback overhead and number of control messages. The adaptive transmission algorithm in Catreux was designed to enhance spectral efficiency for predefined target error rate in orthogonal frequency division multiplexing (OFDM) systems, based on channel time/frequency selectivity indicators. Similar low feedback adaptive approaches have been proposed for narrowband systems, exploiting the channel spatial selectivity to switch between diversity schemes and spatial multiplexing. See, e.g., A. Forenza, M. R. McKay, A. Pandharipande, R. W. Heath. Jr., and I. B. Collings, "Adaptive MIMO transmission for exploiting the capacity of spatially correlated

channels," accepted to the IEEE Trans. on Veh. Tech., Mar. 2007; M. R. McKay, I. B. Collings, A. Forenza, and R. W. Heath. Jr., "Multiplexing/beamforming switching for coded MIMO in spatially correlated Rayleigh channels," accepted to the IEEE Trans. on Veh. Tech., Dec. 2007; A. Forenza, M. R. McKay, R. W. Heath. Jr., and I. B. Collings, "Switching between OSTBC and spatial multiplexing with linear receivers in spatially correlated MIMO channels," Proc. IEEE Veh. Technol. Conf., vol. 3, pp. 1387–1391, May 2006; M. R. McKay, I. B. Collings, A. Forenza, and R. W. Heath Jr., "A throughput-based adaptive MIMO BICM approach for spatially correlated channels," to appear in Proc. IEEE ICC, June 2006

[0208] In this document, we extend the scope of the work presented in various prior publications to DIDO-OFDM systems. See, e.g., R. W. Heath and A. J. Paulraj, "Switching between diversity and multiplexing in MIMO systems," IEEE Trans. Comm., vol. 53, no. 6, pp. 962 – 968, Jun. 2005.S. Catreux, V. Erceg, D. Gesbert, and R. W. Heath Jr., "Adaptive modulation and MIMO coding for broadband wireless data networks," IEEE Comm. Mag., vol. 2, pp. 108–115, June 2002; A. Forenza, M. R. McKay, A. Pandharipande, R. W. Heath Jr., and I. B. Collings, "Adaptive MIMO transmission for exploiting the capacity of spatially correlated channels," IEEE Trans. on Veh. Tech., vol.56, n.2, pp.619-630, Mar. 2007. M. R. McKay, I. B. Collings, A. Forenza, and R. W. Heath Jr., "Multiplexing/beamforming switching for coded MIMO in spatially correlated Rayleigh channels," accepted to the IEEE Trans. on Veh. Tech., Dec. 2007; A. Forenza, M. R. McKay, R. W. Heath Jr., and I. B. Collings, "Switching between OSTBC and spatial multiplexing with linear receivers in spatially correlated MIMO channels," Proc. IEEE Veh. Technol. Conf., vol. 3, pp. 1387–1391, May 2006. M. R. McKay, I. B. Collings, A. Forenza, and R. W. Heath Jr., "A throughput-based adaptive MIMO BICM approach for spatially correlated channels," to appear in Proc. IEEE ICC, June 2006.

[0209] A novel adaptive DIDO transmission strategy is described herein that switches between different numbers of users, numbers of transmit antennas and transmission schemes based on channel quality information as a means to improve the system performance. Note that schemes that adaptively select the users in multiuser MIMO systems were already proposed in M. Sharif and B. Hassibi, "On the capacity of MIMO broadcast channel with partial side information," IEEE Trans. Info. Th., vol. 51, p. 506522, Feb. 2005; and W. Choi, A. Forenza, J. G. Andrews, and R. W. Heath Jr., "Opportunistic space division multiple access with beam selection," to appear in IEEE Trans. on Communications. The opportunistic space division multiple access (OSDMA) schemes in these publications, however, are designed to maximize the sum capacity by exploiting multi-user diversity and they achieve only a fraction of the theoretical capacity of dirty paper codes, since the interference is not completely pre-canceled at the transmitter. In the DIDO transmission algorithm described herein block diagonalization is employed to pre-cancel inter-user interference. The proposed adaptive transmission strategy, however, can be applied to any DIDO system, independently on the type of pre-coding technique.

[0210] The present patent application describes an extension of the embodiments of the invention described above and in the Prior Application, including, but not limited to the following additional features:

1. The training symbols of the Prior Application for channel estimation can be employed by the wireless client devices to evaluate the link-quality metrics in the adaptive DIDO scheme;

2. The base station receives signal characterization data from the client devices as described in the Prior Application. In the current embodiment, the signal characterization data is defined as link-quality metric used to enable adaptation;

3. The Prior Application describes a mechanism to select the number of transmit antennas and users as well as defines throughput allocation. Moreover, different levels of throughput can be dynamically assigned to different clients as in the Prior Application. The current embodiment of the invention defines novel criteria related to this selection and throughput allocation.

b. Embodiments of the Invention

[0211] The goal of the proposed adaptive DIDO technique is to enhance per-user or downlink spectral efficiency by dynamically allocating the wireless resource in time, frequency and space to different users in the system. The general adaptation criterion is to increase throughput while satisfying the target error rate. Depending on the propagation conditions, this adaptive algorithm can also be used to improve the link quality of the users (or coverage) via diversity schemes. The flowchart illustrated in **Figure 21** describes steps of the adaptive DIDO scheme.

[0212] The Base Station (BS) collects the channel state information (CSI) from all the users in 2102. From the received CSI, the BS computes the link quality metrics in time/frequency/space domains in 2104. These link quality metrics are used to select the users to be served in the next transmission as well as the transmission mode for each of the users in 2106. Note that the transmission modes consist of different combinations of modulation/coding and DIDO schemes. Finally, the BS transmits data to the users via DIDO precoding as in 2108.

[0213] At 2102, the Base Station collects the channel state information (CSI) from all the user devices. The CSI is used by the Base Station to determine the instantaneous or statistical channel quality for all the user devices at 2104. In DIDO-OFDM systems the channel quality (or link quality

metric) can be estimated in the time, frequency and space domains. Then, at 2106, the Base Station uses the link quality metric to determine the best subset of users and *transmission mode* for the current propagation conditions. A set of DIDO transmission modes is defined as combinations of DIDO schemes (i.e., antenna selection or multiplexing), modulation/coding schemes (MCSs) and array configuration. At 2108, data is transmitted to user devices using the selected number of users and transmission modes.

[0214] The mode selection is enabled by lookup tables (LUTs) pre-computed based on error rate performance of DIDO systems in different propagation environments. These LUTs map channel quality information into error rate performance. To construct the LUTs, the error rate performance of DIDO systems is evaluated in different propagation scenarios as a function of the SNR. From the error rate curves, it is possible to compute the minimum SNR required to achieve certain pre-defined target error rate. We define this SNR requirement as *SNR threshold*. Then, the SNR thresholds are evaluated in different propagation scenarios and for different DIDO transmission modes and stored in the LUTs. For example, the SER results in Figures 24 and 26 can be used to construct the LUTs. Then, from the LUTs, the Base Station selects the transmission modes for the active users that increase throughput while satisfying predefined target error rate. Finally, the Base Station transmits data to the selected users via DIDO pre-coding. Note that different DIDO modes can be assigned to different time slots, OFDM tones and DIDO substreams such that the adaptation may occur in time, frequency and space domains.

[0215] One embodiment of a system employing DIDO adaptation is illustrated in **Figures 19-20**. Several new functional units are introduced to enable implementation of the proposed DIDO adaptation algorithms. Specifically, in one embodiment, a DIDO configurator 1910 performs a plurality of functions including selecting the number of users, DIDO

transmission schemes (i.e., antenna selection or multiplexing), modulation/coding scheme (MCS), and array configurations based on the channel quality information 1912 provided by user devices.

[0216] The user selector unit 1902 selects data associated with a plurality of users U_1-U_M, based on the feedback information obtained by the DIDO configurator 1910, and provides this information each of the plurality of coding modulation units 1904. Each coding modulation unit 1904 encodes and modulates the information bits of each user and sends them to the mapping unit 1906. The mapping unit 1906 maps the input bits to complex symbols and sends it to the precoding unit 1908. Both the coding modulation units 1904 and the mapping unit 1906 exploit the information obtained from the DIDO configurator unit 1910 to choose the type of modulation/coding scheme to employ for each user. This information is computed by the DIDO configurator unit 1910 by exploiting the channel quality information of each of the users as provided by the feedback unit 1912. The DIDO precoding unit 1908 exploits the information obtained by the DIDO configurator unit 1910 to compute the DIDO precoding weights and precoding the input symbols obtained from the mapping units 1906. Each of the precoded data streams are sent by the DIDO precoding unit 1908 to the OFDM unit 1915 that computes the IFFT and adds the cyclic prefix. This information is sent to the D/A unit 1916 that operates the digital to analog conversion and sends the resulting analog signal to the RF unit 1914. The RF unit 1914 upconverts the baseband signal to intermediate/radio frequency and send it to the transmit antenna.

[0217] The RF units 2008 of each client device receive signals transmitted from the DIDO transmitter units 1914, downconverts the signals to baseband and provide the downconverted signals to the A/D units 2010. The A/D units 2010 then convert the signal from analog to digital and send it to the OFDM units 2013. The OFDM units 2013 remove the cyclic prefix and carries out the

FFT to report the signal to the frequency domain. During the training period the OFDM units 2013 send the output to the channel estimate unit 2004 that computes the channel estimates in the frequency domain. Alternatively, the channel estimates can be computed in the time domain. During the data period the OFDM units 2013 send the output to the receiver unit 2002 which demodulates/decodes the signal to obtain the data 2014. The channel estimate unit 2004 sends the channel estimates to the DIDO feedback generator unit 2006 that may quantize the channel estimates and send it back to the transmitter via the feedback control channel 1912.

[0218] The DIDO configurator 1910 may use information derived at the Base Station or, in a preferred embodiment, uses additionally the output of a DIDO Feedback Generator 2006 (see **Figure 20**), operating at each user device. The DIDO Feedback Generator 2006 uses the estimated channel state 2004 and/or other parameters like the estimated SNR at the receiver to generate a feedback message to be input into the DIDO Configurator 1910. The DIDO Feedback Generator 2006 may compress information at the receiver, may quantize information, and/or use some limited feedback strategies known in the art.

[0219] The DIDO Configurator 1910 may use information recovered from a DIDO Feedback Control Channel 1912. The DIDO Feedback Control Channel 1912 is a logical or physical control channel that is used to send the output of the DIDO Feedback Generator 2006 from the user to the Base Station. The control channel 1912 may be implemented in any number of ways known in the art and may be a logical or a physical control channel. As a physical channel it may comprise a dedicated time/frequency slot assigned to a user. It may also be a random access channel shared by all users. The control channel may be pre-assigned or it may be created by stealing bits in a predefined way from an existing control channel.

[0220] In the following discussion, results obtained through measurements with the DIDO-OFDM prototype are described in real propagation environments. These results demonstrate the potential gains achievable in adaptive DIDO systems. The performance of different order DIDO systems is presented initially, demonstrating that it is possible to increase the number of antennas/user to achieve larger downlink throughput. The DIDO performance as a function of user device's location is then described, demonstrating the need for tracking the changing channel conditions. Finally, the performance of DIDO systems employing diversity techniques is described.

i. Performance of Different Order DIDO Systems

[0221] The performance of different DIDO systems is evaluated with increasing number of transmit antennas $N = M$, where M is the number of users. The performance of the following systems is compared: SISO, DIDO 2 × 2, DIDO 4 × 4, DIDO 6 × 6 and DIDO 8 × 8. DIDO $N \times M$ refers to DIDO with N transmit antennas at the BS and M users.

[0222] **Figure 22** illustrates the transmit/receive antenna layout. The transmit antennas 2201 are placed in squared array configuration and the users are located around the transmit array. In **Figure 22**, T indicates the "transmit" antennas and U refers to the "user devices" 2202.

[0223] Different antenna subsets are active in the 8-element transmit array, depending on the value of N chosen for different measurements. For each DIDO order (N) the subset of antennas that covers the largest real estate for fixed size constraint of the 8-element array was chosen. This criterion is expected to enhance the spatial diversity for any given value of N.

[0224] **Figure 23** shows the array configurations for different DIDO orders that fit the available real estate (i.e., dashed line). The squared dashed box

has dimensions of 24"×24", corresponding to ~ λ × λ at the carrier frequency of 450 MHz.

[0225] Based on the comments related to **Figure 23** and with reference to **Figure 22**, the performance of each of the following systems will now be defined and compared:

SISO with T1 and U1 (2301)
DIDO 2 × 2 with T1,2 and U1,2 (2302)
DIDO 4 × 4 with T1,2,3,4 and U1,2,3,4 (2303)
DIDO 6 × 6 with T1,2,3,4,5,6 and U1,2,3,4,,5,6 (2304)
DIDO 8 × 8 with T1,2,3,4,5,6,7,8 and U1,2,3,4,5,6,7,8 (2305)

[0226] **Figure 24** shows the SER, BER, SE (Spectral Efficiency) and goodput performance as a function of the transmit (TX) power for the DIDO systems described above, with 4-QAM and FEC (Forward Error Correction) rate of 1/2. Observe that the SER and BER performance degrades for increasing values of *N*. This effect is due to two phenomena: for fixed TX power, the input power to the DIDO array is split between increasing number of users (or data streams); the spatial diversity decreases with increasing number of users in realistic (spatially correlated) DIDO channels.

[0227] To compare the relative performance of different order DIDO systems the target BER is fixed to 10^{-4} (this value may vary depending on the system) that corresponds approximately to SER= 10^{-2} as shown in **Figure 24**. We refer to the TX power values corresponding to this target as TX power thresholds (TPT). For any *N*, if the TX power is below the TPT, we assume it is not possible to transmit with DIDO order *N* and we need to switch to lower order DIDO. Also, in **Figure 24**, observe that the SE and goodput performance saturate when the TX power exceeds the TPTs for any value of *N*. From these results, an adaptive transmission strategy may be designed

that switches between different order DIDO to enhance SE or goodput for fixed predefined target error rate.

ii. Performance with Variable User Location

[0228] The goal of this experiment is to evaluate the DIDO performance for different users' location, via simulations in spatially correlated channels. DIDO 2 x 2 systems are considered with 4QAM and an FEC rate of 1/2. User 1 is at a broadside direction from the transmit array, whereas user 2 changes locations from broadside to endfire directions as illustrated in **Figure 25**. The transmit antennas are spaced ~λ/2 and separated ~2.5λ from the users.

[0229] **Figure 26** shows the SER and per-user SE results for different locations of user device 2. The user device's angles of arrival (AOAs) range between 0º and 90º, measured from the broadside direction of the transmit array. Observe that, as the user device's angular separation increases, the DIDO performance improves, due to larger diversity available in the DIDO channel. Also, at target SER= 10^{-2} there is a 10dB gap between the cases AOA2= 0º and AOA2= 90º. This result is consistent to the simulation results obtained in Figure 35 for an angle spread of 10º. Also, note that for the case of AOA1=AOA2= 0º there may be coupling effects between the two users (due to the proximity of their antennas) that may vary their performance from the simulated results in Figure 35.

iii. Preferred Scenario for DIDO 8 X 8

[0230] **Figure 24** illustrated that DIDO 8 X 8 yields a larger SE than lower order DIDO at the expense of higher TX power requirement. The goal of this analysis is to show there are cases where DIDO 8 x 8 outperforms DIDO 2 x 2, not only in terms of peak spectral efficiency (SE), but also in terms of TX power requirement (or TPT) to achieve that peak SE.

[0231] Note that, in i.i.d. (ideal) channels, there is ~6dB gap in TX power between the SE of DIDO 8 × 8 and DIDO 2 × 2. This gap is due to the fact that DIDO 8 × 8 splits the TX power across eight data streams, whereas DIDO 2 X 2 only between two streams. This result is shown via simulation in Figure 32.

[0232] In spatially correlated channels, however, the TPT is a function of the characteristics of the propagation environment (e.g., array orientation, user location, angle spread). For example, Figure 35 shows ~15dB gap for low angle spread between two different user device's locations. Similar results are presented in **Figure 26** of the present application.

[0233] Similarly to MIMO systems, the performance of DIDO systems degrades when the users are located at endfire directions from the TX array (due to lack of diversity). This effect has been observed through measurements with the current DIDO prototype. Hence, one way to show that DIDO 8 × 8 outperforms DIDO 2 × 2 is to place the users at endfire directions with respect to the DIDO 2 × 2 arrays. In this scenario, DIDO 8 × 8 outperforms DIDO 2 × 2 due to the higher diversity provided by the 8-antenna array.

[0234] In this analysis, consider the following systems:

[0235] System 1: DIDO 8 × 8 with 4-QAM (transmit 8 parallel data streams every time slot);

[0236] System 2: DIDO 2 X 2 with 64-QAM (transmit to users X and Y every 4 time slots). For this system we consider four combinations of TX and RX antenna locations: a) T1,T2 U1,2 (endfire direction); b) T3,T4 U3,4 (endfire direction); c) T5,T6 U5,6 (~30° from the endfire direction); d) T7,T8 U7,8 (NLOS (Non-Line of Sight));

[0237] System 3: DIDO 8 × 8 with 64-QAM; and

[0238] System 4: MISO 8 × 1 with 64-QAM (transmit to user X every 8 time slots).

[0239] For all these cases, an FEC rate of ¾ was used.

[0240] The users' locations are depicted in **Figure 27**.

[0241] In **Figure 28** the SER results show a ~15dB gap between Systems 2a and 2c due to different array orientations and user locations (similar to the simulation results in Figure 35). The first subplot in the second row shows the values of TX power for which the SE curves saturate (i.e. corresponding to BER 1e-4). We observe that System 1 yields larger per-user SE for lower TX power requirement (~ 5dB less) than System 2. Also, the benefits of DIDO 8 × 8 versus DIDO 2 ×2 are more evident for the DL (downlink) SE and DL goodput due to multiplexing gain of DIDO 8 × 8 over DIDO 2 × 2. System 4 has lower TX power requirement (8dB less) than System 1, due to the array gain of beamforming (i.e., MRC with MISO 8 × 1). But System 4 yields only 1/3 of per-user SE compared to System 1. System 2 performs worse than System 1 (i.e., yields lower SE for larger TX power requirement). Finally, System 3 yields much larger SE (due to larger order modulations) than System 1 for larger TX power requirement (~ 15dB).

[0242] From these results, the following conclusions may be drawn:

[0243] One channel scenario was identified for which DIDO 8 × 8 outperforms DIDO 2 × 2 (i.e., yields larger SE for lower TX power requirement);

[0244] In this channel scenario, DIDO 8 × 8 yields larger per user SE and DL SE than DIDO 2 × 2 and MISO 8 × 1; and

[0245] It is possible to further increase the performance of DIDO 8 × 8 by using higher order modulations (i.e., 64-QAM rather than 4-QAM) at the expense of larger TX power requirements (~15dB more).

iv. DIDO with Antenna Selection

[0246] Hereafter, we evaluate the benefit of the antenna selection algorithm described in R. Chen, R. W. Heath, and J. G. Andrews, "Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers," accepted to IEEE Trans. on Signal Processing, 2005. We present the results for one particular DIDO system with two users, 4-QAM and FEC rate of 1/2. The following systems are compared in **Figure 27**:

DIDO 2 × 2 with T1,2 and U1,2; and

DIDO 3 × 2 using antenna selection with T1,2,3 and U1,2.

[0247] The transmit antenna's and user device locations are the same as in **Figure 27**.

[0248] **Figure 29** shows that DIDO 3 ×2 with antenna selection may provide ~5dB gain compared to DIDO 2 × 2 systems (with no selection). Note that the channel is almost static (i.e., no Doppler), so the selection algorithms adapts to the path-loss and channel spatial correlation rather than the fast-fading. We should be seeing different gains in scenarios with high Doppler. Also, in this particular experiment it was observed that the antenna selection algorithm selects antennas 2 and 3 for transmission.

iv. SNR thresholds for the LUTs

[0249] In section [0171] we stated that the mode selection is enabled by LUTs. The LUTs can be pre-computed by evaluating the SNR thresholds to achieve certain predefined target error-rate performance for the DIDO transmission modes in different propagation environments. Hereafter, we

provide the performance of DIDO systems with and without antenna selection and variable number of users that can be used as guidelines to construct the LUTs. While Figures 24, 26, 28, 29 were derived from practical measurements with the DIDO prototype, the following Figures are obtained through simulations. The following BER results assume no FEC.

[0250] **Figure 30** shows the average BER performance of different DIDO precoding schemes in i.i.d. channels. The curve labeled as 'no selection' refers to the case when BD is employed. In the same figure the performance of antenna selection (ASel) is shown for different number of extra antennas (with respect to the number of users). It is possible to see that as the number of extra antennas increases, ASel provides better diversity gain (characterized by the slope of the BER curve in high SNR regime), resulting in better coverage. For example, if we fix the target BER to 10^{-2} (practical value for uncoded systems), the SNR gain provided by ASel increases with the number of antennas.

[0251] **Figure 31** shows the SNR gain of ASel as a function of the number of extra transmit antennas in i.i.d. channels. for different targets BER. It is possible to see that, just by adding 1 or 2 antennas, ASel yields significant SNR gains compared to BD. In the following sections, we will evaluate the performance of ASel only for the cases of 1 or 2 extra antennas and by fixing the target BER to 10^{-2} (for uncoded systems).

[0252] **Figure 32** depicts the SNR thresholds as a function of the number of users (M) for BD and ASel with 1 and 2 extra antennas in i.i.d. channels. We observe that the SNR thresholds increase with M due to the larger receive SNR requirement for larger number of users. Note that we assume fixed total transmit power (with variable number of transmit antennas) for any number of users. Moreover, Figure 32 shows that the gain due to antenna selection is constant for any number of users in i.i.d. channels.

Hereafter, we show the performance of DIDO systems in spatially correlated channels. We simulate each user's channel through the COST-259 spatial channel model described in X. Zhuang, F. W. Vook, K. L. Baum, T. A. Thomas, and M. Cudak, "Channel models for link and system level simulations," IEEE 802.16 Broadband Wireless Access Working Group, Sep. 2004.. We generate single-cluster for each user. As a case study, we assume NLOS channels, uniform linear array (ULA) at the transmitter, with element spacing of 0.5 lambda. For the case of 2-user system, we simulate the clusters with mean angles of arrival AOA1 and AOA2 for the first and second user, respectively. The AOAs are measured with respect to the broadside direction of the ULA. When more than two users are in the system, we generate the users' clusters with uniformly spaced mean AOAs in the range $[-\phi_m, \phi_m]$, where we define

$$\Phi_M = \frac{\Delta\phi(M-1)}{2} \qquad (13)$$

with K being the number of users and $\Delta\phi$ is the angular separation between the users' mean AOAs. Note that the angular range $[-\phi_m, \phi_m]$ is centered at the 0° angle, corresponding to the broadside direction of the ULA. Hereafter, we study the BER performance of DIDO systems as a function of the channel angle spread (AS) and angular separation between users, with BD and ASel transmission schemes and different numbers of users.

[0253] **Figure 33** depicts the BER versus per-user average SNR for two users located at the same angular direction (i.e., AOA1 = AOA2 = 0°, with respect to the broadside direction of the ULA), with different values of AS. It is possible to see that as the AS increases the BER performance improves and approaches the i.i.d. case. In fact, higher AS yields statistically less

overlapping between the eigenmodes of the two users and better performance of the BD precoder.

[0254] **Figure 34** shows similar results as Figure 33, but with higher angular separation between the users. We consider AOA1 = 0° and AOA2 = 90° (i.e., 90° angular separation). The best performance is now achieved in the low AS case. In fact, for the case of high angle separation, there is less overlapping between the users' eigenmodes when the angular spread is low. Interestingly, we observe that the BER performance in low AS is better than i.i.d. channels for the same reasons just mentioned.

[0255] Next, we compute the SNR thresholds, for target BER of 10^{-2} in different correlation scenarios. **Figure 35** plots the SNR thresholds as a function of the AS for different values of the mean AOAs of the users. For low users' angular separation reliable transmissions with reasonable SNR requirement (i.e., 18 dB) are possible only for channels characterized by high AS. On the other hand, when the users are spatially separated, less SNR is required to meet the same target BER.

[0256] **Figure 36** shows the SNR threshold for the case of five users. The users' mean AOAs are generated according to the definition in (13), with different values of angular separation $\Delta\phi$. We observe that for $\Delta\phi = 0^\circ$ and AS< 15°, BD performs poorly due to the small angle separation between users, and the target BER is not satisfied. For increasing AS the SNR requirement to meet the fixed target BER decreases. On the other end, for $\Delta\phi = 30^\circ$, the smallest SNR requirement is obtained at low AS, consistently to the results in Figure 35. As the AS increases, the SNR thresholds saturate to the one of i.i.d. channels. Note that_$\Delta\phi = 30^\circ$ with 5 users corresponds to the AOA range of [–60°, 60°], that is typical for base stations in cellular systems with 120° sectorized cells.

[0257] Next, we study the performance of ASel transmission scheme in spatially correlated channels. **Figure 37** compares the SNR threshold of BD and ASel, with 1 and 2 extra antennas, for two user case. We consider two different cases of angular separation between users: {AOA1 = 0°,AOA2 = 0°} and {AOA1 = 0°,AOA2 = 90°}. The curves for BD scheme (i.e., no antenna selection) are the same as in Figure 35. We observe that ASel yields 8 dB and 10 dB SNR gains with 1 and 2 extra antennas, respectively, for high AS. As the AS decreases, the gain due to ASel over BD becomes smaller due to the reduced number of degrees of freedom in the MIMO broadcast channel. Interestingly, for AS= 0° (i.e., close to LOS channels) and the case {AOA1 = 0°,AOA2 = 90°}, ASel does not provide any gain due to the luck of diversity in the space domain. **Figure 38** shows similar results as Figure 37, but for five user case.

[0258] We compute the SNR thresholds (assuming usual target BER of 10^{-2}) as a function of the number of users in the system (M), for both BD and ASel transmission schemes. The SNR thresholds correspond to the average SNR, such that the total transmit power is constant for any M. We assume maximum separation between the mean AOAs of each user's cluster within the azimuth range $[-\phi_m, \phi_m]$ = [–60°, 60°]. Then, the angular separation between users is $\Delta\phi = 120°/(M-1)$.

[0259] **Figure 39** shows the SNR thresholds for BD scheme with different values of AS. We observe that the lowest SNR requirement is obtained for AS= 0.1° (i.e., low angle spread) with relatively small number of users (i.e., $K \leq 20$), due to the large angular separation between users. For $M > 50$, however, the SNR requirement is way above 40 dB, since $\Delta\phi$ is very small, and BD is impractical. Moreover, for AS> 10° the SNR thresholds remain almost constant for any M, and the DIDO system in spatially correlated channels approaches the performance of i.i.d. channels.

[0260] To reduce the values of the SNR thresholds and improve the performance of the DIDO system we apply ASel transmission scheme. **Figure 40** depicts the SNR thresholds in spatially correlated channels with AS= 0.1° for BD and ASel with 1 and 2 extra antennas. For reference we report also the curves for the i.i.d. case shown in Figure 32.. It is possible to see that, for low number of users (i.e., $M \leq 10$), antenna selection does not help reducing the SNR requirement due to the lack of diversity in the DIDO broadcast channel. As the number of users increases, ASel benefits from multiuser diversity yielding SNR gains (i.e., 4 dB for $M = 20$). Moreover, for $M \leq 20$, the performance of ASel with 1 or 2 extra antennas in highly spatially correlated channels is the same.

[0261] We then compute the SNR thresholds for two more channel scenarios: AS= 5° in **Figure 41** and AS= 10° in **Figure 42.** Figure 41 shows that ASel yields SNR gains also for relatively small number of users (i.e., $M \leq 10$) as opposed to Figure 40, due to the larger angle spread. For AS= 10° the SNR thresholds reduce further and the gains due to ASel get higher, as reported in Figure 42.

[0262] Finally, we summarize the results presented so far for correlated channels. **Figure 43** and **Figure 44** show the SNR thresholds as a function of the number of users (M) and angle spread (AS) for BD and ASel schemes, with 1 and 2 extra antennas, respectively. Note that the case of AS= 30° corresponds actually to i.i.d. channels, and we used this value of AS in the plot only for graphical representation. We observe that, while BD is affected by the channel spatial correlation, ASel yields almost the same performance for any AS. Moreover, for AS= 0.1°, ASel performs similarly to BD for low M, whereas outperforms BD for large M (i.e., $M \geq 20$), due to multiuser diversity.

[0263] **Figure 49** compares the performance of different DIDO schemes in terms of SNR thresholds. The DIDO schemes considered are: BD, ASel, BD

with eigenmode selection (BD-ESel) and maximum ratio combining (MRC). Note that MRC, does not pre-cancel interference at the transmitter (unlike the other methods), but does provide larger gain in case the users are spatially separated. In Figure 49 we plot the SNR threshold for target BER= 10–2 for DIDO N × 2 systems when the two users are located at -30° and 30° from the broadside direction of the transmit array, respectively. We observe that for low AS the MRC scheme provides 3 dB gain compared to the other schemes since the users' spatial channels are well separated and the effect of inter-user interference is low. Note that the gain of MRC over DIDO N × 2 are due to array gain. For AS larger than 20° the QR-ASel scheme outperforms the other and yields about 10 dB gain compared to BD 2×2 with no selection. QR-ASel and BD-ESel provide about the same performance for any value of AS.

[0264] Described above is a novel adaptive transmission technique for DIDO systems. This method dynamically switches between DIDO transmission modes to different users to enhance throughput for fixed target error rate. The performance of different order DIDO systems was measured in different propagation conditions and it was observed that significant gains in throughput may be achieved by dynamically selecting the DIDO modes and number of users as a function of the propagation conditions.

III. Pre-compensation of Frequency and Phase Offset

a. Background

[0265] As previously described, wireless communication systems use carrier waves to convey information. These carrier waves are usually sinusoids that are amplitude and/or phase modulated in response to information to be transmitted. The nominal frequency of the sinusoid is known as the carrier frequency. To create this waveform, the transmitter synthesizes one or more sinusoids and uses upconversion to create a modulated signal riding on a sinusoid with the prescribed carrier frequency. This may be done through direct conversion where the signal is directly modulated on the carrier

or through multiple upconversion stages. To process this waveform, the receiver must demodulate the received RF signal and effectively remove the modulating carrier. This requires that the receiver synthesize one or more sinusoidal signals to reverse the process of modulation at the transmitter, known as downconversion. Unfortunately, the sinusoidal signals generated at the transmitter and receiver are derived from different reference oscillators. No reference oscillator creates a perfect frequency reference; in practice there is always some deviation from the true frequency.

[0266] In wireless communication systems, the differences in the outputs of the reference oscillators at the transmitter and receivers create the phenomena known as carrier frequency offset, or simply frequency offset, at the receiver. Essentially there is some residual modulation in the received signal (corresponding to the difference in the transmit and receive carriers), which occurs after downconversion. This creates distortion in the received signal resulting in higher bit error rates and lower throughput.

[0267] There are different techniques for dealing with carrier frequency offset. Most approaches estimate the carrier frequency offset at the receiver then apply a carrier frequency offset correction algorithm. The carrier frequency offset estimation algorithm may be blind using offset QAM (T. Fusco and M. Tanda, "Blind Frequency-offset Estimation for OFDM/OQAM Systems," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 55, pp. 1828–1838, 2007); periodic properties (E. Serpedin, A. Chevreuil, G. B. Giannakis, and P. Loubaton, "Blind channel and carrier frequency offset estimation using periodic modulation precoders," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 48, no. 8, pp. 2389–2405, Aug. 2000); or the cyclic prefix in orthogonal frequency division multiplexing (OFDM) structure approaches (J. J. van de Beek, M. Sandell, and P. O. Borjesson, "ML estimation of time and frequency offset in OFDM systems," Signal Processing, IEEE Transactions on [see also

Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 45, no. 7, pp. 1800–1805, July 1997; U. Tureli, H. Liu, and M. D. Zoltowski, "OFDM blind carrier offset estimation: ESPRIT," IEEE Trans. Commun., vol. 48, no. 9, pp. 1459–1461, Sept. 2000; M. Luise, M. Marselli, and R. Reggiannini, "Low-complexity blind carrier frequency recovery for OFDM signals over frequency-selective radio channels," IEEE Trans. Commun., vol. 50, no. 7, pp. 1182–1188, July 2002).

[0268] Alternatively special training signals may be utilized including a repeated data symbol (P. H. Moose, "A technique for orthogonal frequency division multiplexing frequency offset correction," IEEE Trans. Commun., vol. 42, no. 10, pp. 2908–2914, Oct. 1994); two different symbols (T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997); or periodically inserted known symbol sequences (M. Luise and R. Reggiannini, "Carrier frequency acquisition and tracking for OFDM systems," IEEE Trans. Commun., vol. 44, no. 11, pp. 1590–1598, Nov. 1996). The correction may occur in analog or in digital. The receiver can also use carrier frequency offset estimation to precorrect the transmitted signal to eliminate offset. Carrier frequency offset correction has been studied extensively for multicarrier and OFDM systems due to their sensitivity to frequency offset (J. J. van de Beek, M. Sandell, and P. O. Borjesson, "ML estimation of time and frequency offset in OFDM systems," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 45, no. 7, pp. 1800–1805, July 1997; U. Tureli, H. Liu, and M. D. Zoltowski, "OFDM blind carrier offset estimation: ESPRIT," IEEE Trans. Commun., vol. 48, no. 9, pp. 1459–1461, Sept. 2000; T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997; M. Luise, M. Marselli, and R. Reggiannini, "Low-complexity blind carrier frequency recovery for

OFDM signals over frequency-selective radio channels," IEEE Trans. Commun., vol. 50, no. 7, pp. 1182–1188, July 2002).

[0269] Frequency offset estimation and correction is an important issue for multi-antenna communication systems, or more generally MIMO (multiple input multiple output) systems. In MIMO systems where the transmit antennas are locked to one frequency reference and the receivers are locked to another frequency reference, there is a single offset between the transmitter and receiver. Several algorithms have been proposed to tackle this problem using training signals (K. Lee and J. Chun, "Frequency-offset estimation for MIMO and OFDM systems using orthogonal training sequences," IEEE Trans. Veh. Technol., vol. 56, no. 1, pp. 146–156, Jan. 2007; M. Ghogho and A. Swami, "Training design for multipath channel and frequency offset estimation in MIMO systems," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 54, no. 10, pp. 3957–3965, Oct. 2006, and adaptive tracking C. Oberli and B. Daneshrad, "Maximum likelihood tracking algorithms for MIMOOFDM," in Communications, 2004 IEEE International Conference on, vol. 4, June 20– 24, 2004, pp. 2468–2472). A more severe problem is encountered in MIMO systems where the transmit antennas are not locked to the same frequency reference but the receive antennas are locked together. This happens practically in the uplink of a spatial division multiple access (SDMA) system, which can be viewed as a MIMO system where the different users correspond to different transmit antennas. In this case the compensation of frequency offset is much more complicated. Specifically, the frequency offset creates interference between the different transmitted MIMO streams. It can be corrected using complex joint estimation and equalization algorithms (A. Kannan, T. P. Krauss, and M. D. Zoltowski, "Separation of cochannel signals under imperfect timing and carrier synchronization," IEEE Trans. Veh. Technol., vol. 50, no. 1, pp. 79–96, Jan. 2001), and equalization followed by frequency offset estimation (T. Tang and R. W. Heath, "Joint

frequency offset estimation and interference cancellation for MIMO-OFDM systems [mobile radio]," 2004. VTC2004-Fall. 2004 IEEE 60th Vehicular Technology Conference, vol. 3, pp. 1553–1557, Sept.26–29, 2004; X. Dai, "Carrier frequency offset estimation for OFDM/SDMA systems using consecutive pilots," IEEE Proceedings- Communications, vol. 152, pp. 624–632, Oct.7, 2005). Some work has dealt with the related problem of residual phase off-set and tracking error, where residual phase offsets are estimated and compensated after frequency offset estimation, but this work only consider the uplink of an SDMA OFDMA system (L. Haring, S. Bieder, and A. Czylwik, "Residual carrier and sampling frequency synchronization in multiuser OFDM systems," 2006. VTC 2006-Spring. IEEE 63rd Vehicular Technology Conference, vol. 4, pp. 1937–1941, 2006). The most severe case in MIMO systems occurs when all transmit and receive antennas have different frequency references. The only available work on this topic only deals with asymptotic analysis of estimation error in flat fading channels (O. Besson and P. Stoica, "On parameter estimation of MIMO flat-fading channels with frequency offsets," Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol. 51, no. 3, pp. 602–613, Mar. 2003).

[0270] A case that has not been significantly investigated occurs when the different transmit antennas of a MIMO system do not have the same frequency reference and the receive antennas process the signals independently. This happens in what is known as a distributed input distributed-output (DIDO) communication system, also called the MIMO broadcast channel in the literature. DIDO systems consist of one access point with distributed antennas that transmit parallel data streams (via precoding) to multiple users to enhance downlink throughput, while exploiting the same wireless resources (i.e., same slot duration and frequency band) as conventional SISO systems. Detailed description of DIDO systems was

presented in, S. G. Perlman and T. Cotter, "System and method for distributed input-distributed output wireless communications," United States Patent Application 20060023803, July 2004. There are many ways to implement DIDO precoders. One solution is block diagonalization (BD) described in, for example, Q. H. Spencer, A. L. Swindlehurst, and M. Haardt, "Zero-forcing methods for downlink spatial multiplexing in multiuser MIMO channels," IEEE Trans. Sig. Proc., vol. 52, pp. 461–471, Feb. 2004; K. K. Wong, R. D. Murch, and K. B. Letaief, "A joint-channel diagonalization for multiuser MIMO antenna systems," IEEE Trans. Wireless Comm., vol. 2, pp. 773–786, Jul 2003; L. U. Choi and R. D. Murch, "A transmit preprocessing technique for multiuser MIMO systems using a decomposition approach," IEEE Trans. Wireless Comm., vol. 3, pp. 20–24, Jan 2004; Z. Shen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization," accepted for publication in IEEE Trans. Sig. Proc., Sep. 2005; Z. Shen, R. Chen, J. G. Andrews, R. W. Heath, and B. L. Evans, "Sum capacity of multiuser MIMO broadcast channels with block diagonalization," submitted to IEEE Trans. Wireless Comm., Oct. 2005; R. Chen, R. W. Heath, and J. G. Andrews, "Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers," accepted to IEEE Trans. on Signal Processing, 2005.

[0271] In DIDO systems, transmit precoding is used to separate data streams intended for different users. Carrier frequency offset causes several problems related to the system implementation when the transmit antenna radio frequency chains do not share the same frequency reference. When this happens, each antenna is effectively transmits at a slightly different carrier frequency. This destroys the integrity of the DIDO precoder resulting in each user experiencing extra interference. Propose below are several solutions to this problem. In one embodiment of the solution, the DIDO transmit antennas share a frequency reference through a wired, optical, or

wireless network. In another embodiment of the solution, one or more users estimate the frequency offset differences (the relative differences in the offsets between pairs of antennas) and send this information back to the transmitter. The transmitter then precorrects for the frequency offset and proceeds with the training and precoder estimation phase for DIDO. There is a problem with this embodiment when there are delays in the feedback channel. The reason is that there may be residual phase errors created by the correction process that are not accounted for in the subsequent channel estimation. To solve this problem, one additional embodiment uses a novel frequency offset and phase estimator that can correct this problem by estimating the delay. Results are presented based both on simulations and practical measurements carried out with a DIDO-OFDM prototype.

[0272] The frequency and phase offset compensation method proposed in this document may be sensitive to estimation errors due to noise at the receiver. Hence, one additional embodiment proposes methods for time and frequency offset estimation that are robust also under low SNR conditions.

[0273] There are different approaches for performing time and frequency offset estimation. Because of its sensitivity to synchronization errors, many of these approaches were proposed specifically for the OFDM waverform.

[0274] The algorithms typically do not exploit the structure of the OFDM waveform thus they are generic enough for both single carrier and multicarrier waveforms. The algorithm described below is among a class of techniques that employ known reference symbols, e.g. training data, to aid in synchronization. Most of these methods are extensions of Moose's frequency offset estimator (see P. H. Moose, "A technique for orthogonal frequency division multiplexing frequency offset correction," IEEE Trans. Commun., vol. 42, no. 10, pp. 2908–2914, Oct. 1994.). Moose proposed to use two repeated training signals and derived the frequency offset using the phase difference between both received signals. Moose's method can only correct for the fractional frequency offset. An extension of the Moose method was proposed

by Schmidl and Cox (T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997.). Their key innovation was to use one periodic OFDM symbol along with an additional differentially encoded training symbol. The differential encoding in the second symbol enables integer offset correction. Coulson considered a similar setup as described in T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997, and provided a detailed discussion of algorithms and analysis as described in A. J. Coulson, "Maximum likelihood synchronization for OFDM using a pilot symbol: analysis," IEEE J. Select. Areas Commun., vol. 19, no. 12, pp. 2495–2503, Dec. 2001.; A. J. Coulson, "Maximum likelihood synchronization for OFDM using a pilot symbol: algorithms,"IEEE J. Select. Areas Commun., vol. 19, no. 12, pp. 2486–2494, Dec. 2001. One main difference is that Coulson uses repeated maximum length sequences to provide good correlation properties. He also suggests using chirp signals because of their constant envelope properties in the time and frequency domains. Coulson considers several practical details but does not include integer estimation. Multiple repeated training signals were considered by Minn et. al. in H. Minn, V. K. Bhargava, and K. B. Letaief, "A robust timing and frequency synchronization for OFDM systems," IEEE Trans. Wireless Commun., vol. 2, no. 4, pp. 822–839, July 2003, but the structure of the training was not optimized. Shi and Serpedin show that the training structure has some optimality form the perspective of frame synchronization (K. Shi and E. Serpedin, "Coarse frame and carrier synchronization of OFDM systems: a new metric and comparison," IEEE Trans. Wireless Commun., vol. 3, no. 4, pp. 1271–1284, July 2004). One embodiment of the invention uses the Shi and Serpedin approach to perform frame synchronization and fractional frequency offset estimation.

[0275] Many approaches in the literature focus on frame synchronization and fractional frequency offset correction. Integer offset correction is solved

using an additional training symbol as in T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997. For example, Morrelli et. al. derived an improved version of T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," IEEE Trans. Commun., vol. 45, no. 12, pp. 1613–1621, Dec. 1997, in M. Morelli, A. N. D'Andrea, and U. Mengali, "Frequency ambiguity resolution in OFDM systems," IEEE Commun. Lett., vol. 4, no. 4, pp. 134–136, Apr. 2000. An alternative approach using a different preamble structure was suggested by Morelli and Mengali (M. Morelli and U. Mengali, "An improved frequency offset estimator for OFDM applications," IEEE Commun. Lett., vol. 3, no. 3, pp. 75–77, Mar. 1999). This approach uses the correlations between M repeated identical training symbols to increase the range of the fractional frequency offset estimator by a factor of M. This is the best linear unbiased estimator and accepts a large offset (with proper design) but does not provide good timing synchronization.

System Description

[0276] One embodiment of the invention uses pre-coding based on channel state information to cancel frequency and phase offsets in DIDO systems. See Figure 11 and the associated description above for a description of this embodiment.

[0277] In one embodiment of the invention, each user employs a receiver equipped with frequency offset estimator/compensator. As illustrated in **Figure 45**, in one embodiment of the invention, a system including the receiver includes a plurality of RF units 4508, a corresponding plurality of A/D units 4510, a receiver equipped with a frequency offset estimator/compensator 4512 and a DIDO feedback generator unit 4506.

[0278] The RF units 4508 receive signals transmitted from the DIDO transmitter units, downconvert the signals to baseband and provide the downconverted signals to the A/D units 4510. The A/D units 4510 then convert the signal from analog to digital and send it to the frequency offset

estimator/compensator units 4512. The frequency offset estimator/compensator units 4512 estimate the frequency offset and compensate for it, as described herein, and then send the compensated signal to the OFDM units 4513. The OFDM units 4513 remove the cyclic prefix and operate the Fast Fourier Transform (FFT) to report the signal to the frequency domain. During the training period the OFDM units 4513 send the output to the channel estimate unit 4504 that computes the channel estimates in the frequency domain. Alternatively, the channel estimates can be computed in the time domain. During the data period the OFDM units 4513 send the output to the DIDO receiver unit 4502 which demodulates/decodes the signal to obtain the data. The channel estimate unit 4504 sends the channel estimates to the DIDO feedback generator unit 4506 that may quantize the channel estimates and send them back to the transmitter via the feedback control channel, as illustrated.

Description of One Embodiment of an Algorithm for a DIDO 2 × 2 Scenario

[0279] Described below are embodiments of an algorithm for frequency/phase offset compensation in DIDO systems. The DIDO system model is initially described with and without frequency/phase offsets. For the sake of the simplicity, the particular implementation of a DIDO 2 × 2 system is provided. However, the underlying principles of the invention may also be implemented on higher order DIDO systems.

DIDO System Model w/o Frequency and Phase Offset

[0280] The received signals of DIDO 2 × 2 can be written for the first user as

$$r_1[t] = h_{11}(w_{11}x_1[t] + w_{21}x_2[t]) + h_{12}(w_{12}x_1[t] + w_{22}x_2[t]) \quad (1)$$

and for the second user as

$$r_2[t] = h_{21}(w_{11}x_1[t] + w_{21}x_2[t]) + h_{22}(w_{12}x_1[t] + w_{22}x_2[t]) \quad (2)$$

where t is the discrete time index, h_{mn} hand w_{mn} are the channel and the DIDO precoding weights between the m-th user and n-th transmit antenna, respectively, and x_m is the transmit signal to user m. Note that h_{mn} and w_{mn} are not a function of t since we assume the channel is constant over the period between training and data transmission.

[0281] In the presence of frequency and phase offset, the received signals are expressed as

$$r_1[t] = e^{j(\omega_{T1}-\omega_{T1})T_s(t-t_{11})}h_{11}(w_{11}x_1[t] + w_{21}x_2[t]) + e^{j(\omega_{T1}-\omega_{T2})T_s(t-t_{12})}h_{12}(w_{12}x_1[t] + w_{22}x_2[t]) \quad (3)$$

and

$$r_2[t] = e^{j(\omega_{T1}-\omega_{T1})T_s(t-t_{21})}h_{21}(w_{11}x_1[t] + w_{21}x_2[t]) + e^{j(\omega_{T2}-\omega_{T2})T_s(t-t_{22})}h_{22}(w_{12}x_1[t] + w_{22}x_2[t]) \quad (4)$$

where T_s is the symbol period, $\omega_{Tn} = 2\Pi f_{Tn}$ for the n-th transmit antenna, $\omega_{Um} = 2\Pi f_{Um}$ for the m-th user, and f_{Tn} and f_{Um} are the actual carrier frequencies (affected by offset) for the n-th transmit antenna and m-th user, respectively. The values t_{mn} denote random delays that cause phase offset over the channel h_{mn}. **Figure 46** depicts the DIDO 2 × 2 system model.

[0282] For the time being, we use the following definitions:

$$\Delta\omega_{mn} = \omega_{Um} - \omega_{Tn} \quad (5)$$

to denote the frequency offset between the m-th user and the n-th transmit antenna.

Description of One Embodiment of the Invention

[0283] A method according to one embodiment of the invention is illustrated in **Figure 47**. The method includes the following general steps (which include sub-steps, as illustrated): training period for frequency offset estimation 4701; training period for channel estimation 4702; data transmission via DIDO precoding with compensation 4703. These steps are described in detail below.

(a) Training Period for Frequency Offset Estimation (4701)

[0284] During the first training period the base station sends one or more training sequences from each transmit antennas to one of the users (4701a). As described herein "users" are wireless client devices. For the DIDO 2 × 2 case, the signal received by the *m*-th user is given by

$$r_m[t] = e^{j\Delta\omega_{m1}T_s(t-t_{m1})}h_{m1}p_1[t] + e^{j\Delta\omega_{m2}T_s(t-t_{m2})}h_{m2}p_2[t] \quad (6)$$

where p_1 and p_2 are the training sequences transmitted from the first and second antennas, respectively.

[0285] The *m*-th user may employ any type of frequency offset estimator (i.e., convolution by the training sequences) and estimates the offsets $\Delta\omega_{m1}$ and $\Delta\omega_{m2}$. Then, from these values the user computes the frequency offset between the two transmit antennas as

$$\Delta\omega_T = \Delta\omega_{m2} - \Delta\omega_{m1} = \omega_{T1} - \omega_{T2} \quad (7)$$

Finally, the value in (7) is fed back to the base station (4701b).

[0286] Note that p_1 and p_2 in (6) are designed to be orthogonal, so that the user can estimate $\Delta\omega_{m1}$ and $\Delta\omega_{m2}$. Alternatively, in one embodiment, the same training sequence is used over two consecutive time slots and the user

estimates the offset from there. Moreover, to improve the estimate of the offset in (7) the same computations described above can be done for all users of the DIDO systems (not just for the *m*-th user) and the final estimate may be the (weighted) average of the values obtained from all users. This solution, however, requires more computational time and amount of feedback. Finally, updates of the frequency offset estimation are needed only if the frequency offset varies over time. Hence, depending on the stability of the clocks at the transmitter, this step 4701 of the algorithm can be carried out on a long-term basis (i.e., not for every data transmission), resulting in reduction of feedback overhead.

(b) Training Period for Channel Estimation (4702)

[0287] During the second training period, the base station first obtains the frequency offset feedback with the value in (7) from the *m*-th user or from the plurality of users. The value in (7) is used to pre-compensate for the frequency offset at the transmit side. Then, the base station sends training data to all the users for channel estimation (4702a).

[0288] For DIDO 2 × 2 systems, the signal received at the first user is given by

$$r_1[t] = e^{j\Delta\omega_{11}T_s(t-\tilde{t}_{11})}h_{11}p_1[t] + e^{j\Delta\omega_{12}T_s(t-\tilde{t}_{12})}h_{12}e^{-j\Delta\omega_T T_s t}p_2[t] \qquad (8)$$

and at the second user by

$$r_2[t] = e^{j\Delta\omega_{21}T_s(t-\tilde{t}_{21})}h_{21}p_1[t] + e^{j\Delta\omega_{22}T_s(t-\tilde{t}_{22})}h_{22}e^{-j\Delta\omega_T T_s t}p_2[t] \qquad (9)$$

where $\tilde{t}_{mn} = t_{mn} + \Delta t$ and Δt is random or known delay between the first and second transmissions of the base station. Moreover, p_1 and p_2 are the training sequences transmitted from the first and second antennas, respectively, for frequency offset and channel estimation.

[0289] Note that the pre-compensation is applied only to the second antennas in this embodiment.

Expanding (8) we obtain

$$r_1[t] = e^{j\Delta\omega_{11}T_s t} e^{j\theta_{11}} [h_{11} p_1[t] + e^{j(\theta_{12}-\theta_{11})} h_{12} p_2[t]] \quad (10)$$

and similarly for the second user

$$r_2[t] = e^{j\Delta\omega_{21}T_s t} e^{j\theta_{21}} [h_{21} p_1[t] + e^{j(\theta_{22}-\theta_{21})} h_{22} p_2[t]] \quad (11)$$

where $\theta_{mn} = -\Delta\omega_{mn} T_s \tilde{\tau}_{mn}$.

[0290] At the receive side, the users compensate for the residual frequency offset by using the training sequences p_1 and p_2. Then the users estimate via training the vector channels (4702b)

$$h_1 = \begin{bmatrix} h_{11} \\ e^{j(\theta_{12}-\theta_{11})} h_{12} \end{bmatrix} \qquad h_2 = \begin{bmatrix} h_{21} \\ e^{j(\theta_{22}-\theta_{21})} h_{22} \end{bmatrix} \quad (12)$$

[0291] These channel in (12) or channel state information (CSI) is fed back to the base station (4702b) that computes the DIDO precoder as described in the following subsection.

(c) DIDO Precoding with Pre-compensation (4703)

[0292] The base station receives the channel state information (CSI) in (12) from the users and computes the precoding weights via block diagonalization (BD) (4703a), such that

$$\mathbf{w}_1^T \mathbf{h}_2 = 0, \qquad \mathbf{w}_2^T \mathbf{h}_1 = 0 \quad (13)$$

where the vectors h_1 are defined in (12) and $\mathbf{w}_m = [w_{m1}, w_{m2}]$. Note that the invention presented in this disclosure can be applied to any other DIDO precoding method besides BD. The base station also pre-compensates for the frequency offset by employing the estimate in (7) and phase offset by

estimating the delay (Δt_o) between the second training transmission and the current transmission (4703a). Finally, the base station sends data to the users via the DIDO precoder (4703b).

[0293] After this transmit processing, the signal received at user 1 is given by

$$\begin{aligned} r_1[t] &= e^{j\Delta\omega_{11}T_s(t-\tilde{\tau}_{11}-\Delta t_o)}h_{11}[w_{11}x_1[t]+w_{21}x_2[t]] \\ &= e^{j\Delta\omega_{12}T_s(t-\tilde{\tau}_{12}-\Delta t_o)}h_{12}e^{-j\Delta\omega_{T}T_s(t-\Delta t_o)}[w_{12}x_1[t]+w_{22}x_2[t]] \\ &= \gamma_1[t]\left[h_{11}\left(w_{11}x_1[t]+w_{21}x_2[t]\right)+e^{j(\Delta\omega_{11}t_{11}-\Delta\omega_{12}t_{12})}h_{12}\left(w_{12}x_1[t]+w_{22}x_2[t]\right)\right] \\ &= \gamma_1[t]\left[\left(h_{11}w_{11}+e^{j(\theta_{12}-\theta_{11})}h_{12}w_{12}\right)x_1[t]+\left(h_{11}w_{21}+e^{j(\theta_{12}-\theta_{11})}h_{12}w_{22}\right)x_2[t]\right] \end{aligned}$$

(14)

where $\gamma_1[t]=e^{j\Delta\omega_{11}T_s(t-\tilde{\tau}_{11}-\Delta t_o)}$. Using the property (13) we obtain

$$r_1[t]=\gamma_1[t]\mathbf{w}_1^T\mathbf{h}_1x_1[t]. \quad (15)$$

Similarly, for user 2 we get

$$\begin{aligned} r_2[t] &= e^{j\Delta\omega_{21}T_s(t-\tilde{\tau}_{21}-\Delta t_o)}h_{21}[w_{11}x_1[t]+w_{21}x_2[t]] \\ &\quad + e^{j\Delta\omega_{22}T_s(t-\tilde{\tau}_{22}-\Delta t_o)}h_{22}e^{-j\Delta\omega_{T}T_s(t-\Delta t_o)}[w_{12}x_1[t]+w_{22}x_2[t]] \end{aligned}$$

(16)

and expanding (16)

$$r_2[t]=\gamma_2[t]\mathbf{w}_2^T\mathbf{h}_2x_2[t] \quad (17)$$

where $\gamma_2[t]=e^{j\Delta\omega_{21}T_s(t-\tilde{\tau}_{21}-\Delta t_o)}$.

[0294] Finally, the users compute the residual frequency offset and the channel estimation to demodulate the data streams $x_1[t]$ and $x_2[t]$ (4703c).

Generalization to DIDO $N \times M$

[0295] In this section, the previously described techniques are generalized to DIDO systems with N transmit antennas and M users.

i. Training Period for Frequency Offset Estimation

[0296] During the first training period, the signal received by the m-th user as a result of the training sequences sent from the N antennas is given by

$$r_m[t] = \sum_{n=1}^{N} e^{j\Delta\omega_{mn} t_s (t - t_{mn})} h_{mn} p_n[t] \qquad (18)$$

where p_n is the training sequences transmitted from the n-th antenna.

[0297] After estimating the offsets $\Delta\omega_{mn}, \forall n = 1,\ldots,N$, the m-th user computes the frequency offset between the first and the n-th transmit antenna as

$$\Delta\omega_{T,1n} = \Delta\omega_{mn} - \Delta\omega_{m1} = \omega_{T1} - \omega_{Tn}. \qquad (19)$$

Finally, the values in (19) are fed back to the base station.

ii. Training Period for Channel Estimation

[0298] During the second training period, the base station first obtains the frequency offset feedback with the value in (19) from the m-th user or from the plurality of users. The value in (19) is used to pre-compensate for the frequency offset at the transmit side. Then, the base station sends training data to all the users for channel estimation.

For DIDO $N \times M$ systems, the signal received at the m-th user is given by

$$\begin{aligned} r_m[t] &= e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}h_{m1}p_1[t] + \sum_{n=2}^{N} e^{j\Delta\omega_{mn}T_s(t-\tilde{t}_{mn})}h_{mn}e^{-j\Delta\omega_{n1}T_s t}p_n[t] \\ &= e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}\left[h_{m1}p_1[t] + \sum_{n=2}^{N} e^{j(\theta_{mn}-\theta_{m1})}h_{mn}p_n[t]\right] \\ &= e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}\sum_{n=1}^{N} e^{j(\theta_{mn}-\theta_{m1})}h_{mn}p_n[t] \end{aligned} \quad (20)$$

where $\theta_{mn} = -\Delta\omega_{mn}T_s\tilde{t}_{mn}, \tilde{t}_{mn} = t_{mn} + \Delta t$ and Δt is random or known delay between the first and second transmissions of the base station. Moreover, p_n is the training sequence transmitted from the n-th antenna for frequency offset and channel estimation.

[0299] At the receive side, the users compensate for the residual frequency offset by using the training sequences p_n. Then, each users m estimates via training the vector channel

$$\mathbf{h}_m = \begin{bmatrix} h_{m1} \\ e^{j(\theta_{m2}-\theta_{m1})}h_{m2} \\ \vdots \\ e^{j(\theta_{mN}-\theta_{m1})}h_{mN} \end{bmatrix} \quad (21)$$

and feeds back to the base station that computes the DIDO precoder as described in the following subsection.

iii. DIDO Precoding with Pre-compensation

[0300] The base station receives the channel state information (CSI) in (12) from the users and computes the precoding weights via block diagonalization (BD), such that

$$\mathbf{w}_m^T\mathbf{h}_l = 0, \qquad \forall m \neq l, \qquad m = 1,\ldots,M \quad (22)$$

where the vectors $\boldsymbol{h}_m$ are defined in (21) and $\mathbf{w}_m = [w_{m1}, w_{m2}, \ldots, w_{mN}]$. The base station also pre-compensates for the frequency offset by employing the

estimate in (19) and phase offset by estimating the delay (Δt_{o}) between the second training transmission and the current transmission. Finally, the base station sends data to the users via the DIDO precoder.

[0301] After this transmit processing, the signal received at user *i* is given by

$$\begin{aligned}
r_i[t] &= e^{j\Delta\omega_{i1}T_s(t-\tilde{t}_{i1}-\Delta t_o)}h_{i1}\sum_{m=1}^{M}w_{m1}x_m[t]+ \\
&+\sum_{n=2}^{N}e^{j\Delta\omega_{in}T_s(t-\tilde{t}_{in}-\Delta t_o)}h_{in}e^{-j\Delta\omega_{i1n}T_s(t-\Delta t_o)}\sum_{m=1}^{M}w_{mn}x_m[t] \\
&= e^{j\Delta\omega_{i1}T_s(t-\Delta t_o)}e^{-j\lambda\omega_{i1}T_s\tilde{t}_{i1}}h_{i1}\sum_{m=1}^{M}w_{m1}x_m[t] \\
&+\sum_{n=2}^{N}e^{j\Delta\omega_{i1}T_s(t-\Delta t_o)}e^{-j\lambda\omega_{in}T_s\tilde{t}_{in}}h_{in}\sum_{m=1}^{M}w_{mn}x_m[t] \\
&= \gamma_i[t]\left[h_{i1}\sum_{m=1}^{M}w_{m1}x_m[t]+\sum_{n=2}^{N}e^{j(\theta_{in}-\theta_{i1})}h_{in}\sum_{m=1}^{M}w_{m1}x_m[t]\right] \\
&= \gamma_i[t]\left[\sum_{n=1}^{N}e^{j(\theta_{in}-\theta_{i1})}h_{in}\sum_{m=1}^{M}w_{mn}x_m[t]\right] \\
&= \gamma_i[t]\sum_{m=1}^{M}\left[\sum_{n=1}^{N}e^{j(\theta_{in}-\theta_{i1})}h_{in}w_{mn}\right]x_m[t] \\
&= \gamma_i[t]\sum_{m=1}^{M}\mathbf{w}_m^T\mathbf{h}_i x_m[t]
\end{aligned} \tag{23}$$

Where $\gamma_i[n]=e^{j\lambda\omega_{i1}T_s(t-\tilde{t}_{i1}-\Delta t_o)}$. Using the property (22) we obtain

$$r_i[t]=\gamma_i[t]\mathbf{w}_i^T\mathbf{h}_i x_i[t] \tag{24}$$

[0302] Finally, the users compute the residual frequency offset and the channel estimation to demodulate the data streams $x_i[t]$.

Results

[0303] **Figure 48** shows the SER results of DIDO 2 × 2 systems with and without frequency offset. It is possible to see that the proposed method completely cancels the frequency/phase offsets yielding the same SER as systems without offsets.

[0304] Next, we evaluate the sensitivity of the proposed compensation method to frequency offset estimation errors and/or fluctuations of the offset in time. Hence, we re-write (14) as

$$\begin{aligned} r_1[t] = {} & e^{j\Delta\omega_{11}T_s(t-\tilde{t}_{11}-\Delta t_o)} h_{11}\left[w_{11}x_1[t]+w_{21}x_2[t]\right] \\ & + e^{j\Delta\omega_{12}T_s(t-\tilde{t}_{12}-\Delta t_o)} h_{12} e^{-j(\Delta\omega_1+2\Pi\in)T_s(t-\Delta t_o)}\left[w_{12}x_1[t]+w_{22}x_2[t]\right] \end{aligned} \quad (25)$$

where $\in$ indicates the estimation error and/or variation of the frequency offset between training and data transmission. Note that the effect of $\in$ is to destroy the orthogonality property in (13) such that the interference terms in (14) and (16) are not completely pre-canceled at the transmitter. As a results of that, the SER performance degrades for increasing values of $\in$.

[0305] **Figure 48** shows the SER performance of the frequency offset compensation method for different values of $\in$. These results assume T_s = 0.3 ms (i.e., signal with 3 KHz bandwidth). We observe that for $\in$ = 0.001 Hz (or less) the SER performance is similar to the no offset case.

f. Description of One Embodiment of an Algorithm for Time and Frequency Offset Estimation

[0306] Hereafter, we describe additional embodiments to carry out time and frequency offset estimation (4701b in Figure 47). The transmit signal structure under consideration is illustrated in H. Minn, V. K. Bhargava, and K. B. Letaief, "A robust timing and frequency synchronization for OFDM systems," IEEE Trans. Wireless Commun., vol. 2, no. 4, pp. 822–839, July 2003, and studied in more detail in K. Shi and E. Serpedin, "Coarse frame and

carrier synchronization of OFDM systems: a new metric and comparison," IEEE Trans. Wireless Commun., vol. 3, no. 4, pp. 1271–1284, July 2004. Generally sequences with good correlation properties are used for training. For example, for our system, Chu sequences are used which are derived as described in D. Chu, "Polyphase codes with good periodic correlation properties (corresp.)," IEEE Trans. Inform. Theory. vol. 18, no. 4, pp. 531–532, July 1972. These sequences have an interesting property that they have perfect circular correlations. Let L_{cp} denote the length of the cyclic prefix and let N_t denote the length of the component training sequences. Let $N_t = M_t$, where M_t is the length of the training sequence. Under these assumptions the transmitted symbol sequence for the preamble can be written as

$$\begin{aligned} s[n] &= t[n - N_t] && \text{for } n = -1, \ldots, -L_{cp} \\ s[n] &= t[n] && \text{for } n = 0, \ldots, N_t - 1 \\ s[n] &= t[n - N_t] && \text{for } n = N_t, \ldots, 2N_t - 1 \\ s[n] &= -t[n - 2N_t] && \text{for } n = 2N_t, \ldots, 3N_t - 1 \\ s[n] &= t[n - 3N_t] && \text{for } n = 3N_t, \ldots, 4N_t - 1. \end{aligned}$$

Note that the structure of this training signal can be extended to other lengths but
repeating the block structure. For example, to use 16 training signals we consider a structure such as:

$$[CP, B, B, -B, B, B, B, -B, B, -B, -B, B, -B, B, B, -B, B,].$$

By using this structure and letting $N_t = 4\ M_t$ all the algorithms to be described can be employed without modification. Effectively we are repeating the training sequence. This is especially useful in cases where a suitable training signal may not be available.

[0307] Consider the following received signal, after matched filtering and downsampling to the symbol rate:

$$r[n] = e^{2\pi\varepsilon\iota}\sum_{l=0}^{L} h[l]s[n-l-\Delta] + v[n]$$

where ε is the unknown discrete-time frequency offset, Δ is the unknown frame offset, h[l] are the unknown discrete-time channel coefficients, and v[n] is additive noise. To explain the key ideas in the following sections the presence of additive noise is ignored.

i. Coarse Frame Synchronization

[0308] The purpose of coarse frame synchronization is to solve for the unknown frame offset Δ. Let us make the following definitions

$$\begin{aligned}
\mathbf{r}_1[n] &:= [r[n], r[n+1], \ldots, r[n+N_c-1]]^T, \\
\bar{\mathbf{r}}_1[n] &:= [r[n+L_{cp}], r[n+1], \ldots, r[n+N_t-1]]^T, \\
\mathbf{r}_2[n] &:= [r[n+N_t], r[n+1+N_t], \ldots, r[n+2N_t-1]]^T, \\
\bar{\mathbf{r}}_2[n] &:= [r[n+L_{cp}+N_t], r[n+1+L_{cp}+N_t], \ldots, r[n+L_{cp}+2N_t-1]]^T, \\
\mathbf{r}_3[n] &:= [r[n+2N_t], r[n+1+2N_t], \ldots, r[n+3N_t-1]]^T, \\
\bar{\mathbf{r}}_3[n] &:= [r[n+L_{cp}+2N_t], r[n+L_{cp}+1+2N_t], \ldots, r[n+L_{cp}+3N_t-1]]^T, \\
\mathbf{r}_4[n] &:= [r[n+3N_t], r[n+1+3N_t], \ldots, r[n+4N_t-1]]^T, \\
\bar{\mathbf{r}}_4[n] &:= [r[n+L_{cp}+3N_t], r[n+L_{cp}+1+3N_t], \ldots, r[n+L_{cp}+4N_t-1]]^T.
\end{aligned}$$

The proposed coarse frame synchronization algorithm is inspired from the algorithm in K. Shi and E. Serpedin, "Coarse frame and carrier synchronization of OFDM systems: a new metric and comparison," IEEE Trans. Wireless Commun., vol. 3, no. 4, pp. 1271–1284, July 2004, derived from a maximum likelihood criterion.

Method 1 – Improved coarse frame synchronization: *the coarse frame synchronization estimator solves the following optimization*

$$\hat{\Delta} = \arg\max_{k\in\mathcal{Z}} \frac{|P_1(k)| + |P_2(k)| + |P_3(k)|}{\|\mathbf{r}_1\|^2 + \|\mathbf{r}_2\|^2 + \|\mathbf{r}_3\|^2 + \|\mathbf{r}_4\|^2 + \frac{1}{2}(\|\bar{\mathbf{r}}_1\|^2 + \|\bar{\mathbf{r}}_2\|^2 + \|\bar{\mathbf{r}}_3\|^2 + \|\bar{\mathbf{r}}_4\|^2)}$$

where

$$\begin{aligned} P_1[k] &= \mathbf{r}_1^*[k]\mathbf{r}_2[k] - \mathbf{r}_3^*[k]\mathbf{r}_4[k] - \bar{\mathbf{r}}_2^*[k]\bar{\mathbf{r}}_3[k] \\ P_2[k] &= \mathbf{r}_2^*[k]\mathbf{r}_1[k] - \mathbf{r}_4^*[k]\mathbf{r}_3[k] \\ P_3[k] &= \bar{\mathbf{r}}_1^*[k]\bar{\mathbf{r}}_4[k]. \end{aligned}$$

Let the corrected signal be defined as

$$r_c[n] = r[n - \hat{\Delta} - \lceil L_{cp}/4 \rceil].$$

The additional correction term is used to compensate for small initial taps in the channel and can be adjusted based on the application. This extra delay will be included henceforth in the channel.

ii. Fractional Frequency Offset Correction

[0309] The fractional frequency offset correction follows the coarse frame synchronization block.

Method 2 – Improved fractional frequency offset correction: *the fractional frequency offset is the solution to*

$$\hat{\epsilon}_f = \frac{\text{phase}P_1[\hat{\Delta}]}{2\pi N_t}.$$

This is known as a fractional frequency offset because the algorithm can only correct for offsets

$$|\epsilon_f| < \frac{1}{2N_t}.$$

This problem will be solved in the next section. Let the fine frequency offset corrected signal be defined as

$$r_f[n] = e^{-j2\pi\hat{\epsilon}}r_c[n].$$

[0310] Note that the Methods 1 and 2 are an improvement to K. Shi and E. Serpedin, "Coarse frame and carrier synchronization of OFDM systems: a new metric and comparison," IEEE Trans. Wireless Commun., vol. 3, no. 4, pp.1271–1284, July 2004 that works better in frequency-selective channels. One specific innovation here is the use of both r and $\overline{r}$ as described above. The use of $\overline{r}$ improves the prior estimator because it ignores the samples that would be contaminated due to inter-symbol interference.

iii. Integer Frequency Offset Correction

[0311] To correct for the integer frequency offset, it is necessary to write an equivalent system model for the received signal after fine frequency offset correction. Absorbing remaining timing errors into the channel, the received signal in the absence of noise has the following structure:

$$r_f[n] = e^{j2\pi\frac{nk}{N_s}}\sum_{l=0}^{L_{cp}} g[l]s[n-l]$$

for n = 0, 1, . . . , $4N_t - 1$. The integer frequency offset is *k* while the unknown equivalent channel is g[l].

Method 3 - Improved Integer frequency offset correction: *the integer frequency offset is the solution to*

$$\hat{k} = \arg\max_{v=0,1,\ldots,N_s-1} \mathbf{r}^*\mathbf{D}[k]\mathbf{S}(\mathbf{S}^*\mathbf{S})^{-1}\mathbf{S}^*\mathbf{D}[k]^*\mathbf{r}$$

where

$$\mathbf{r} = \mathbf{D}[k]\mathbf{S}\mathbf{g}$$

$$\mathbf{D}[k] := \mathrm{diag}\left\{1, e^{j2\pi\frac{k}{N_t}}, \ldots, e^{j2\pi\frac{k(4N_t-1)}{N_t}}\right\}$$

$$\mathbf{S} := \begin{bmatrix} s[0] & s[-1] & \cdots & \cdots & s[-L_{cp}] \\ s[1] & s[0] & s[-1] & \cdots & s[-L_{cp}+1] \\ s[4N_t-1] & s[4N_t-2] & s[4N_t-3] & \cdots & s[4N_t-1-L_{cp}] \end{bmatrix}$$

$$\mathbf{g} := \begin{bmatrix} g[0] \\ g[1] \\ \vdots \\ g[L_{cp}] \end{bmatrix}$$

This gives the estimate of the total frequency offset as

$$\hat{\epsilon} = \frac{\hat{k}}{N_t} + \hat{\epsilon}_f.$$

Practically, Method 3 has rather high complexity. To reduce complexity the following observations can be made. First of all, the product S $\mathbf{S}(\mathbf{S}^*\mathbf{S})^{-1}\mathbf{S}^*$ can be precomputed. Unfortunately, this still leaves a rather large matrix multiplication. An alternative is to exploit the observation that with the proposed training sequences, $\mathbf{S}^*\mathbf{S} \approx \mathbf{I}$. This leads to the following reduced complexity method.

Method 4 – Low-complexity improved integer frequency offset correction: *a low complexity integer frequency offset estimator solves*

$$\hat{k} = \arg \max_{m=0,1,\ldots,N_t-1} (\mathbf{S}^* \mathbf{D}[k]^m \mathbf{r})^* (\mathbf{S}^* \mathbf{D}[k]^m \mathbf{r}).$$

iv. Results

In this section we compare the performance of the different proposed estimators.

[0312] First, in **Figure 50** we compare the amount of overhead required for each method. Note that both of the new methods reduce the overhead required by 10x to 20x. To compare the performance of the different estimators, Monte Carlo experiments were performed. The setup considered is our usual NVIS transmit waveform constructed from a linear modulation with a symbol rate of 3K symbols per second, corresponding to a passband bandwidth of 3kHz, and raised cosine pulse shaping. For each Monte Carlo realization, the frequency offset is generated from a uniform distribution on $[-f_{max}, f_{max}]$.

[0313] A simulation with a small frequency offset of f_{max} = 2Hz and no integer offset correction is illustrated in **Figure 51**. It can be seen from this performance comparison that performance with $N_t / M_t = 1$ is slightly degraded from the original estimator, though still substantially reduces overhead. Performance with $N_t / M_t = 4$ is much better, almost 10dB. All the curves experience a knee at low SNR points due to errors in the integer offset estimation. A small error in the integer offset can create a large frequency error and thus a large mean squared error. Integer offset correction can be turned off in small offsets to improve performance.

[0314] In the presence of multipath channels, the performance of frequency offset estimators generally degrades. Turning off the integer offset estimator, however, reveals quite good performance in **Figure 52**. Thus, in multipath channels it is even more important to perform a robust coarse

correction followed by an improved fine correction algorithm. Notice that the offset performance with $N_t/M_t = 4$ is much better in the multipath case.

[0315] Embodiments of the invention may include various steps as set forth above. The steps may be embodied in machine-executable instructions which cause a general-purpose or special-purpose processor to perform certain steps. For example, the various components within the Base Stations/APs and Client Devices described above may be implemented as software executed on a general purpose or special purpose processor. To avoid obscuring the pertinent aspects of the invention, various well known personal computer components such as computer memory, hard drive, input devices, etc., have been left out of the figures.

[0316] Alternatively, in one embodiment, the various functional modules illustrated herein and the associated steps may be performed by specific hardware components that contain hardwired logic for performing the steps, such as an application-specific integrated circuit ("ASIC") or by any combination of programmed computer components and custom hardware components.

[0317] In one embodiment, certain modules such as the Coding, Modulation and Signal Processing Logic 903 described above may be implemented on a programmable digital signal processor ("DSP") (or group of DSPs) such as a DSP using a Texas Instruments' TMS320x architecture (e.g., a TMS320C6000, TMS320C5000, . . . etc). The DSP in this embodiment may be embedded within an add-on card to a personal computer such as, for example, a PCI card. Of course, a variety of different DSP architectures may be used while still complying with the underlying principles of the invention.

[0318] Elements of the present invention may also be provided as a machine-readable medium for storing the machine-executable instructions. The machine-readable medium may include, but is not limited to, flash memory, optical disks, CD-ROMs, DVD ROMs, RAMs, EPROMs, EEPROMs, magnetic or optical cards, propagation media or other type of machine-readable media suitable for storing electronic instructions. For example, the present invention may be downloaded as a computer program which may be transferred from a remote computer (e.g., a server) to a requesting computer (e.g., a client) by way of data signals embodied in a carrier wave or other propagation medium via a communication link (e.g., a modem or network connection).

[0319] Throughout the foregoing description, for the purposes of explanation, numerous specific details were set forth in order to provide a thorough understanding of the present system and method. It will be apparent, however, to one skilled in the art that the system and method may be practiced without some of these specific details. Accordingly, the scope of the present invention should be judged in terms of the claims which follow.

[0320] Moreover, throughout the foregoing description, numerous publications were cited to provide a more thorough understanding of the present invention.

Claims

1. A system for compensating for frequency and phase offsets for MU-MAS communication comprising:

one or more coding modulation units to encode and modulate information bits for each of a plurality of wireless client devices to produce encoded and modulated information bits;

one or more mapping units to map the encoded and modulated information bits to complex symbols; and

a MU-MAS frequency/phase offset aware precoding unit to exploit channel state information to compute MU-MAS frequency/phase offset aware precoding weights, the MU-MAS frequency/phase offset aware precoding unit precoding the complex symbols obtained from the mapping units using the MU-MAS frequency/phase offset aware precoding weights to pre-cancel frequency/phase offset and/or inter-user interference.

2. The system as in claim 1 further comprising: one or more orthogonal frequency division multiplexing (OFDM) units to receive the precoded signals from the MU-MAS frequency/phase offset aware precoding unit and to modulate the precoded signals according to an OFDM standard.

3. The system as in claim 2 wherein the OFDM standard comprises computing an inverse fast Fourier transform (IFFT) and adding a cyclic prefix.

4. The system as in claim 2 further comprising one or more D/A units to perform digital to analog (D/A) conversion on the output of the OFDM units to generate an analog baseband signal; and one or more radio frequency (RF) units to upconvert the baseband signal to radio frequency and transmit the signals using a corresponding one or more transmit antennas.

5. The system as in claim 1 wherein the MU-MAS frequency/phase offset aware precoding unit is implemented as a minimum mean square error (MMSE), a weighted MMSE, a zero-forcing (ZF) or a block diagonalization (BD) precoder.

6. The system as in claim 1 wherein the MU-MAS communication comprises DIDO communication and wherein the system comprises a DIDO frequency/phase offset aware precoding unit to exploit channel state information to compute DIDO frequency/phase offset aware precoding weights, the DIDO frequency/phase offset aware precoding unit precoding the complex symbols obtained from the mapping units using the weights to pre-cancel frequency/phase offset and/or inter-user interference.

7. A wireless client device for use in a system for compensating for frequency and phase offsets for MU-MAS communication comprising:

one or more RF units to receive signals transmitted from one or more MU-MAS transmitter units, and downconvert the signals to baseband;

one or more analog to digital (A/D) conversion units to receive the downconverted signals and to convert the signals from analog signals to digital signals;

a frequency/phase offset estimation/compensation unit that estimates the frequency and/or phase offset and feedbacks this information to the transmitter for pre-compensation;

one or more OFDM units to remove cyclic prefix and perform a fast Fourier transform (FFT) on the digital signals to report the signals in the frequency domain;

a channel estimate unit which receives a signal output from the one or more OFDM units during a training period and responsively calculates channel estimate data; and a feedback generator unit to transmit the channel estimate data or a training signal to a base station for use in precoding signals prior to transmission to the wireless client device.

8. The system as in claim 7 where the channel estimates are computed in the time domain by using the input to the OFDM units.

9. The wireless client device as in claim 7 wherein the feedback generator unit further comprises logic for quantizing the frequency offset and channel estimates prior to transmission to the base station.

10. The wireless client device as in claim 7 further comprising: a receiver unit which receives outputs from the OFDM units and responsively computes the receiver and demodulates/decodes the signal to obtain the estimate of the transmitted data.

11. The wireless client device as in claim 10 wherein the receiver unit is a minimum mean square error (MMSE) receiver, zero-forcing (ZF) receiver, a maximum likelihood (ML) or a MAP receiver.

12. The system as in claim 7 wherein the MU-MAS communication comprises DIDO communication and wherein the one or more RF units receive signals transmitted from one or more DIDO transmitter units, and downconvert the signals to baseband.

13. The system as in claim 7 wherein the frequency/phase offset estimation/compensation unit carries out means for coarse frame synchronization, means for fractional frequency offset correction, means for integer frequency offset correction and/or means for integer frequency offset correction.

14. A method for compensating for in-phase and quadrature (I/Q) imbalances in a multiple antenna system (MAS) with multi-user (MU) transmissions ("MU-MAS") comprising:

transmitting a training signal between each antenna of a base station and each of a plurality of wireless client devices, analyzing each training signal to generate channel characterization data, and receiving the channel characterization data at the base station;

computing a plurality of MU-MAS precoder weights based on the channel characterization data, the MU-MAS precoder weights calculated to pre-cancel interference due to I/Q gain and phase imbalances and/or inter-user interference;

precoding data using the MU-MAS precoder weights to generate precoded data signals for each antenna of the base station; and

transmitting the precoded data signals through each antenna of the base station to each respective client device.

15. The method as in claim 14, wherein the base station is an access point coupling the wireless client devices to a wide area network.

16. The method as in claim 14 further comprising:

demodulating the data streams at each user device using zero-forcing (ZF), minimum mean-square error (MMSE) or maximum likelihood (ML) receiver to suppress residual interference.

17. The method as in claim 14 wherein precoding is performed using block diagonalization (BD) techniques.

18. The method of claim 14 wherein the MU-MAS system is a distributed-input distributed-output (DIDO) communication system and wherein the MU-MAS precoder weights are DIDO precoder weights.

19. The method as in claim 18 wherein the precoder weights are calculated to cancel inter-user interference but not ICI and wherein the wireless client devices include receivers with filters for cancelling the ICI.

20. A system for compensating for in-phase and quadrature (I/Q) imbalances for MU-MAS communication comprising:

one or more coding modulation units to encode and modulate information bits for each of a plurality of wireless client devices to produce encoded and modulated information bits;

one or more mapping units to map the encoded and modulated information bits to complex symbols; and

a MU-MAS IQ-aware precoding unit to exploit channel state information to compute MU-MAS IQ-aware precoding weights, the MU-MAS IQ-aware precoding unit precoding the complex symbols obtained from the mapping units using the weights to pre-cancel interference due to I/Q gain and phase imbalances and/or inter-user interference.

21. The system as in claim 20 further comprising:

one or more orthogonal frequency division multiplexing (OFDM) units to receive the precoded signals from the MU-MAS IQ-aware precoding unit and to modulate the precoded signals according to an OFDM standard.

22. The system as in claim 21 wherein the OFDM standard comprises computing an inverse fast Fourier transform (IFFT) and adding a cyclic prefix.

23. The system as in claim 21 further comprising one or more D/A units to perform digital to analog (D/A) conversion on the output of the OFDM units to generate an analog baseband signal; and

one or more radio frequency (RF) units to upconvert the baseband signal to radio frequency and transmit the signals using a corresponding one or more transmit antennas.

24. The system as in claim 20 wherein the MU-MAS IQ-aware precoding unit is implemented as a minimum mean square error (MMSE), a weighted MMSE or a zero-forcing (ZF) precoder.

25. The system as in claim 20 wherein the MU-MAS precoding unit is implemented as a minimum mean square error (MMSE), a weighted MMSE, a zero-forcing (ZF) or a block diagonalization (BD) precoder.

26. The system as in claim 20 wherein the MU-MAS communication comprises DIDO communication and wherein the system comprises a DIDO IQ-aware precoding unit to exploit channel state information to compute DIDO IQ-aware precoding weights, the DIDO IQ-aware precoding unit precoding the complex symbols obtained from the mapping units using the weights to pre-cancel interference due to I/Q gain and phase imbalances and/or inter-user interference.

27. A wireless client device for use in a system for compensating for in-phase and quadrature (I/Q) imbalances for MU-MAS communication comprising:

one or more RF units to receive signals transmitted from one or more MU-MAS transmitter units, and downconvert the signals to baseband;

one or more analog to digital (A/D) conversion units to receive the downconverted signals and to convert the signals from analog signals to digital signals;

one or more OFDM units to remove cyclic prefix and perform a fast Fourier transform (FFT) on the digital signals to report the signals in the frequency domain;

an I/Q-aware channel estimate unit which receives a signal output from the one or more OFDM units during a training period and responsively calculates channel estimate data; and

a feedback generator unit to transmit the channel estimate data or a training signals to a base station for use in precoding signals prior to transmission to the wireless client device.

28. The system as in claim 27 where the channel estimates are computed in the time domain by using the input to the OFDM units.

29. The wireless client device as in claim 27 wherein the feedback generator unit further comprises logic for quantizing the channel estimates prior to transmission to the base station.

30. The wireless client device as in claim 27 further comprising:

an I/Q-aware receiver unit which receives outputs from the OFDM units and responsively computes the IQ receiver and demodulates/decodes the signal to obtain the estimate of the transmitted data.

31. The wireless client device as in claim 29 wherein the I/Q-aware receiver unit is a minimum mean square error (MMSE) receiver, zero-forcing (ZF) receiver, a maximum likelihood (ML) or a MAP receiver.

32. The wireless client device as in claim 29 wherein the I/Q-aware receiver unit comprises an MMSE or ZF filter to cancel inter-carrier interference (ICI) caused by I/Q imbalance on mirror tones.

33. The wireless client device as in claim 29 wherein the I/Q-aware receiver unit comprises a nonlinear detector (i.e., ML) to jointly detect symbols on mirror tones.

34. The wireless client device as in claim 29 wherein the I/Q aware channel estimate unit calculates coefficients usable by the I/Q-aware receiver unit to remove inter-carrier interference (ICI).

35. The system as in claim 26 wherein the MU-MAS communication comprises DIDO communication and wherein the one or more RF units receive signals transmitted from one or more DIDO transmitter units, and downconvert the signals to baseband.

36. A wireless client device for use in a system for dynamically adapting the communication characteristics of a MU-MAS communication system comprising:

one or more RF units to receive signals transmitted from one or more MU-MAS transmitter units, and downconvert the signals to baseband;

one or more analog to digital (A/D) conversion units to receive the downconverted signals and to convert the signals from analog signals to digital signals;

one or more OFDM units to remove cyclic prefix and perform a fast Fourier transform (FFT) on the digital signals to report the signals in the frequency domain;

a channel estimator which receives a signal output from the one or more OFDM units during a training period and responsively calculates the link quality metrics; and

a feedback generator unit to transmit the link quality metrics to a base station for use in modulation/coding, precoding signals and user selection prior to transmission to the wireless client device.

37. The wireless client device as in claim 36 where the channel estimates are computed in the time domain by using the input to the OFDM units.

38. The wireless client device as in claim 36 wherein the feedback generator unit further comprises logic for quantizing the channel estimates and or link-quality metrics prior to transmission to the base station.

39. The wireless client device as in claim 36 further comprising:

a receiver unit which receives outputs from the OFDM units and responsively demodulates/decodes the signal to obtain the estimate of the transmitted data.

40. The wireless client device as in claim 38 wherein the receiver unit is a minimum mean square error (MMSE) receiver, zero-forcing (ZF) receiver, a maximum likelihood (ML) or a MAP receiver.

41. The wireless client device as in claim 37 wherein the MU-MAS configurator unit or channel estimator unit, respectively, employs polarization and/or pattern diversity techniques as a means to reduce the array size, while obtaining diversity over the wireless link.

42. The wireless client device as in claim 36 wherein communication occurs via NVIS and/or groundwave as a means to increase diversity and downlink throughput.

43. The wireless client device as in claim 36 wherein pattern diversity are employed to communicate via ground-wave with certain users and NVIS with other users.

44. The wireless client device as in claim 43 wherein each client exploits spatial separation of the ground-wave and NVIS links as a means to increase the spatial diversity of the link.

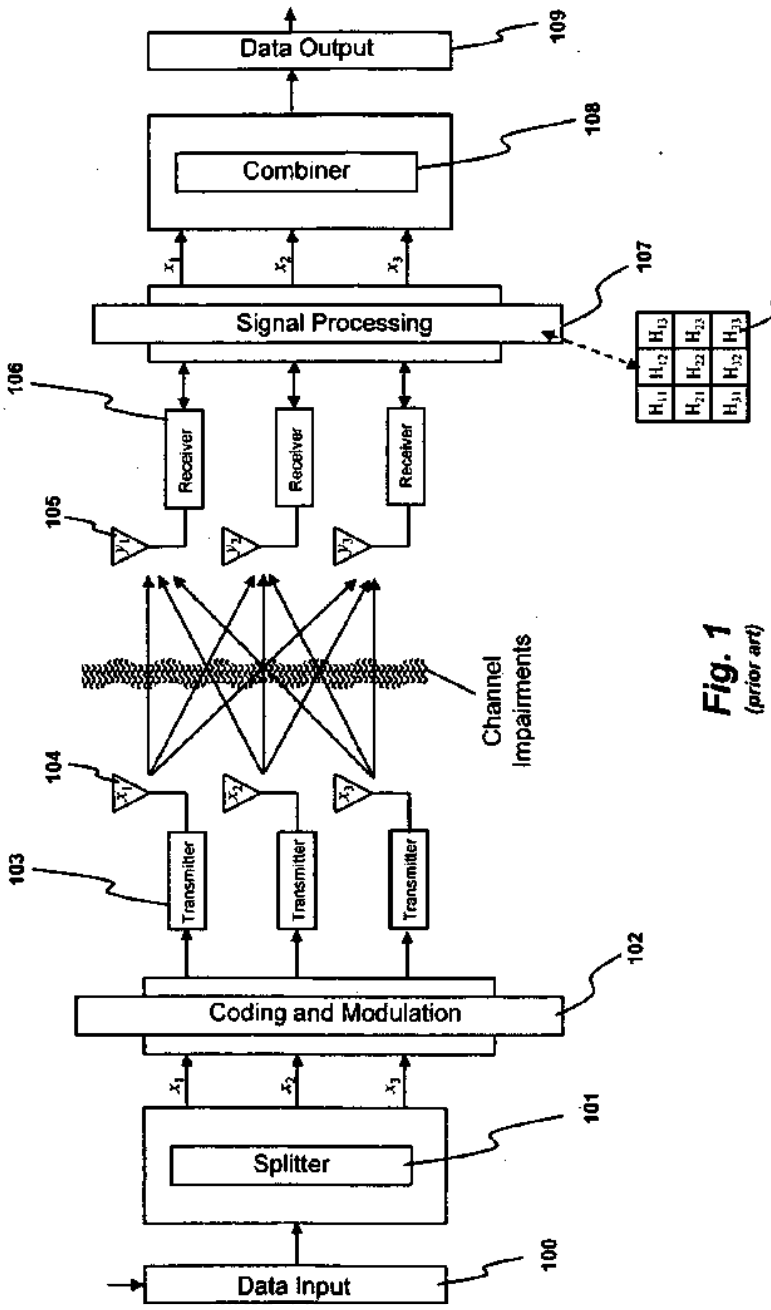
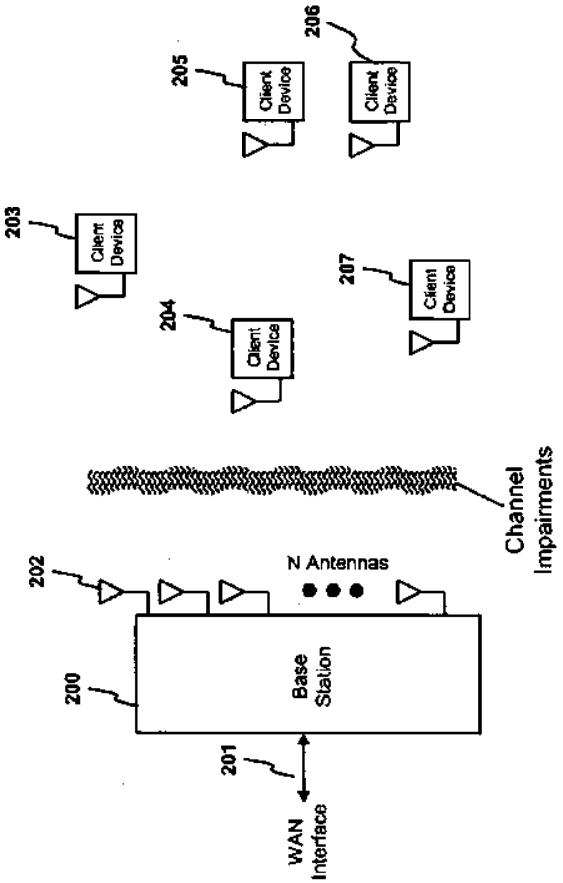
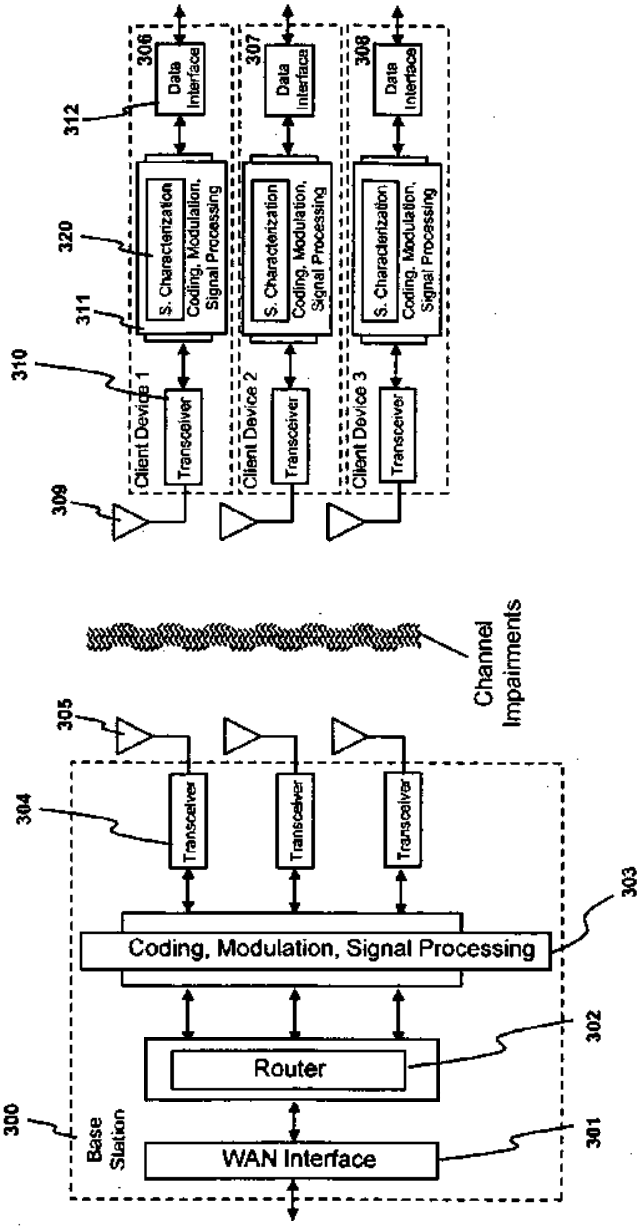


Fig. 1
(prior art)



N-antenna Base Station with Single-antenna Client Devices

Fig. 2



3-Antenna Base Station with 3 Single-Antenna Client Devices

Fig. 3

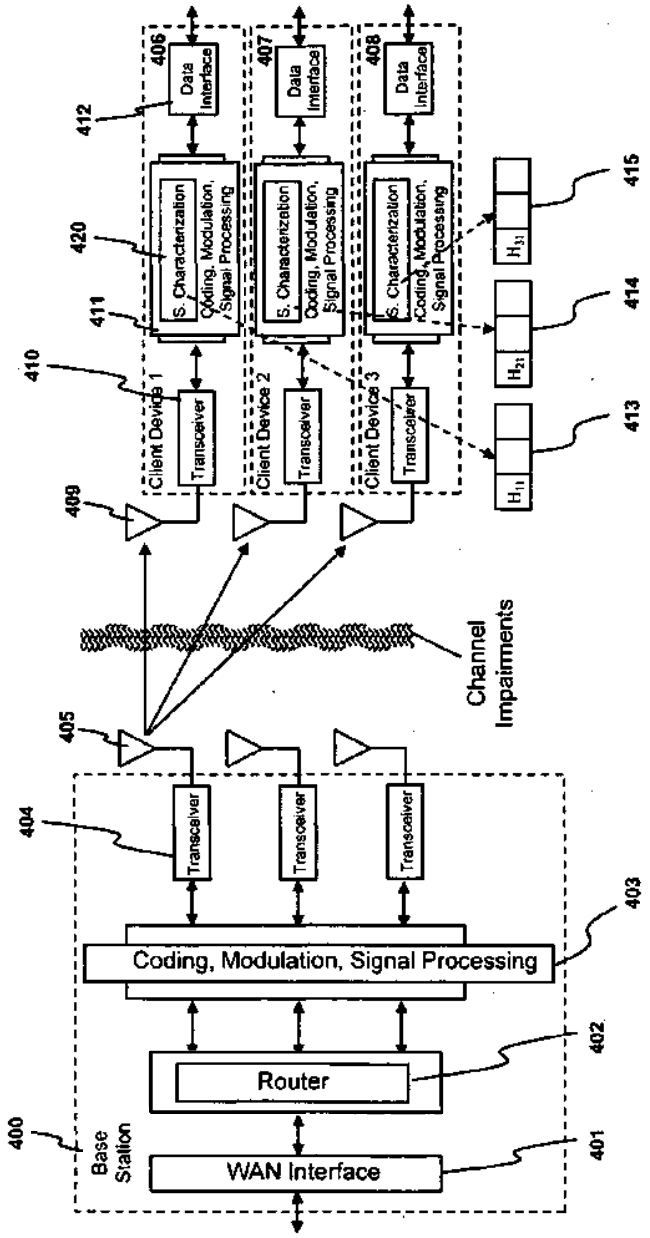

Base Station
400
WAN Interface
401
Router
402
Coding, Modulation, Signal Processing
403
Transceiver
404
405
Channel Impairments
Client Device 1
Client Device 2
Client Device 3
Transceiver
410
411
Data Interface
412
S. Characterization
Coding, Modulation, Signal Processing
420
413
414
415
406
407
408
409
Antenna 1 Pilot Signal Transmission

Fig. 4

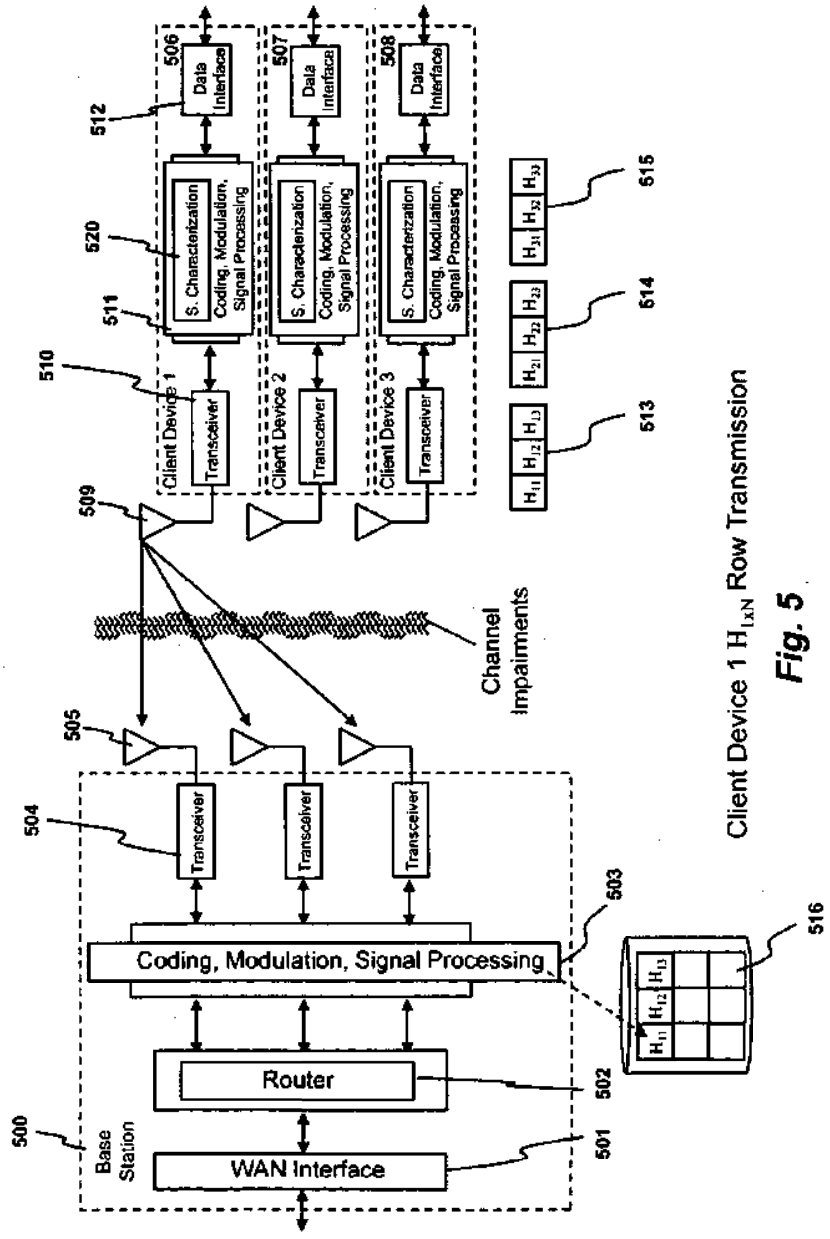
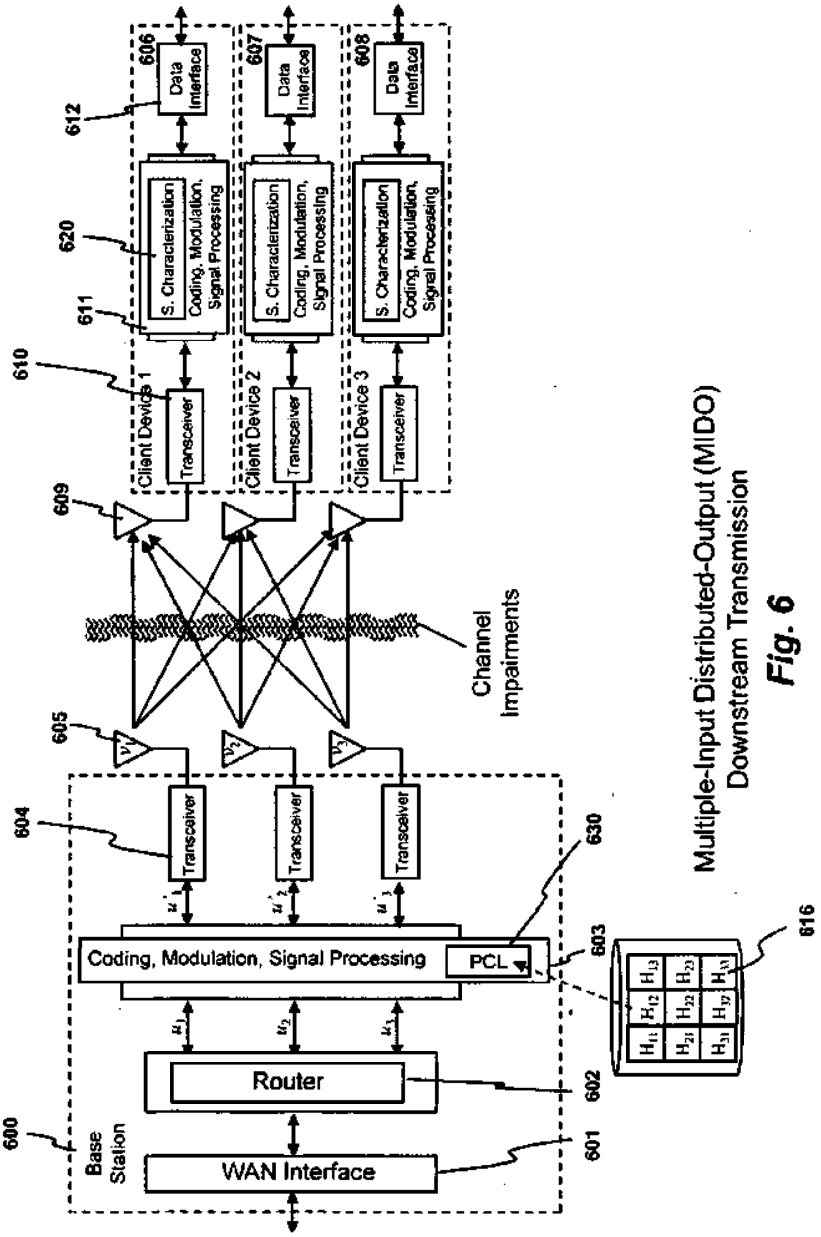

Client Device 1
Client Device 2
Client Device 3
Data Interface
S. Characterization
Coding, Modulation, Signal Processing
Transceiver
Channel Impairments
Coding, Modulation, Signal Processing
Router
WAN Interface
Base Station
500
501
502
503
504
505
506
507
508
509
510
511
512
513
514
515
516
520
Client Device 1 H_{1xN} Row Transmission

Fig. 5


Multiple-Input Distributed-Output (MIDO)
Downstream Transmission
Fig. 6
Base Station
600
WAN Interface
601
Router
602
Coding, Modulation, Signal Processing
PCL
603
Transceiver
604
605
Channel Impairments
609
Client Device 1
Client Device 2
Client Device 3
610
611
612
S. Characterization
Coding, Modulation, Signal Processing
620
Data Interface
606
607
608
630
616

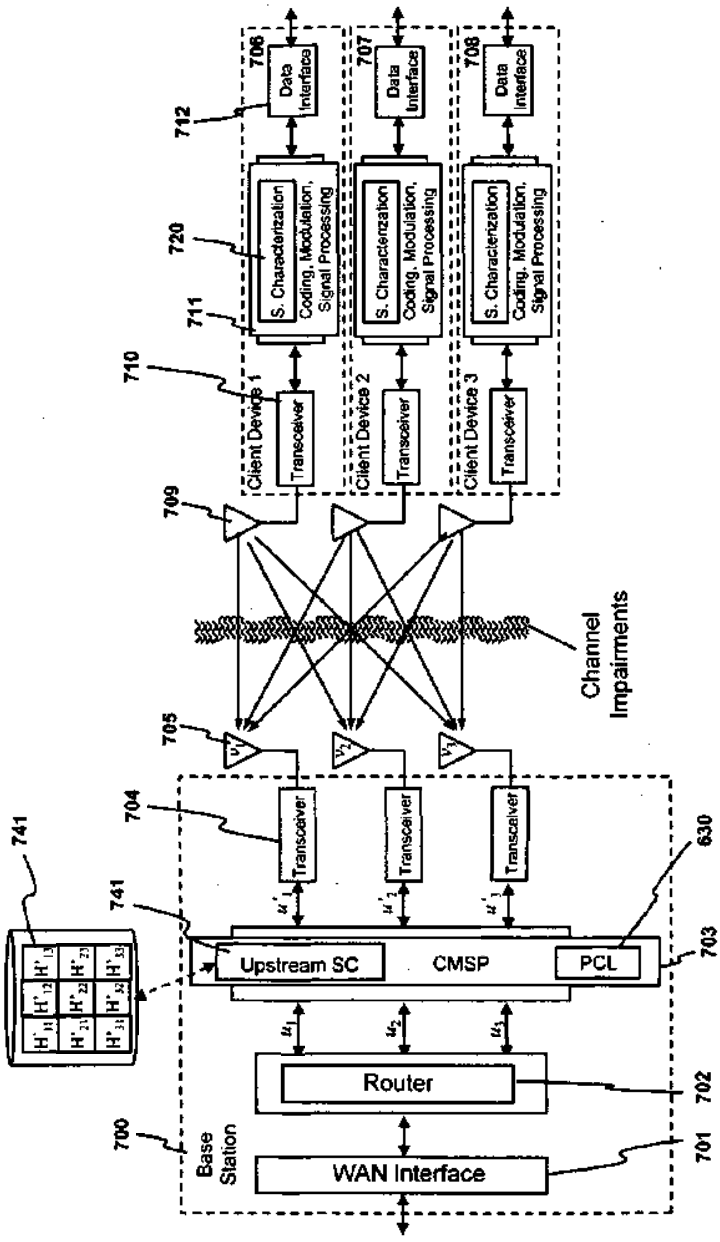
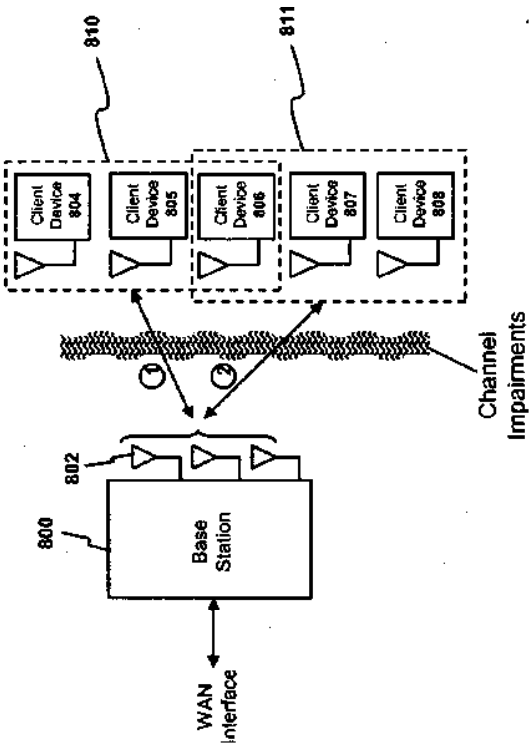
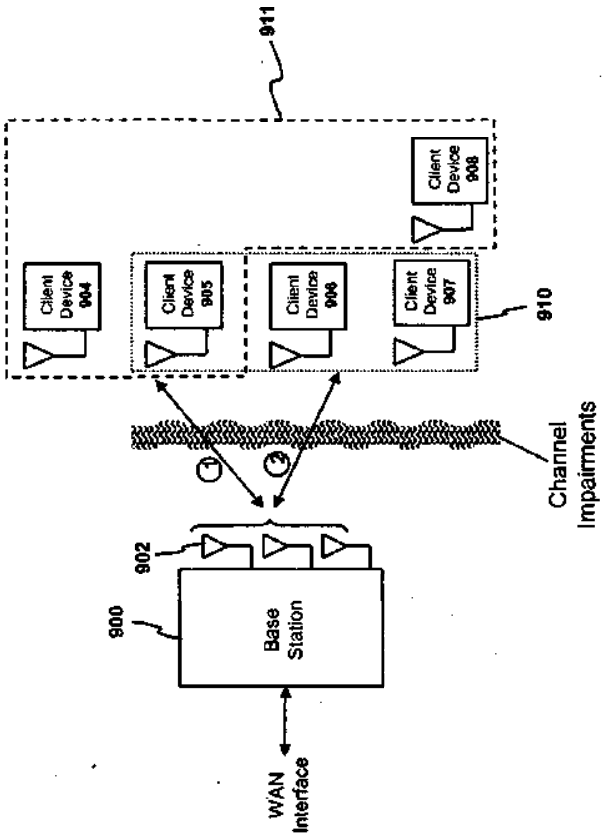

MIMO Upstream Transmission
Client Device 1
Client Device 2
Client Device 3
Data Interface
S. Characterization
Coding, Modulation, Signal Processing
Transceiver
Channel Impairments
Upstream SC
CMSP
PCL
Router
WAN Interface
Base Station
712
720
711
710
709
705
704
741
630
703
702
701
700

Fig. 7



Cycling With Different Client Groups

Fig. 8



Grouping Clients Based on Proximity

Fig. 9

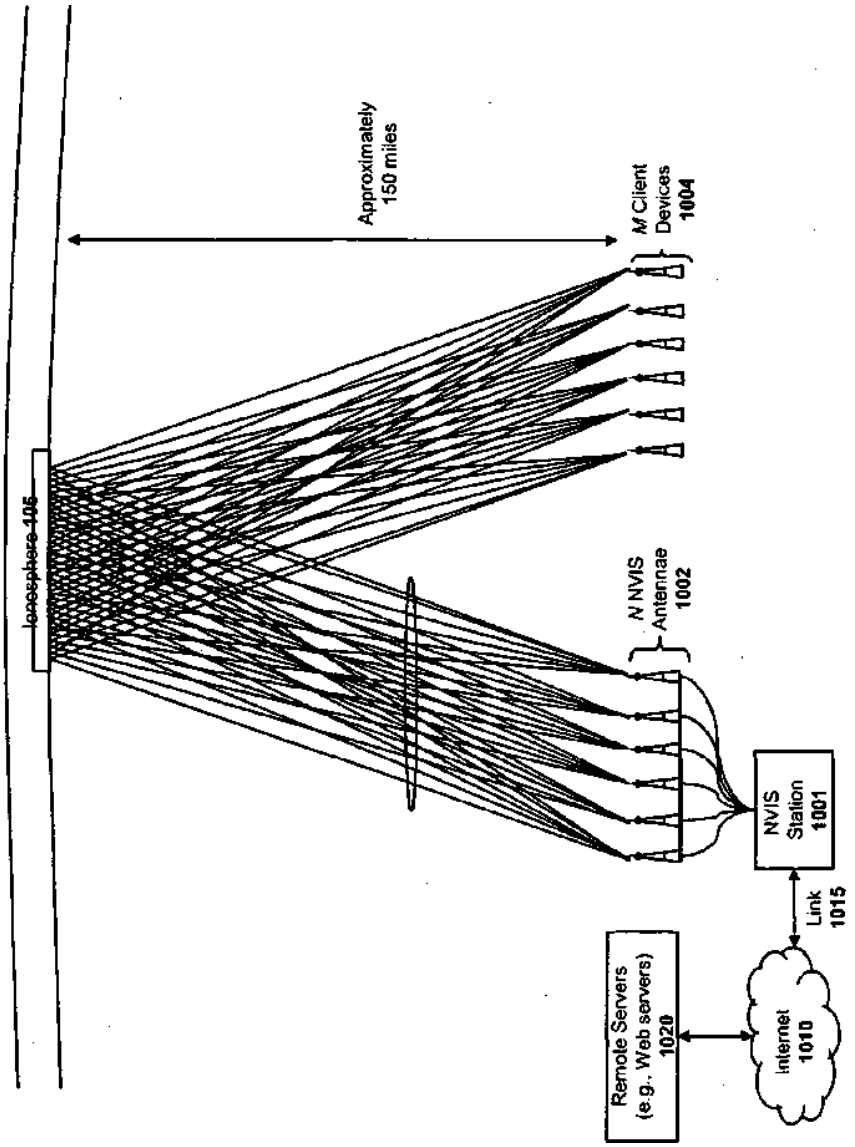

Approximately
150 miles
M Client
Devices
1004
Ionosphere 105
N NVIS
Antennae
1002
NVIS
Station
1001
Link
1015
Remote Servers
(e.g., Web servers)
1020
Internet
1010

Fig. 10

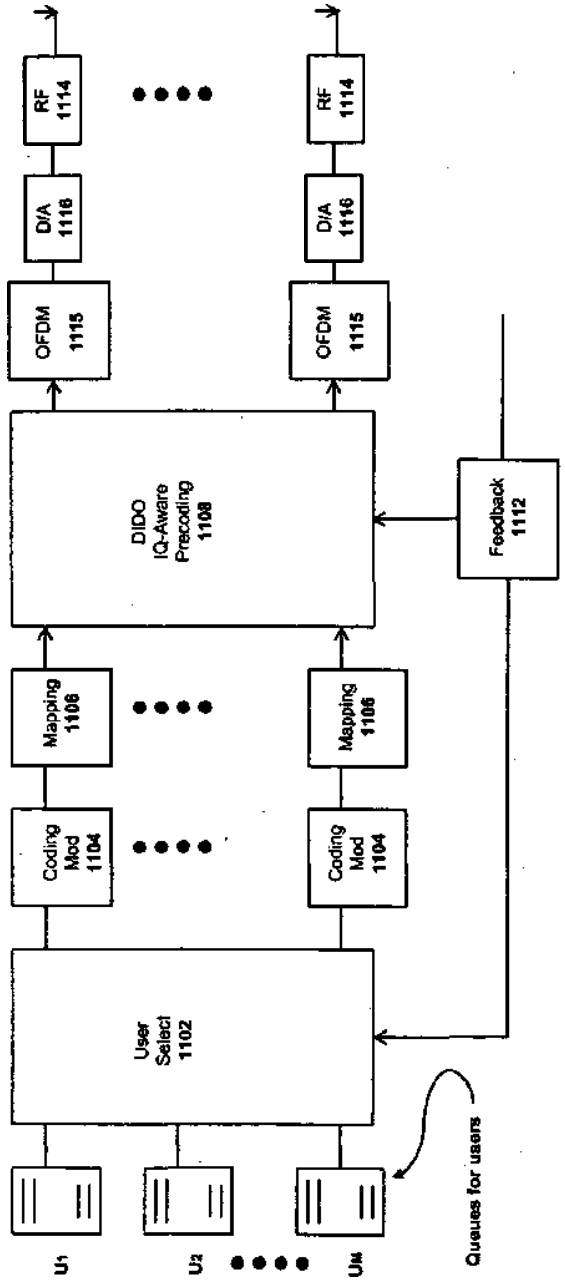

RF 1114
RF 1114
D/A 1116
D/A 1116
OFDM 1115
OFDM 1115
DIDO IQ-Aware Precoding 1108
Feedback 1112
Mapping 1106
Mapping 1106
Coding Mod 1104
Coding Mod 1104
User Select 1102
Queues for users
U1
U2
UM

FIG. 11

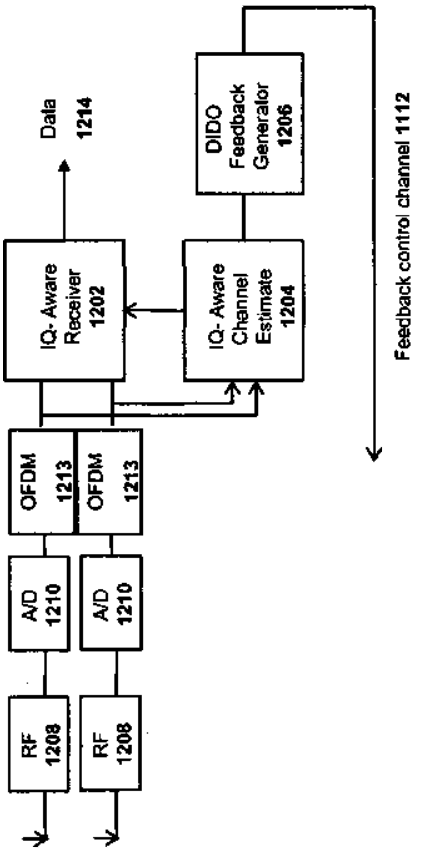

Data
1214
DIDO
Feedback
Generator
1206
Feedback control channel 1112
IQ- Aware
Receiver
1202
IQ- Aware
Channel
Estimate
1204
OFDM
1213
OFDM
1213
A/D
1210
A/D
1210
RF
1208
RF
1208

FIG. 12

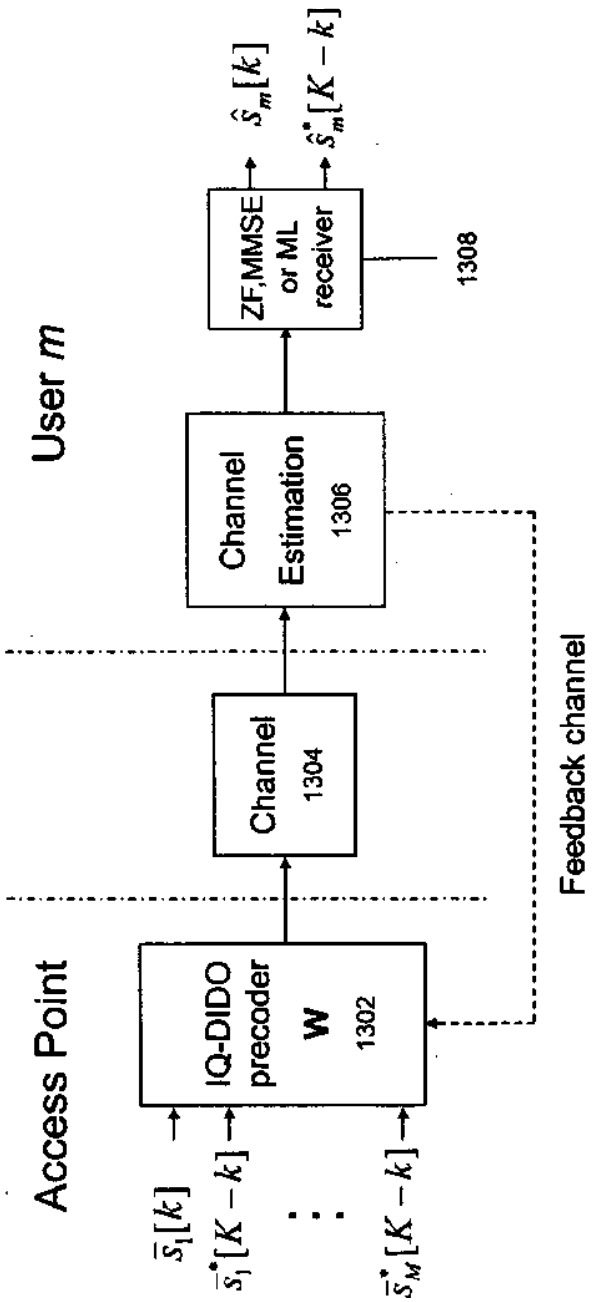

Access Point
IQ-DIDO precoder W 1302
Channel 1304
Channel Estimation 1306
User m
ZF,MMSE or ML receiver
1308
Feedback channel
$\bar{s}_1[k]$
$\bar{s}_1^*[K-k]$
$\bar{s}_M^*[K-k]$
$\hat{s}_m[k]$
$\hat{s}_m^*[K-k]$

FIG. 13

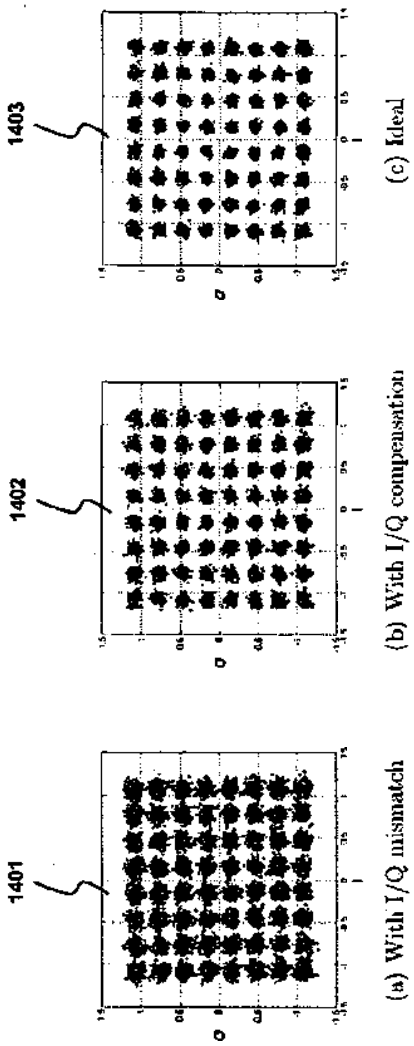

1403
(c) Ideal
1402
(b) With I/Q compensation
1401
(a) With I/Q mismatch

FIG. 14

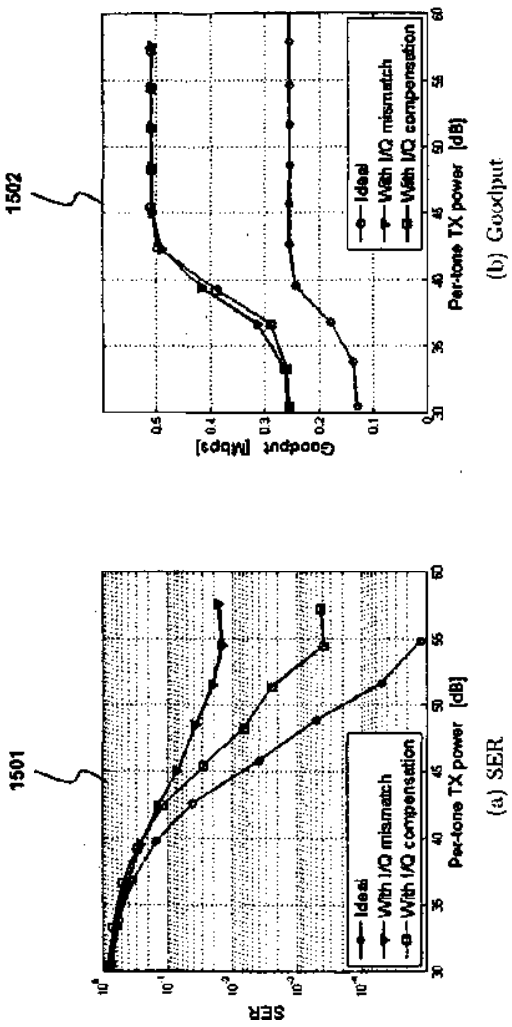

1502
Goodput [Mbps]
Per-tone TX power [dB]
Ideal
With I/Q mismatch
With I/Q compensation
(b) Goodput
1501
SER
Per-tone TX power [dB]
Ideal
With I/Q mismatch
With I/Q compensation
(a) SER

FIG. 15

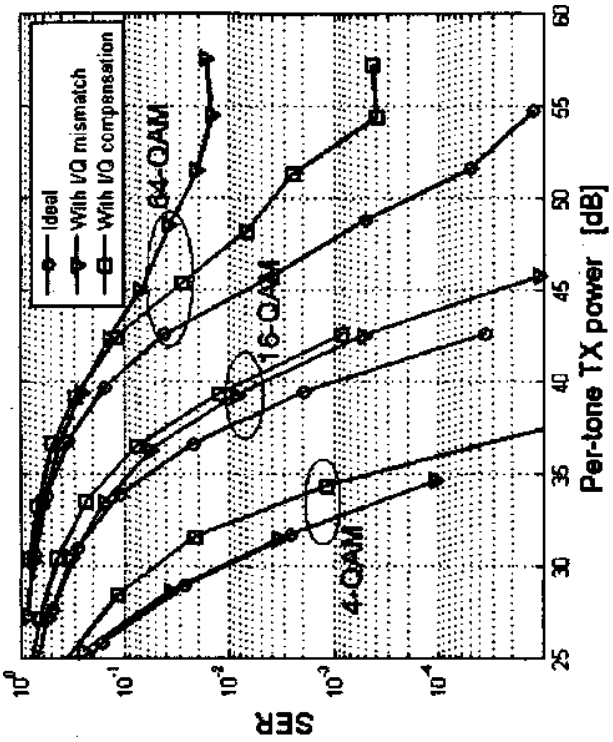

Ideal
With I/Q mismatch
With I/Q compensation
64-QAM
16-QAM
4-QAM
SER
Per-tone TX power [dB]

FIG. 16

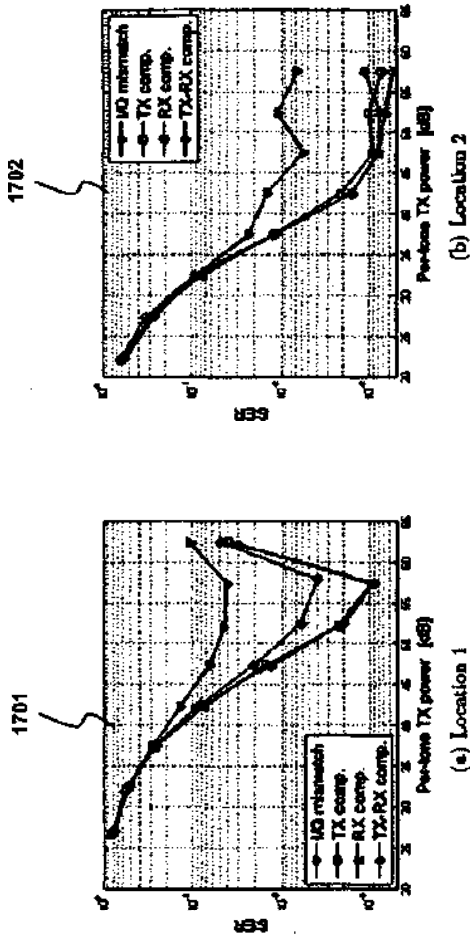

1702
1701
I/Q mismatch
TX comp.
RX comp.
TX-RX comp.
BER
Per-tone TX power [dB]
(b) Location 2
(a) Location 1

FIG. 17

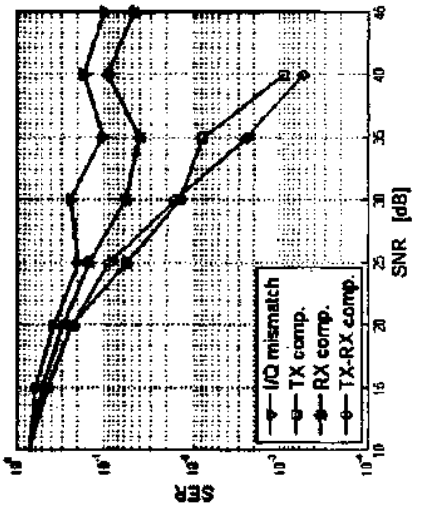


FIG. 18

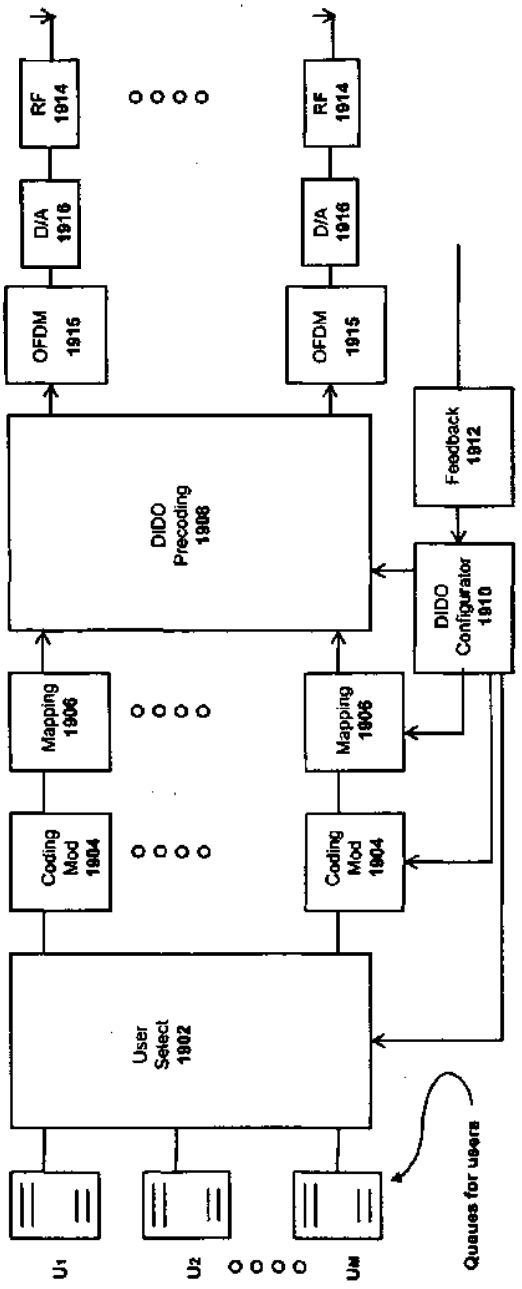

RF 1914
RF 1914
D/A 1916
D/A 1916
OFDM 1915
OFDM 1915
DIDO Precoding 1908
Feedback 1912
DIDO Configurator 1910
Mapping 1906
Mapping 1906
Coding Mod 1904
Coding Mod 1904
User Select 1902
Queues for users
U1
U2
UM

FIG. 19

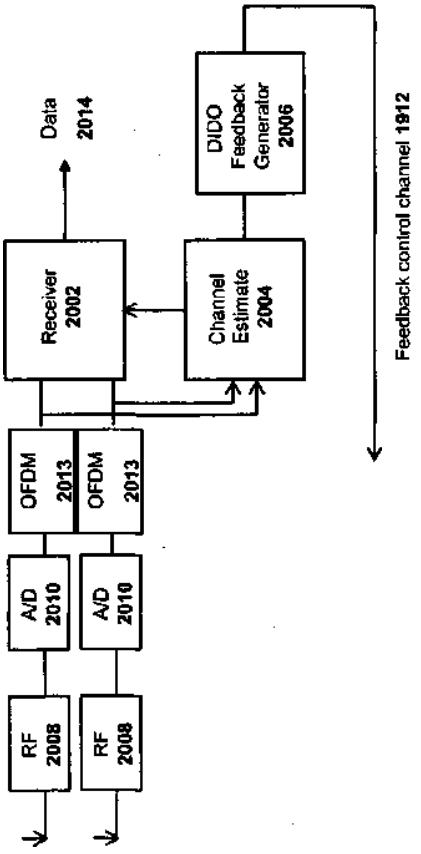

Data
2014
DIDO
Feedback
Generator
2006
Feedback control channel 1912
Receiver
2002
Channel
Estimate
2004
OFDM
2013
OFDM
2013
A/D
2010
A/D
2010
RF
2008
RF
2008

FIG. 20

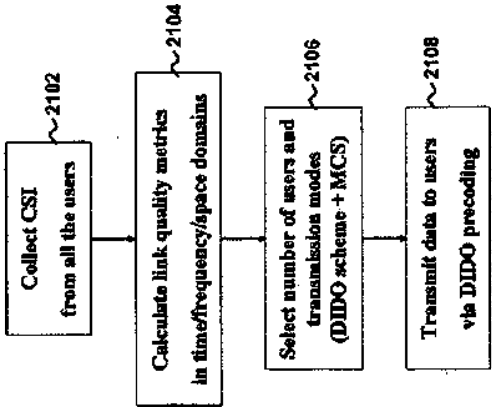

2102
Collect CSI from all the users
2104
Calculate link quality metrics in time/frequency/space domains
2106
Select number of users and transmission modes (DIDO scheme + MCS)
2108
Transmit data to users via DIDO precoding

FIG. 21

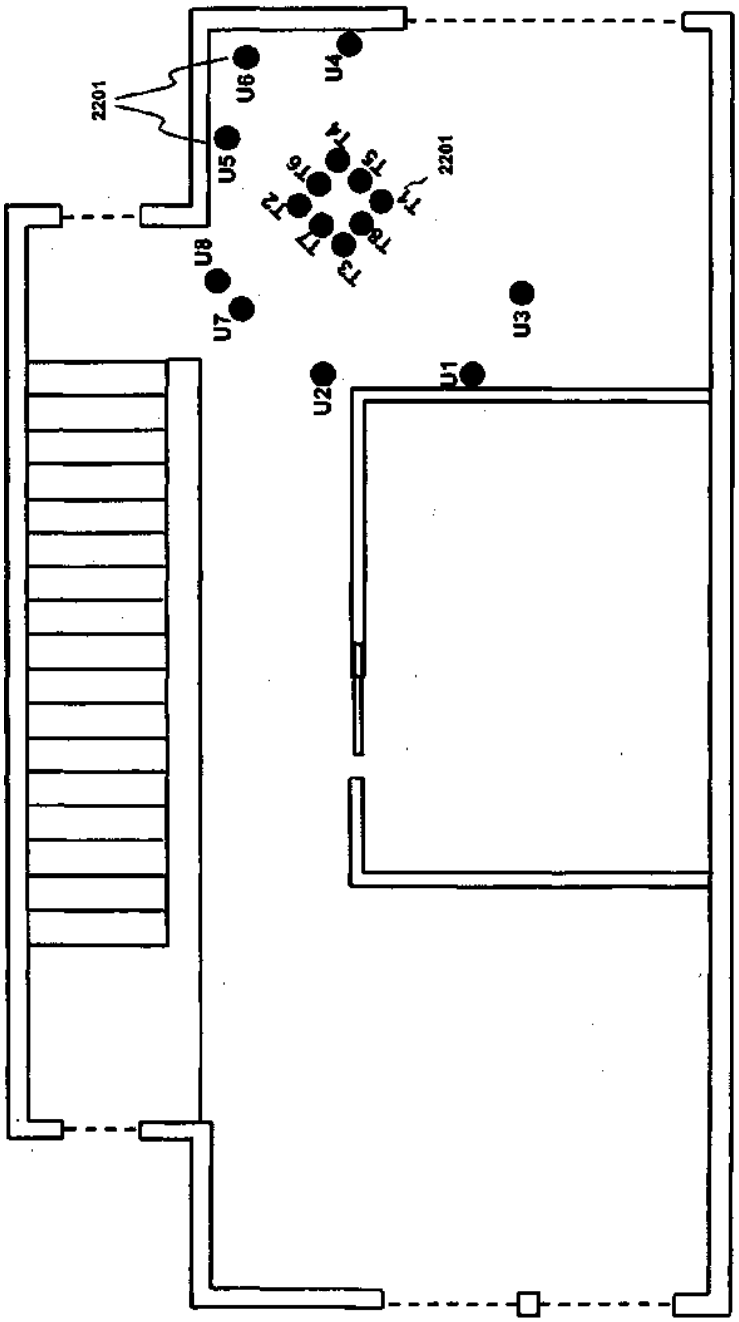


FIG. 22

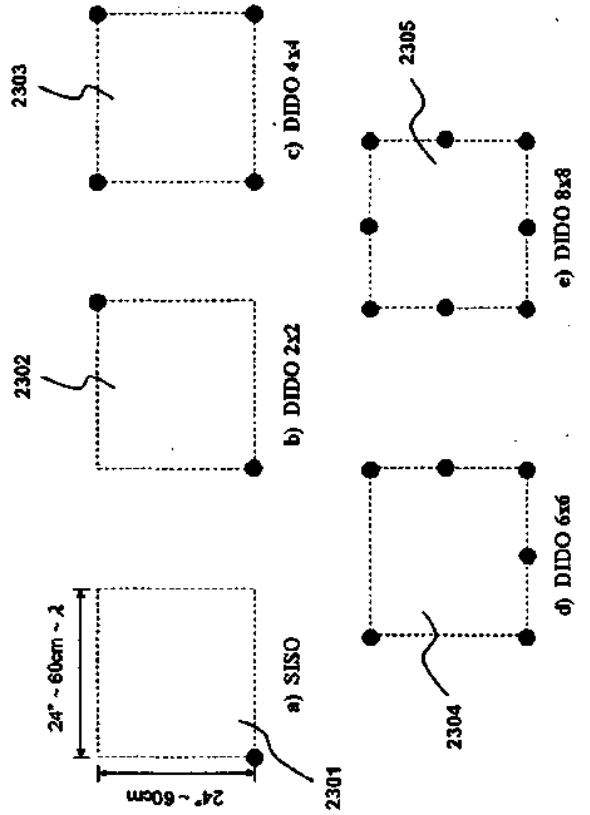

2303
c) DIDO 4x4
2305
e) DIDO 8x8
2302
b) DIDO 2x2
d) DIDO 6x6
24" ~ 60cm ~ λ
a) SISO
24" ~ 60cm
2301
2304

FIG. 23

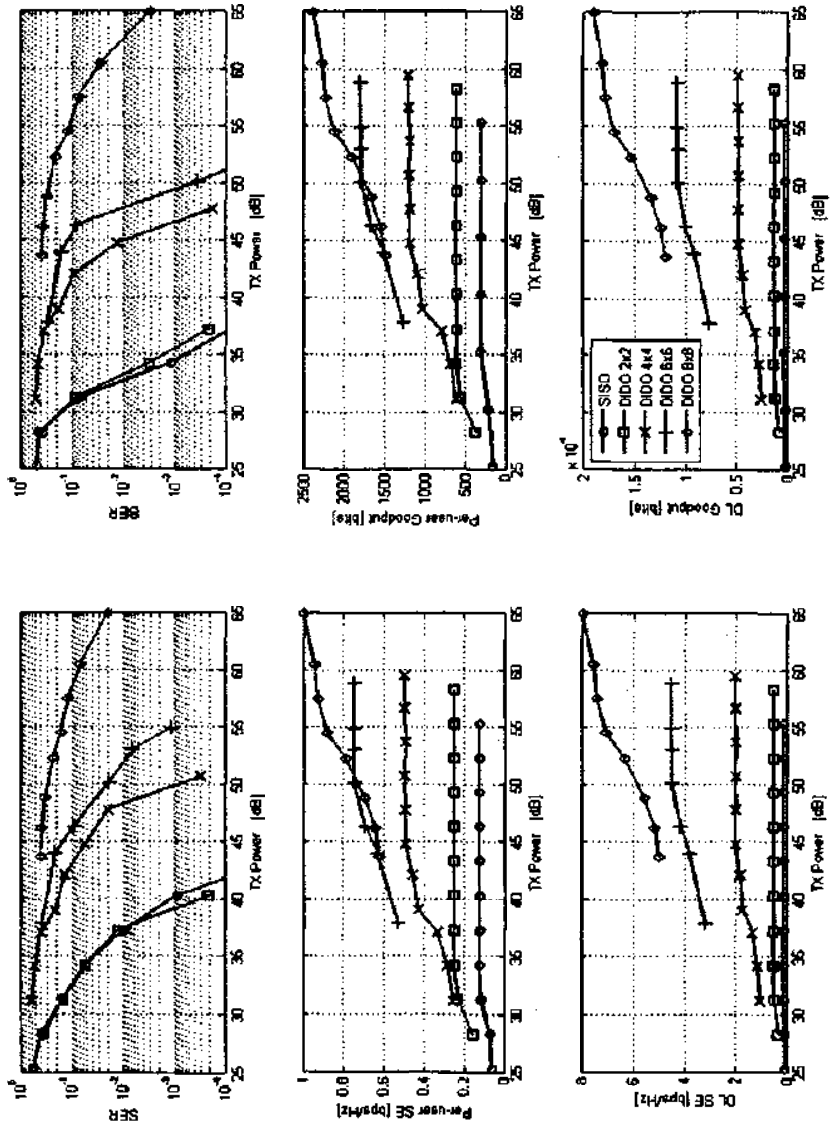

SER
TX Power [dB]
Per-user Goodput [bits]
DL Goodput [bits]
SISO
DIDO 2x2
DIDO 4x4
DIDO 8x8
Per-user SE [bps/Hz]
DL SE [bps/Hz]

FIG. 24

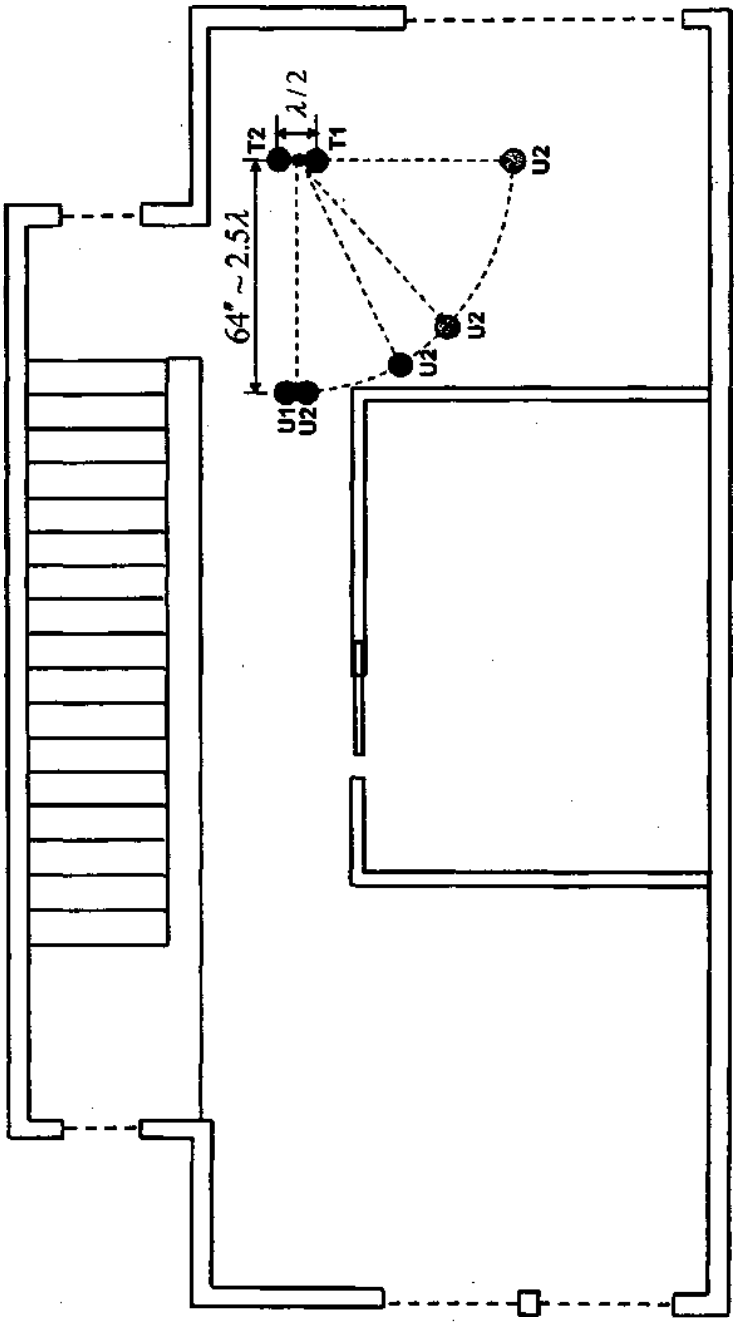


FIG. 25

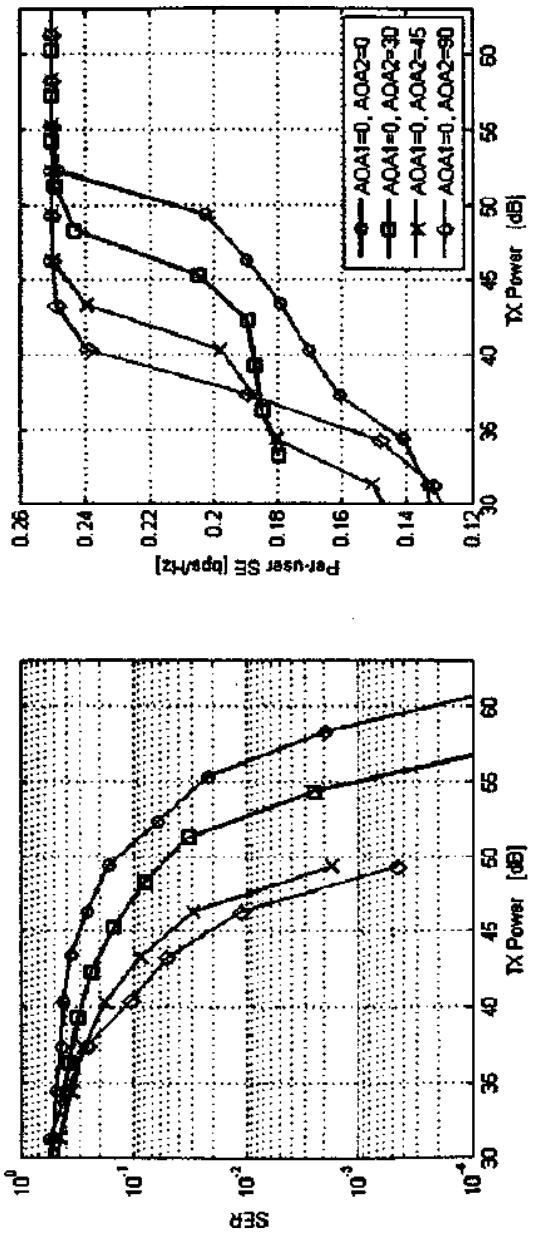

Per-user SE [bps/Hz]
TX Power [dB]
AOA1=0, AOA2=0
AOA1=0, AOA2=30
AOA1=0, AOA2=45
AOA1=0, AOA2=90
SER
TX Power [dB]

FIG. 26

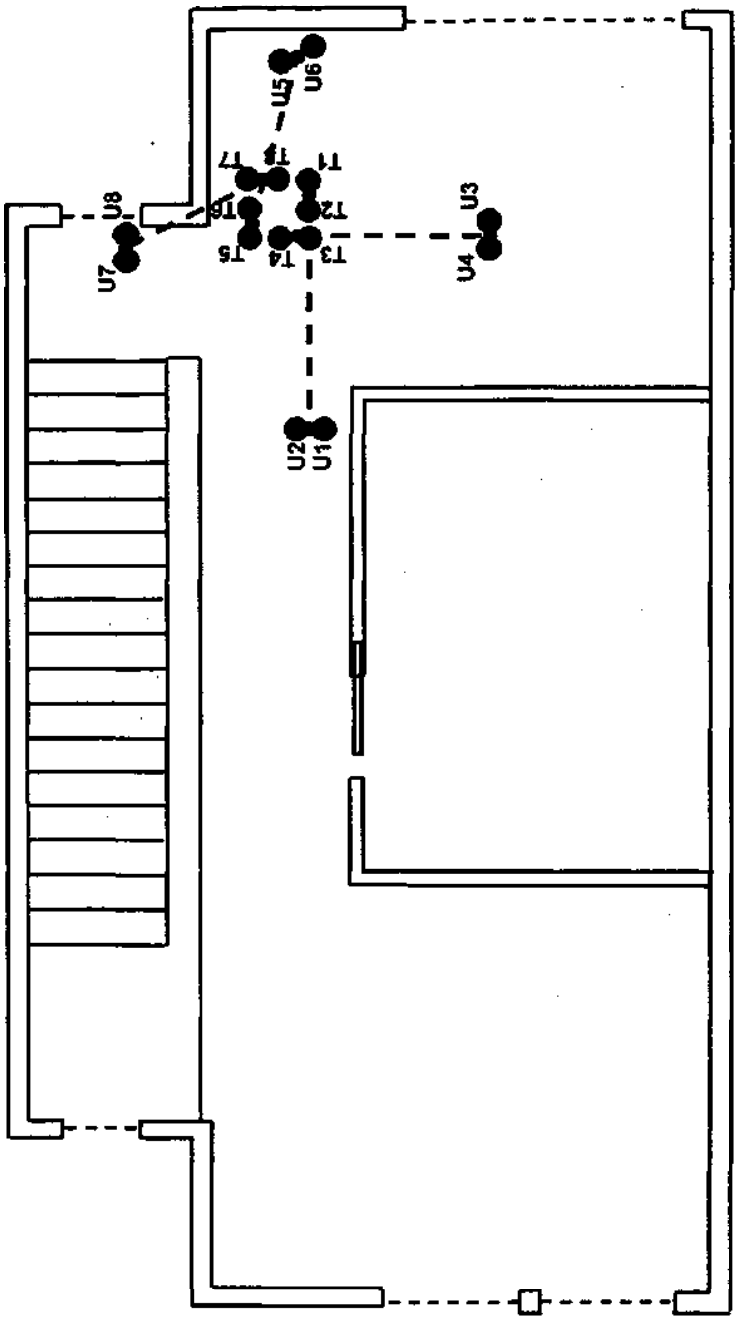


FIG. 27

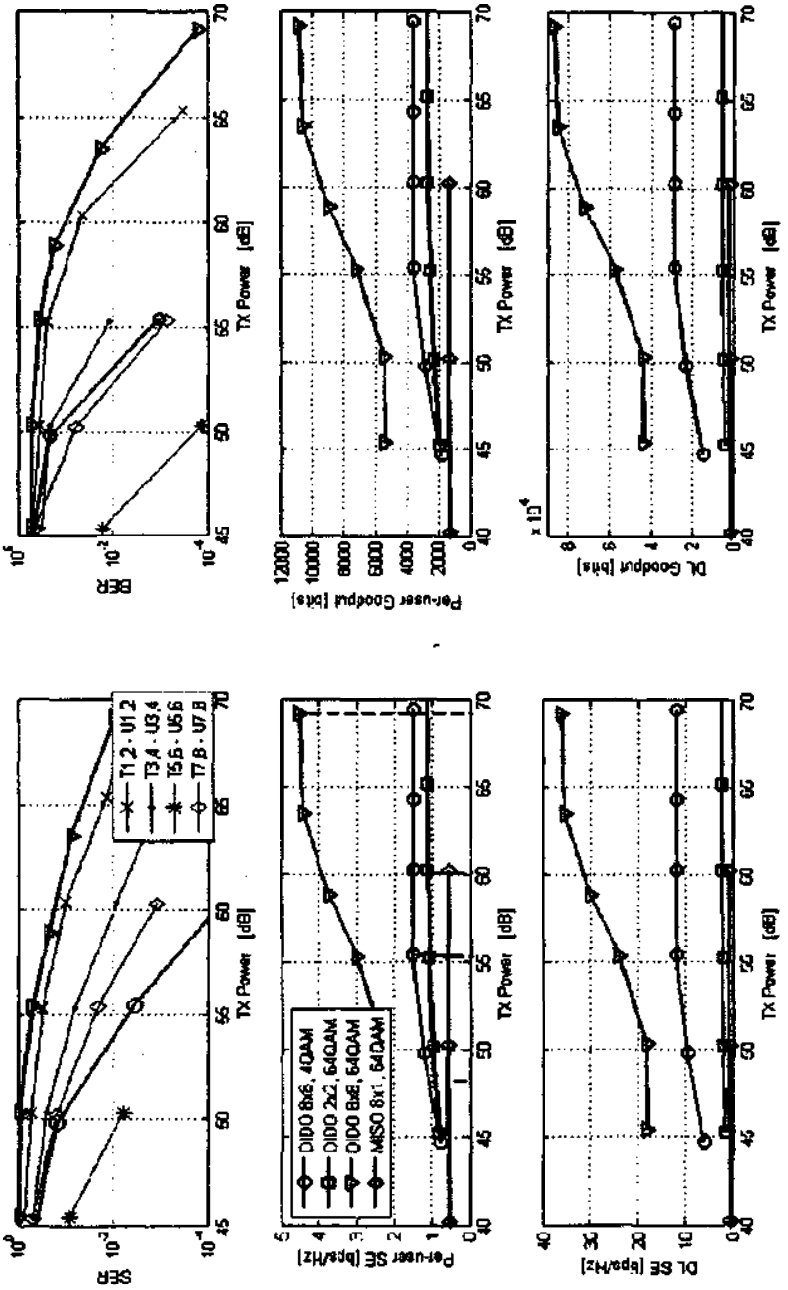

BER
Tx Power [dB]
Per-user Goodput [bits]
DL Goodput [bits]
SER
T1,2 - U1,2
T3,4 - U3,4
T5,6 - U5,6
T7,8 - U7,8
Per-user SE [bps/Hz]
DL SE [bps/Hz]
DIDO 8x8, 4QAM
DIDO 2x2, 64QAM
DIDO 8x8, 64QAM
MISO 8x1, 64QAM

FIG. 28

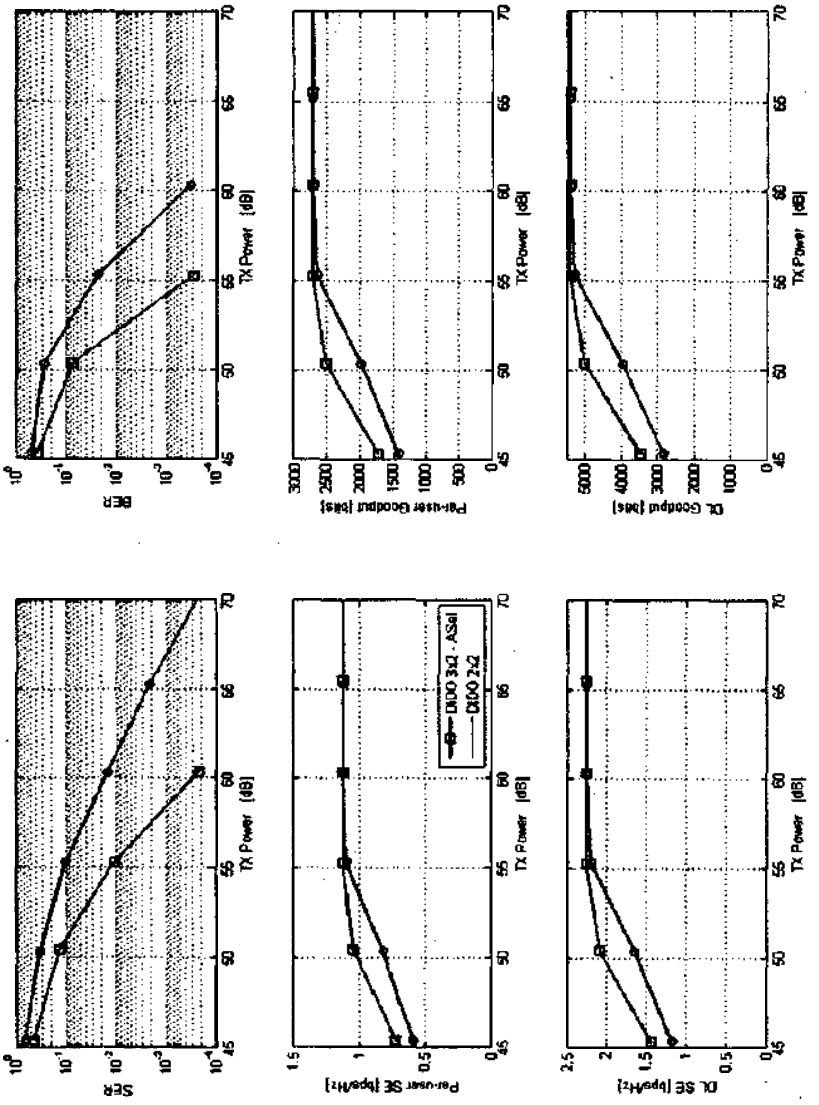

BER
SER
Per-user Goodput [bits]
DL Goodput [bits]
Per-user SE [bps/Hz]
DL SE [bps/Hz]
TX Power [dB]
DIDO 3x2 - ASel
DIDO 2x2

FIG. 29

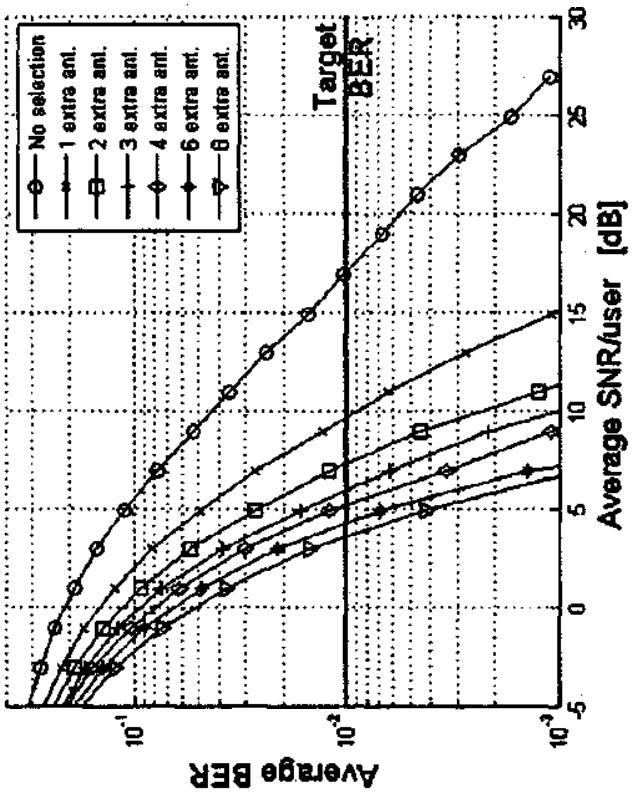

No selection
1 extra ant.
2 extra ant.
3 extra ant.
4 extra ant.
6 extra ant.
8 extra ant.
Target BER
Average BER
Average SNR/user [dB]
10^-1
10^-2
10^-3
-5
0
5
10
15
20
25
30

FIG. 30

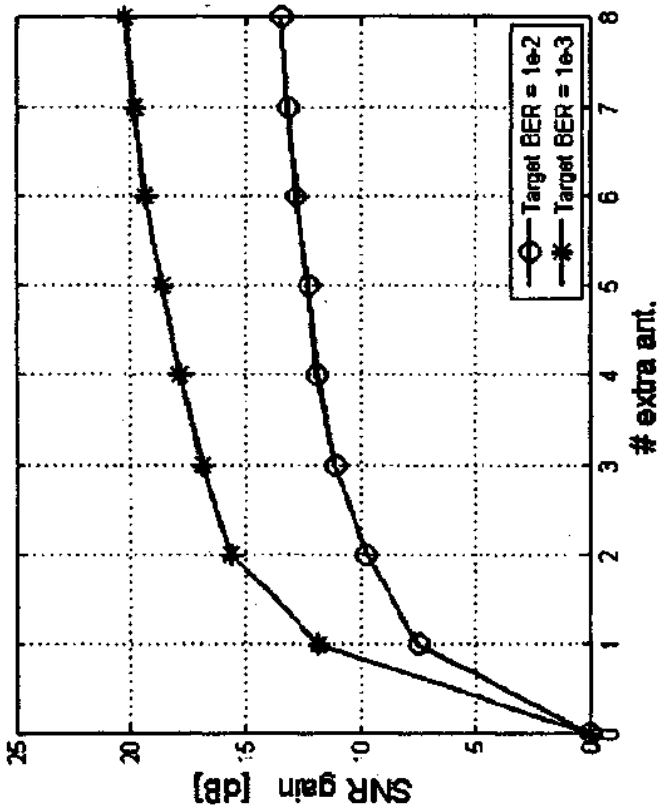

SNR gain [dB]
extra ant.
Target BER = 1e-2
Target BER = 1e-3
0
5
10
15
20
25
1
2
3
4
5
6
7
8

FIG. 31

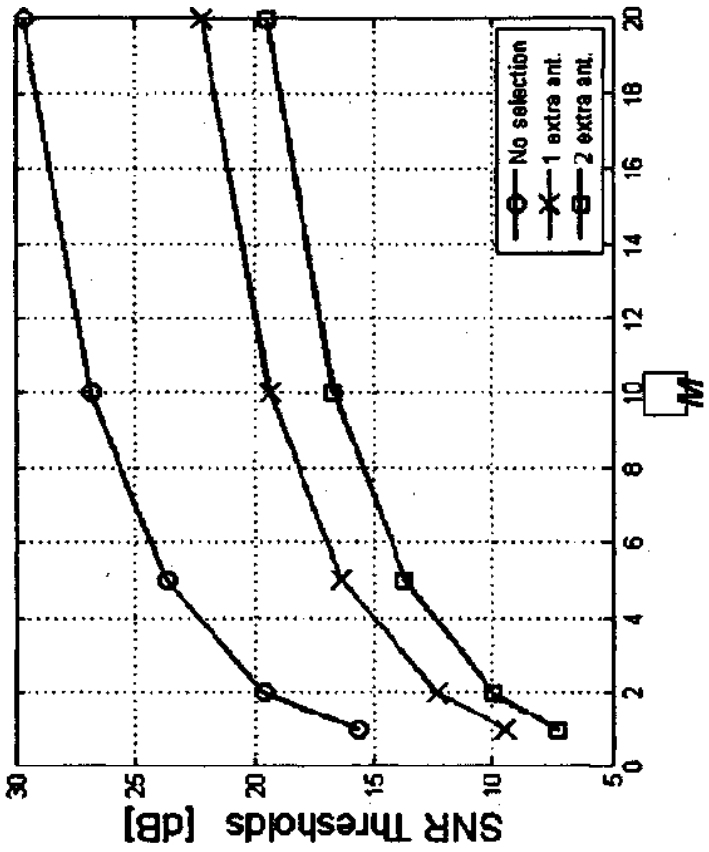


FIG. 32

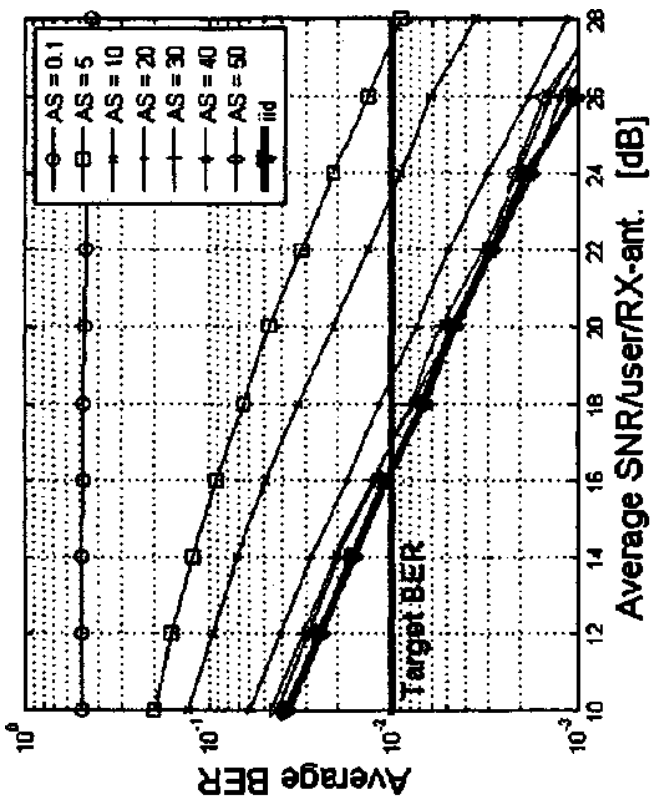

AS = 0.1
AS = 5
AS = 10
AS = 20
AS = 30
AS = 40
AS = 50
iid
Target BER
Average BER
Average SNR/user/RX-ant. [dB]
10^0
10^-1
10^-2
10^-3
10
12
14
16
18
20
22
24
26
28

FIG. 33

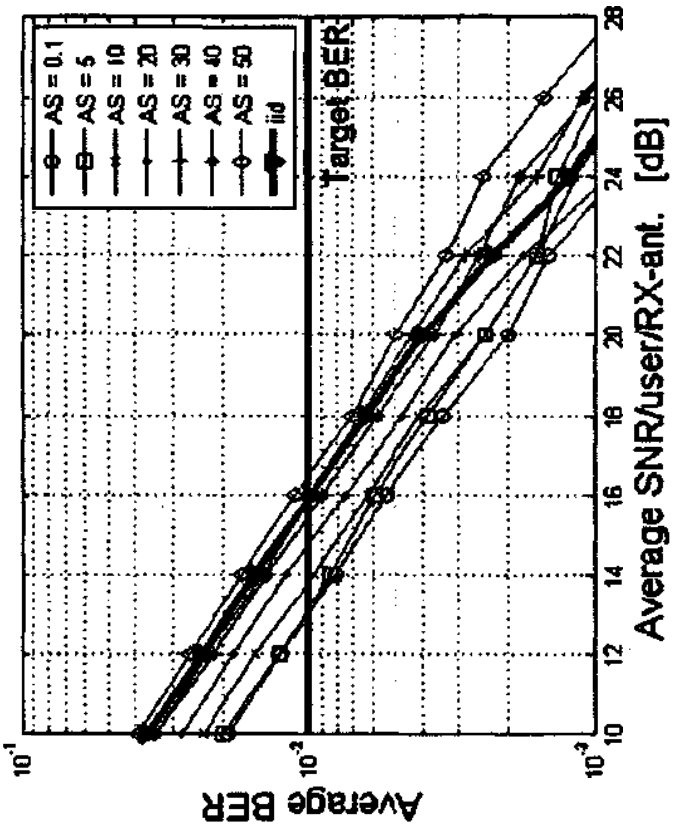

AS = 0.1
AS = 5
AS = 10
AS = 20
AS = 30
AS = 40
AS = 50
iid
Target BER
Average BER
Average SNR/user/RX-ant. [dB]

FIG. 34

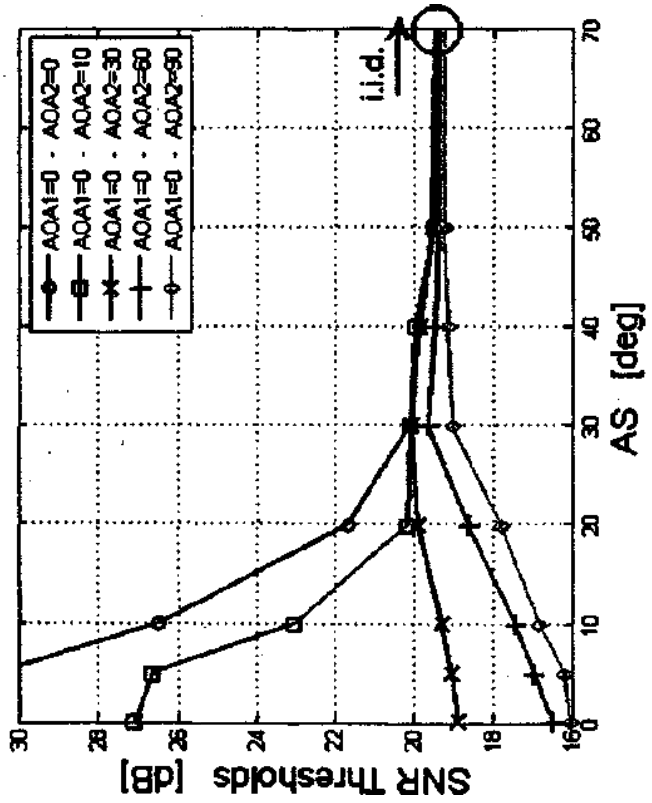

AOA1=0 - AOA2=0
AOA1=0 - AOA2=10
AOA1=0 - AOA2=30
AOA1=0 - AOA2=60
AOA1=0 - AOA2=90
i.i.d.
SNR Thresholds [dB]
AS [deg]

FIG. 35

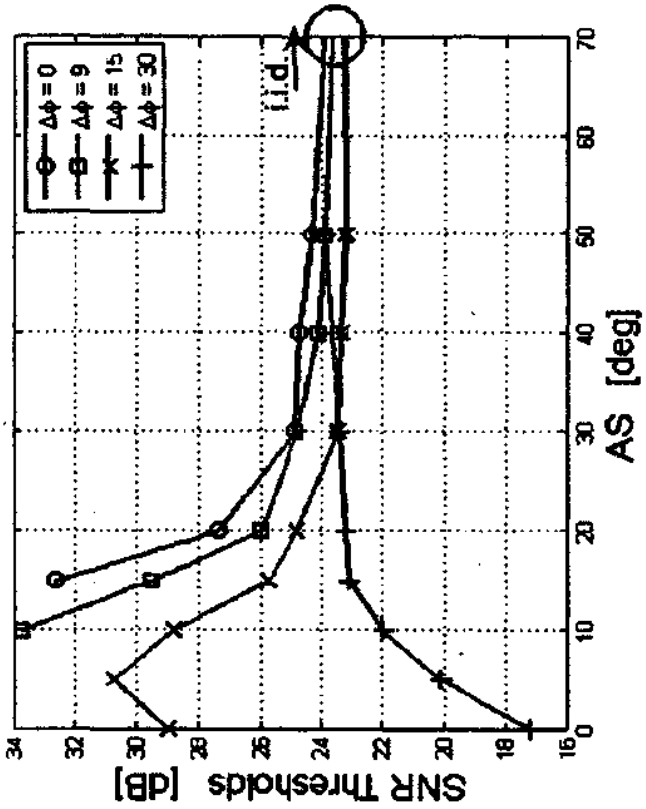

SNR Thresholds [dB]
AS [deg]
i.i.d.
Δφ = 0
Δφ = 9
Δφ = 15
Δφ = 30
34
32
30
28
26
24
22
20
18
16
0
10
20
30
40
50
60
70

FIG. 36

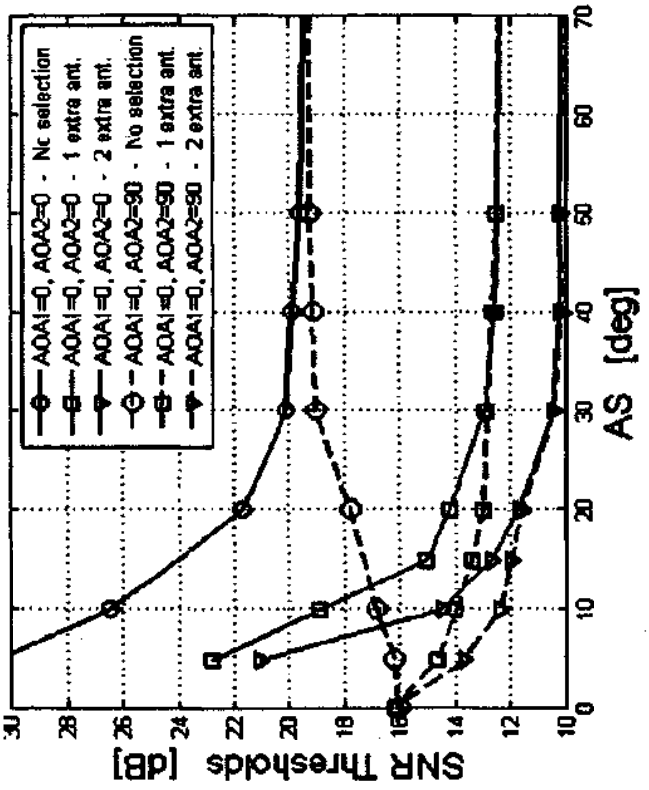

AOA1=0, AOA2=0 - No selection
AOA1=0, AOA2=0 - 1 extra ant.
AOA1=0, AOA2=0 - 2 extra ant.
AOA1=0, AOA2=90 - No selection
AOA1=0, AOA2=90 - 1 extra ant.
AOA1=0, AOA2=90 - 2 extra ant.
SNR Thresholds [dB]
AS [deg]
30
28
26
24
22
20
18
16
14
12
10
0
10
20
30
40
50
60
70

FIG. 37

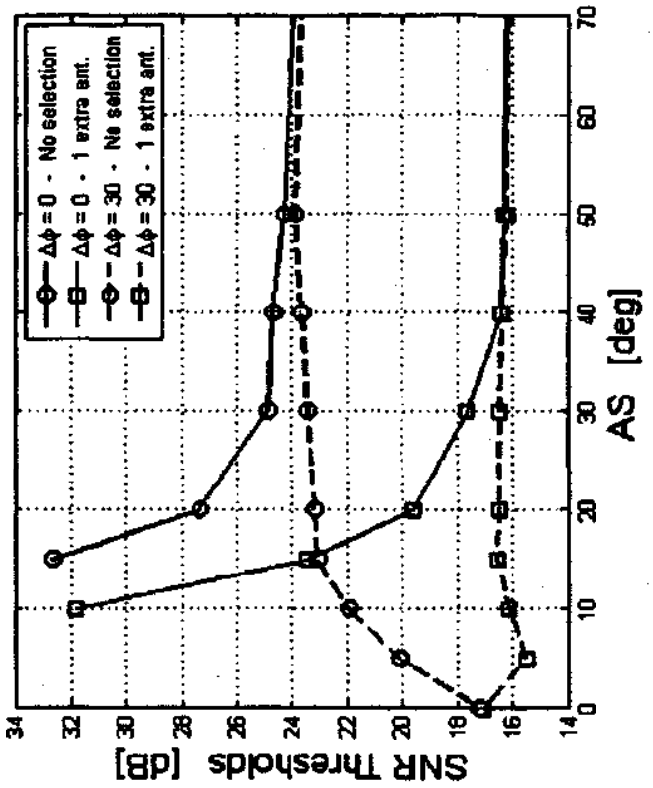

Δφ = 0 - No selection
Δφ = 0 - 1 extra ant.
Δφ = 30 - No selection
Δφ = 30 - 1 extra ant.
SNR Thresholds [dB]
AS [deg]
34
32
30
28
26
24
22
20
18
16
14
0
10
20
30
40
50
60
70

FIG. 38

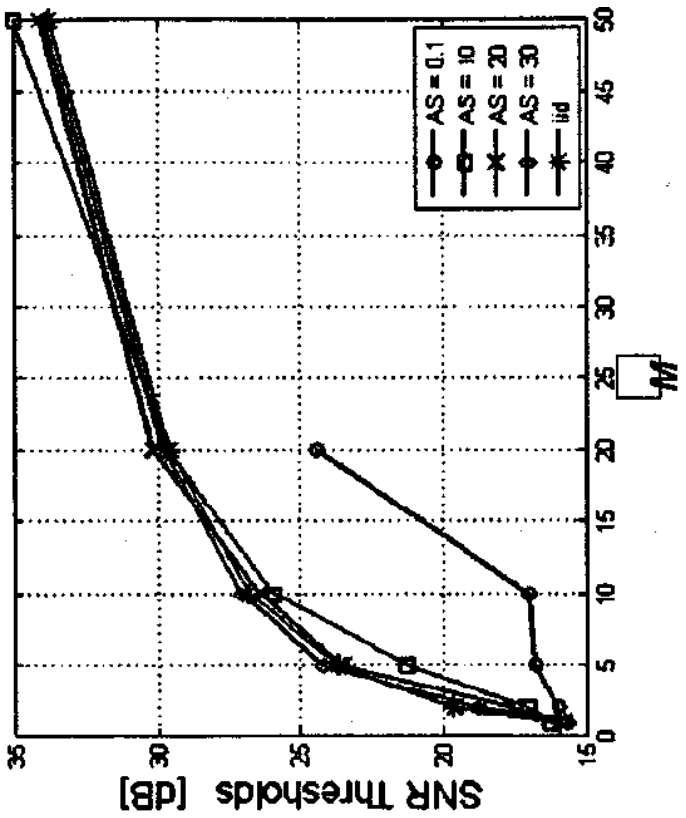


FIG. 39

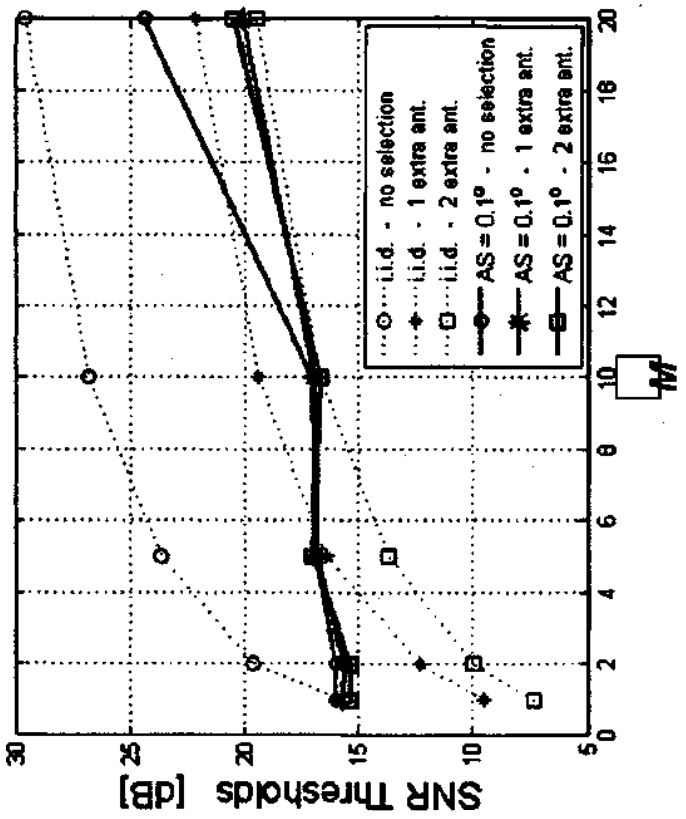

SNR Thresholds [dB]
M
i.i.d. - no selection
i.i.d. - 1 extra ant.
i.i.d. - 2 extra ant.
AS = 0.1° - no selection
AS = 0.1° - 1 extra ant.
AS = 0.1° - 2 extra ant.

FIG. 40

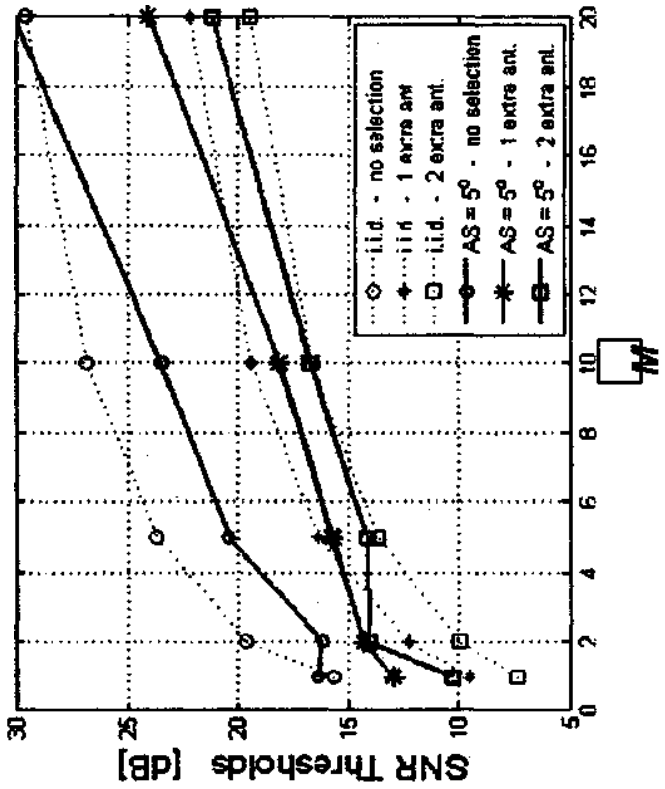

SNR Thresholds [dB]
M
i.i.d. - no selection
i.i.d. - 1 extra ant
i.i.d. - 2 extra ant.
AS = 5° - no selection
AS = 5° - 1 extra ant.
AS = 5° - 2 extra ant.

FIG. 41

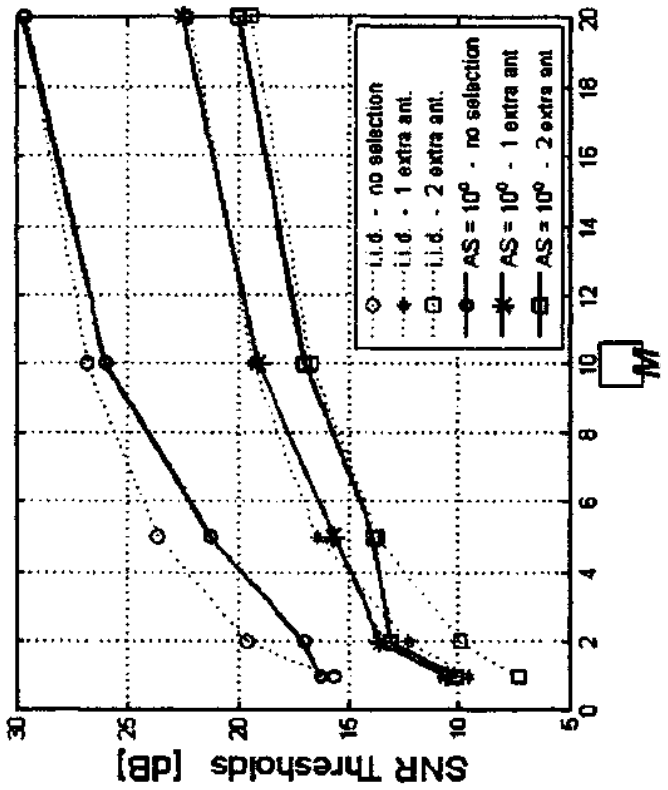

SNR Thresholds [dB]
M
i.i.d. - no selection
i.i.d. - 1 extra ant.
i.i.d. - 2 extra ant.
AS = 10^0 - no selection
AS = 10^0 - 1 extra ant
AS = 10^0 - 2 extra ant

FIG. 42

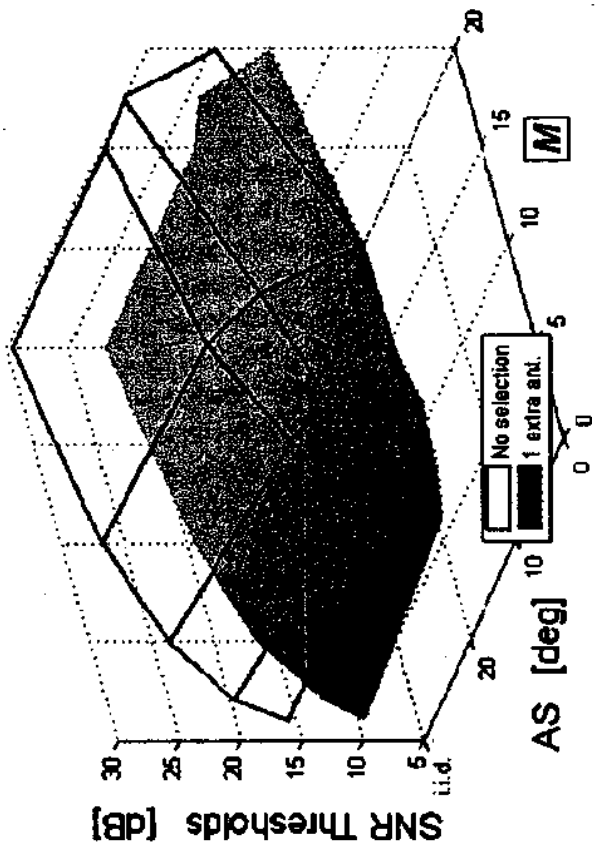

SNR Thresholds [dB]
30
25
20
15
10
5
i.i.d.
AS [deg]
5
10
15
20
M
No selection
1 extra ant.
0
0

FIG. 43

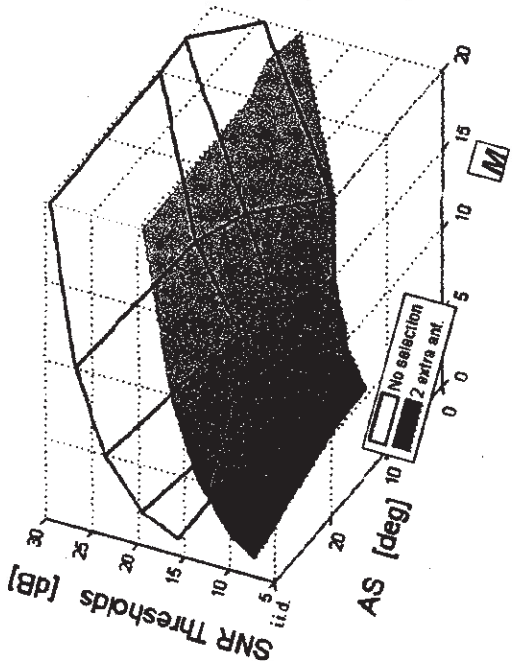

SNR Thresholds [dB]
30
25
20
15
10
5
i.i.d.
AS [deg]
10
20
M
0
5
10
15
20
No selection
2 extra ant.

FIG. 44

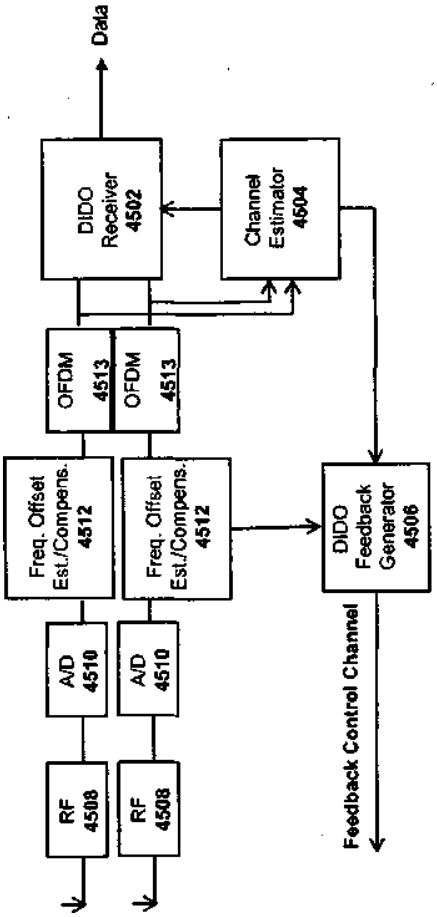

Data
DIDO Receiver 4502
Channel Estimator 4504
OFDM 4513
OFDM 4513
Freq. Offset Est./Compens. 4512
Freq. Offset Est./Compens. 4512
DIDO Feedback Generator 4506
A/D 4510
A/D 4510
RF 4508
RF 4508
Feedback Control Channel

FIG. 45

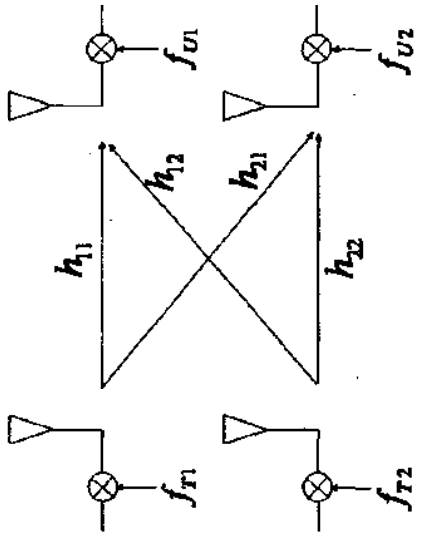


FIG. 46

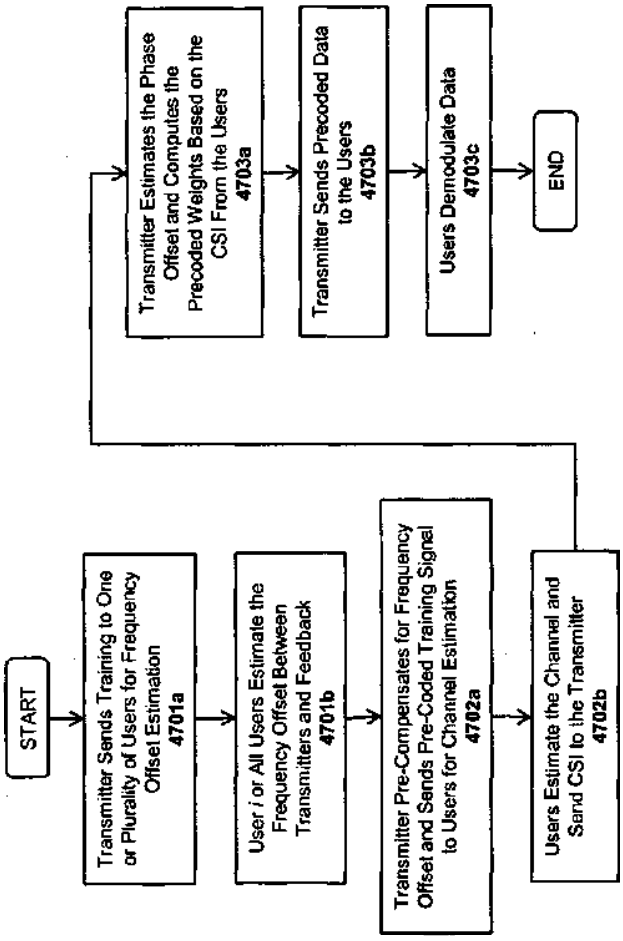

START
Transmitter Sends Training to One or Plurality of Users for Frequency Offset Estimation
4701a
User i or All Users Estimate the Frequency Offset Between Transmitters and Feedback
4701b
Transmitter Pre-Compensates for Frequency Offset and Sends Pre-Coded Training Signal to Users for Channel Estimation
4702a
Users Estimate the Channel and Send CSI to the Transmitter
4702b
Transmitter Estimates the Phase Offset and Computes the Precoded Weights Based on the CSI From the Users
4703a
Transmitter Sends Precoded Data to the Users
4703b
Users Demodulate Data
4703c
END

FIG. 47

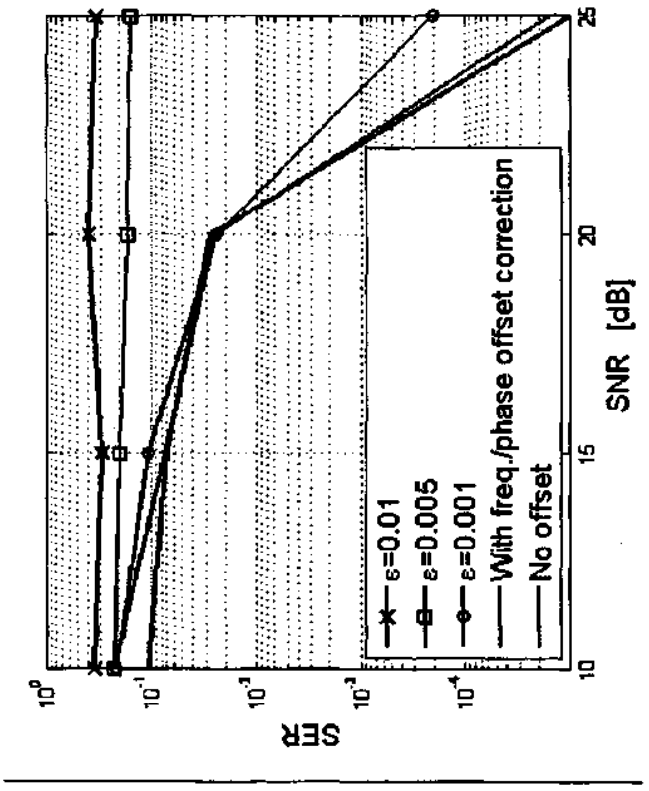

SER
SNR [dB]
ε=0.01
ε=0.005
ε=0.001
With freq./phase offset correction
No offset

FIG. 48

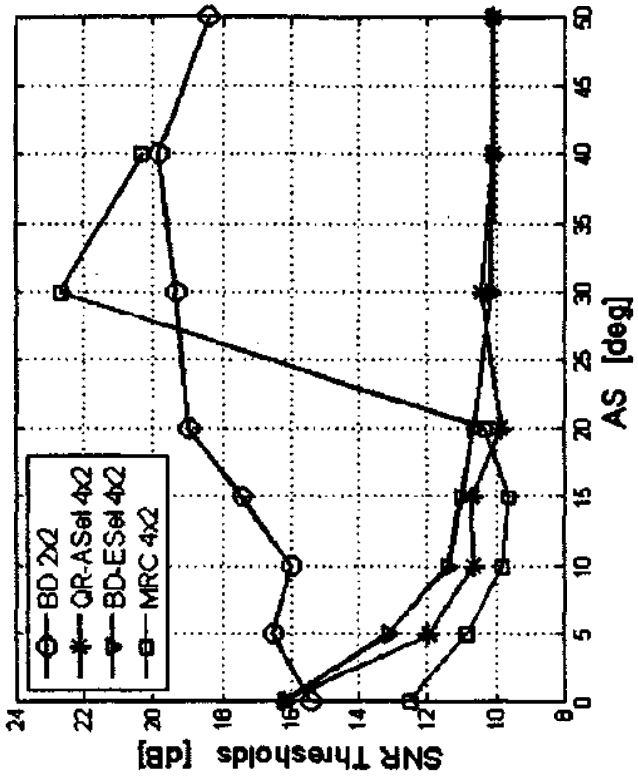

SNR Thresholds [dB]
AS [deg]
BD 2x2
QR-ASel 4x2
BD-ESel 4x2
MRC 4x2
24
22
20
18
16
14
12
10
8
0
5
10
15
20
25
30
35
40
45
50

FIG. 49

Method	# of Symbols	Run Time (1,000 iterations)	Max Correct. Offset
Old	14208	600s	?
New ($N_t/\lfloor M \rfloor = 1$)	288	60s	1500Hz
New ($N_t/\lfloor M \rfloor = 4$ length N_t)	1056	600s	1500Hz

FIG. 50

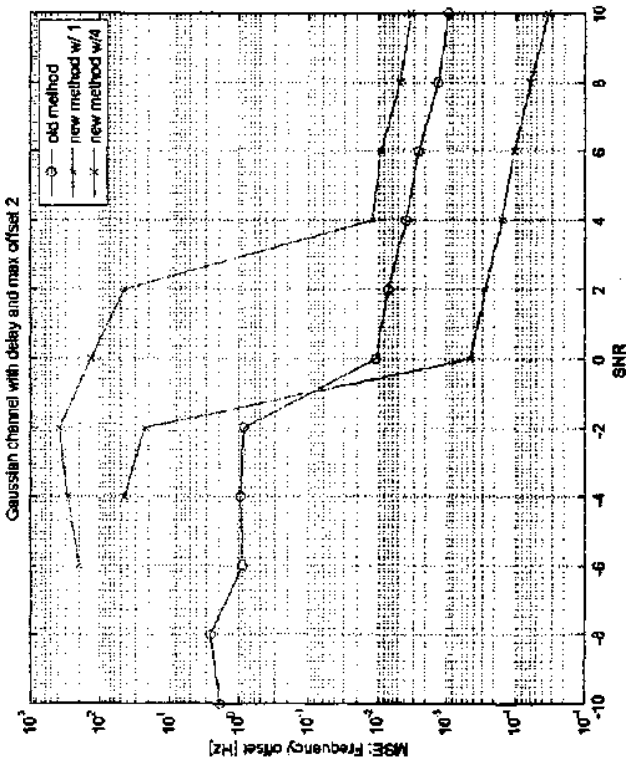

Gaussian channel with delay and max offset 2
old method
new method w/ 1
new method w/4
MSE: Frequency offset [Hz]
SNR

FIG. 51

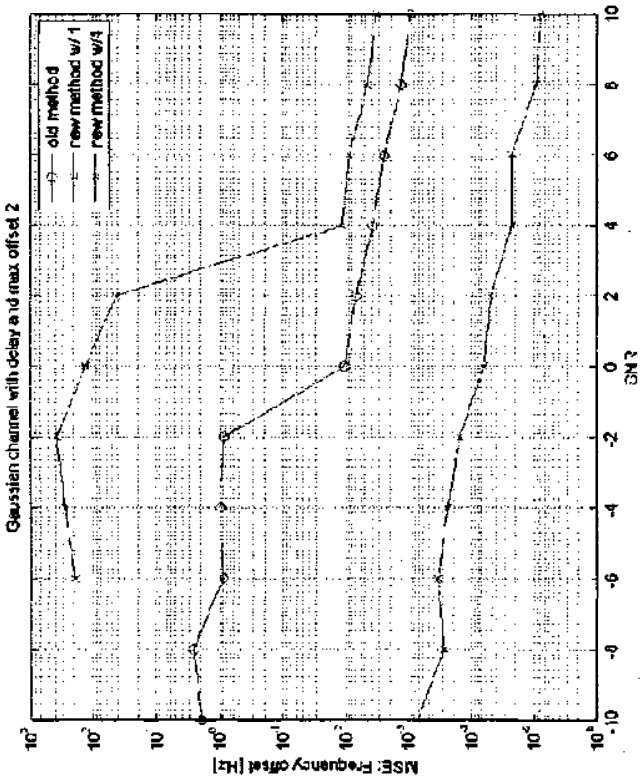
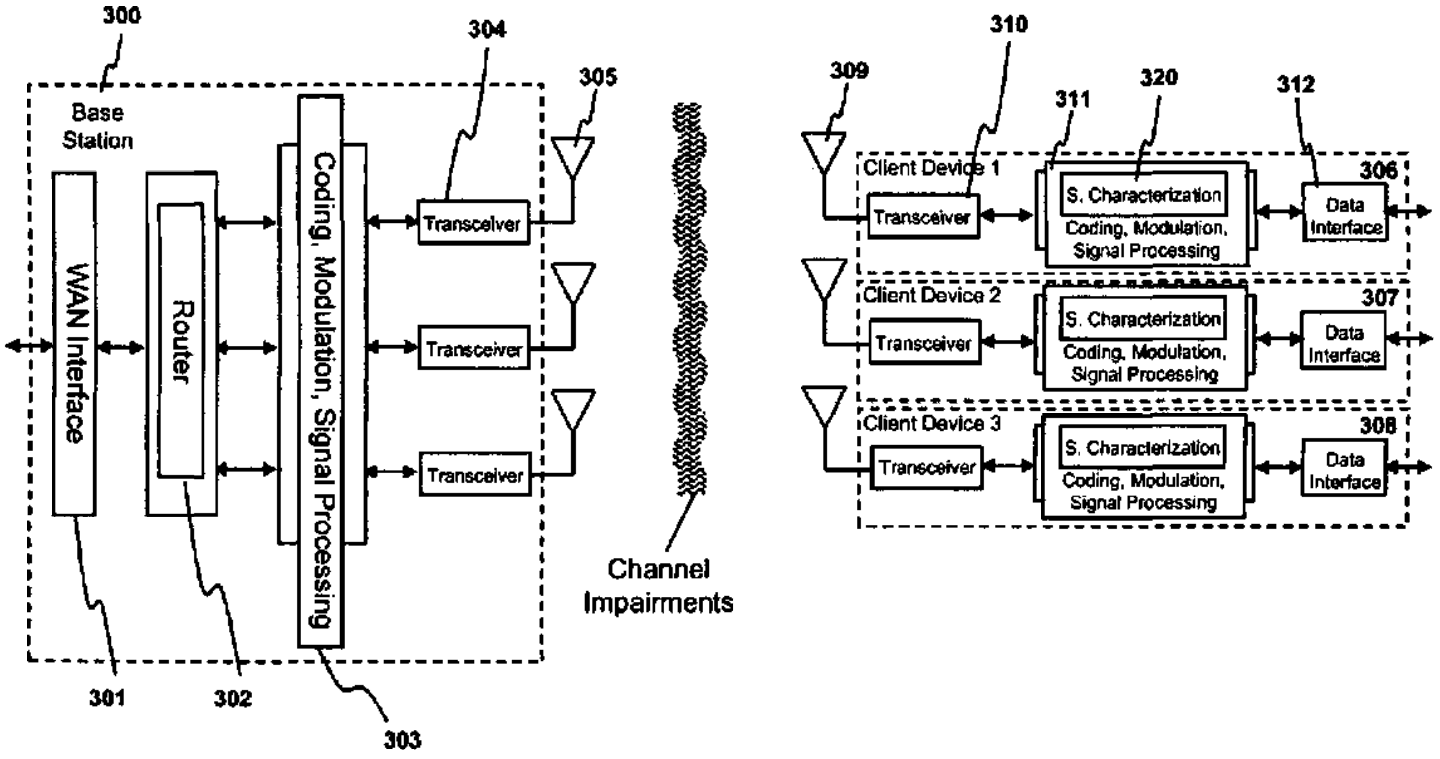

Gaussian channel with delay and max offset 2
old method
new method w/ 1
new method w/4
MSE: Frequency offset [Hz]
SNR

FIG. 52



3-Antenna Base Station with 3 Single-Antenna Client Devices

CN103501193B

Method for compensating for MU-MAS and dynamically adapting to MU-MAS

Abstract

A method for compensating for MU-MAS and dynamically adapting to MU-MAS is described. The method for compensating for the MU-MAS comprises: transmitting a training signal from each antenna of a base station to one or each of a plurality of wireless client devices, one or each of the client devices analyzing each training signal to generate frequency offset compensation data, and receiving the frequency offset compensation data at the base station; computing MU-MAS precoder weights based on the frequency offset compensation data to pre-cancel the frequency offset at the transmitter; transmitting the training signal and analyzing the training signal on a link between each antenna and one or each of the plurality of wireless client devices to obtain channel characterization data at the base station; and receiving the channel characterization data at the base station; computing a plurality of MU-MAS precoder weights based on the channel characterization data to pre-cancel frequency and phase offset and/or inter-user interference; precoding data using the MU-MAS precoder weights to generate precoded data signals; and transmitting the precoded data signals through the antenna to the each respective client device.

Classifications

H04B7/0452 Multi-user MIMO systems

View 8 more classifications

China

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Other languages

Chinese

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J·斯佩克

Worldwide applications

2008 CA CN KR KR CN CA KR CN 2011 RU RU 2013 JP 2014 JP 2015 JP 2016 RU JP

Application CN201310407419.4A events

2007-08-20

Priority to US11/894,394

2008-08-20

Application filed by 瑞登有限责任公司

2014-01-08

Publication of CN103501193A

2017-04-12

Application granted

2017-04-12

Publication of CN103501193B

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Description

Compensation MU-MAS communications, the method for dynamically adapting MU-MAS

The application be the applying date on 08 20th, 2008, it is Application No. 200880102933.4, entitled " distributed defeated The divisional application of the application for a patent for invention of the system and method for entering distributed output radio communication ".

Priority request

The application is the continuation application of the application NO.10/902,978 that July 30 in 2004 submits to.

Technical field

The present invention relates generally to field of wireless communications.Especially, the present invention relates to be used to use dividing for Space-time coding techniques The system and method that cloth is input into the radio communication of distributed output.

Background technology

The space-time code of signal of communication

Space multiplexing and space-time code are known newer development in wireless technology.Due to there is several antennas to be used in each Terminal, so a kind of space-time code of specific type is referred to as " multi-input multi output " (MIMO).By using multiple days Sending and receiving, multiple independent radio waves can simultaneously be transmitted line in identical frequency range.Following article is carried The general introduction of MIMO is supplied.

IEEE member David Gesbert, IEEE member Mansoor Shafi, IEEE member Da-shan Shiu, The IEEE JOURNAL ON SELECTED of the senior member Ayman Naguib of IEEE member Peter J.Smith and IEEE AREAS IN COMMUNICATIONS, VOL.21, NO.3, APRIL2003：“From theory to Practice ：An Overview of MIMO Space-Time Coded Wireless Systems”。

IEEE member David Gesbert, IEEE member Helmut Bolcskei, Dhananijay A.Gore and IEEE IEEE TRANSCTIONS ON COMMUNICATIONS, VOL.50, the NO.12 of member Arogyaswami J.Paulraj, DECEMBER2000：”Outdoor MIMO Wireless Channels：Models and Performance Prediction”。

Substantially, MIMO technology is based on the spatially distributed day for producing coordinate spaces data flow in common band The application of line.Radio wave is propagated by this way, such that it is able to separating and demodulating individual signals in receiver, even if they In identical transmitted in band, this can cause the communication channel of independent (namely efficiently separating) on multiple statistical significances. Therefore, compare with the standard wireless communication system for making great efforts to suppress multipath signal (that is, multiple delay time signals of same frequency, and shake There is modification in width and phase place), MIMO may rely on irrelevant or weak related multipath signal, realize in given frequency band compared with High throughput and the signal to noise ratio for improving.Example shows, under

conditions of power and signal to noise ratio (SNR) are suitable, MIMO technology reality Show higher handling capacity (throughput), and traditional non-mimo system can only realize relatively low handling capacity.High pass is public Department's (high pass is maximum wireless technology supplier) websitehttp://www.cdmatech.com/products/what_mimo_ delivers.jsp : This function is described on the page of subscript entitled " What MIMO Delivers " : “MIMO is the only multiple antenna technique that increases spectral capacity by delivering two or more times the peak data rate of a system per channel or Per MHz of spectrum.To be more specific, for wireless LAN or applications QUALCOMM's fourth generation MIMO technology delivers speeds Of315Mbps in36MHz of spectrum or8.8Mbps/MHz.Compare this to the peak capacity Of802.11a/g (even with beam-forming or diversity techniques) which delivers Only54Mbps in17MHz of spectrum or3.18Mbps/MHz ".

On generally, due to several reasons, mimo system limit less than the actual property of 10 antennas facing to each device (because Improvement in this network is less than 10 × throughput) :

1. physical limit : There must be enough intervals between mimo antenna on setter, so as to each receive The signal of statistical iteration.Although even if the improvement of MIMO handling capacities still can be seen in the antenna spacing of wavelength fraction, When antenna is more nearly, efficiency deteriorates rapidly, which results in relatively low MIMO throughput multiplication devices.

See, for example, below with reference to document :

[1] D.-S.Shiu, G.J.Foschini, M.J.Gans, and J.M.Kahn, " Fading correlation and its effect on the eapacity of multielement antenna systems,”IEEE Trans.Comm., vol.48, no.3, pp.502-513, Mar.2000.

[2] V.Pohl, V.Jungnickel, T.Haustein, and C.von Helmolt, " Antenna spacing In MIMO indoor ehannels, " Proc.IEEE Veh.Technol.Conf., vol.2, pp.749-753, May2002.

[3] M.Stoytchev, H.Safar, A.L.Moustakas, and S.Simon, " Compact antenna Arrays for MIMO applications, " Proc.IEEE Antennas and Prop.Symp., vol.3, pp.708- 711, July2001.

[4] A.Forenza and R.W.Heath Jr., " Impact of antenna geometry on MIMO Communication in indoor clustered channels, " Proc.IEEE Antennas and Prop.Symp., vol.2, pp.1700-1703, June2004.

Additionally, for little antenna spacing, coupling effect each other may reduce the performance of mimo system.

See, for example, below with reference to document：

[5] M.J.Fakhereddin and K.R.Dandekar, " Combined effect of polarization Diversity and mutual coupling on MIMO capacity, " Proc.IEEE Antennas and Prop.Symp., vol.2, pp.495-498, June2003.

[7] P.N.Fletcher, M.Dean, and A.R.Nix, " Mutual coupling in multi-element array antennas and its influence on MIMO channel capacity,"IEEE Electronics Letters, vol.39, pp.342-344, Feb.2003.

[8] V.Jungnickel, V.Pohl, and C.Von Helmolt, " Capaeity of MIMO systems With closely spaced antennas, " IEEE Comm.Lett., vol.7, pp.361-363, Aug.2003.

[10] J.W.Wallace and M.A.Jensen, " Termination-dependent diversity performance of coupled antennas：Network theory analysis,"IEEE Trans.Antennas Propagat., vol.52, pp.98-105, Jan.2004.

[13] C.Waldschmidt, S.Schulteis, and W.Wiesbeck, " Complete RF system Model for analysis of compact MIMO arrays, " IEEE Trans.on Veh.Technol., vol.53, Pp.579-586, May2004.

[14]M.L.Morris and M.A.Jensen,"Network model for MIMO systems with Coupled antennas and noisy amplifiers, " IEEE Trans.Antennas Propagat., vol.53, Pp.545-552, Jan.2005.

And, when antenna it is crowded to together when, antenna generally has to be made less, and this can also affect antenna to imitate Rate.

See, for example, below with reference to document：

[15] H.A.Wheeler, " Small antennas, " IEEE Trans.Antennas Propagat., Vol.AP-23, n.4, pp.462-469, July1975.

[16] J.S.McLean, " Are-examination of the fundamental limits on the Radiation Q of electrically small antennas, " IEEE Trans.Antennas Propagat., Vol.44, n.5, pp.672-676, May1996.

Finally, with lower frequency and longer wavelength, the physical size of MIMO devices just becomes to be difficult to process.One pole The example at end is that, in HF wave bands, here MIMO devices antenna must 10 meters separated from each other or bigger distance.

2. noise is limited.The receiver of each MIMO/transmitter subsystem produces the noise of certain level.When more and more This subsystem when closing on placement mutually, ambient noise will rise.Meanwhile, when needs are recognized from multiple-antenna MIMO system When going out more unlike signals, it is desirable to lower ambient noise.

3. cost and power limit.Although cost and power consumption are not focuses in some MIMO applications, typical wireless In product, when developing a kind of successful product, cost and power consumption are all it is critical that restraining factors.For each MIMO days Line, needs detached RF subsystems, including detached analog-to-digital (A/D) sum-mould (D/A) converter.Unlike being come with Moore's Law Many aspects (observation of the experience aspect done by cofounder Gordon mole of Intel of the digital display circuit of measurement scale As a result, the transistor size on the integrated circuit of microdevice will quadruple per every about 24 months ; Source : http:// Www.intel.com/technology/mooreslaw/), generally there is so intensive analog subsystem certain physics to tie Structure size and power requirement, its size is proportional to cost and power linear.Therefore, compare with single antenna devices, multiple antennas MIMO devices will become extremely expensive and with surprising energy consumption.

Used as result above, most of mimo systems expected from today are in the grade of 2 to 4 antennas, to cause to gulp down The rising of 2 to 4 times of the amount of telling and the rising of some SNR (signal to noise ratio) caused due to the diversity benefit of multiaerial system. Anticipate 10 antennas mimo system (especially because shorter wavelength and nearer

antenna spacing in higher Microwave Frequency In rate), but it is very unactual more than 10 antennas in addition to special for some and insensitive to cost application 's.

Virtual antenna array

A kind of special applications of the technology of MIMO type are virtual antenna arrays.Euroscience technical field research cooperative groups This system, EURO, Barcelona, Spain, the day 15-17 of in January, 2003 are suggested in the research file for knitting proposition : Center For Telecommunication Research, King's College London, UK : "A step towards MIMO : Virtual Antenna Arrays ", Mischa Dohler & Hamid Aghvami.

As described in file, virtual antenna array is cooperation wireless device system (such as cell phone), and it is detached Communicate in communication channel (if when they mutually close on enough), rather than in their main communication channels with they Base station communication so that the work of collaborative ground is (for example, if they are the GSM cell phones in UHF waveband, then this can be with It is the wireless wave band of industrial scientific medical (ISM) of 5GHz).By in mutual relaying scope (except in base station range) Forwarding information between several devices, just look like they be that physically the work of a device with multiple antennas is the same so that Single antenna devices potentially realize the throughput hoisting as MIMO.

However, in fact, such system it is extremely difficult realization and use is limited.First of all, it is necessary to keep each device present A minimum of two different communication paths are lifted with realizing throughput, and the availability of its second repeated link is often uncertain 's.It is additionally, since them at least to there is the second communication subsystem and have bigger calculating demand, therefore the device is more expensive, Physical size is bigger, and consumes more power.Additionally, potentially through multiple communication links, the system is depended on very The complicated systematic live collaboration of institute.Finally, due to simultaneous channel usage increases (for example, using MIMO technology Simultaneous phone call transmission), for the computation burden of each device is also increased (generally with the linear increasing of channel usage Plus and exponentially increases), this is very unpractiaca to the portable unit with strict power and size limitation.

The content of the invention

Describe a kind of for the frequency in the multiaerial system (MAS) that (" MU-MAS ") is sent with multi-user (MU) The system and method compensated with phase offset.For example, included according to the method for one embodiment of the present invention：Will be from base One or each wireless client device that the training signal of each antenna of standing is sent into multiple wireless client devices, client's dress One in putting or each customer set up analyze each training signal to generate frequency offset compensation data, and receive in base station Frequency offset compensation data；MU-MAS precoders weight is calculated based on the frequency offset compensation data and is launched with pre-elimination Frequency shift (FS) at machine；Precoding is carried out to training signal using the MU-MAS precoders weight, to generate base station is directed to The precoding training signal of each antenna；Training signal after the precoding of each antenna of the base station is sent to into institute Each wireless client device in multiple wireless client devices is stated, each customer set up analyzes each training signal to generate letter Road characteristic, and receive the channel characteristics data in the base station；Multiple MU- are calculated based on the channel characteristics data MAS precoding weights, the MU-MAS precoder weights be calculated to for it is pre- eliminate frequency and phase offset and/or user it Between interference；Precoding is carried out to data using MU-MAS precoders weight, to generate for the pre- of each antenna in base station Data-signal after coding；And the pre-coded data signal after the precoding is sent to it by each antenna of base station The client device of each.

Description of the drawings

With reference to accompanying drawing, below in detail description can obtain the present invention is better understood from, wherein：

Fig. 1 shows the mimo system of prior art.

Fig. 2 shows the N antenna base stations communicated with multiple single antenna customer set ups.

Fig. 3 shows the base station of three antennas communicated with three single antenna customer set ups.

Fig. 4 shows the training signal technology used in one embodiment of the present of invention.

Fig. 5 shows the channel characteristics data that base station is transferred to from customer set up according to an embodiment of the invention.

Fig. 6 shows distributed output (" the MIDO ") downlink transfer of multiple input according to an embodiment of the invention.

Fig. 7 shows multi-input multi output (" MIMO ") uplink according to an embodiment of the invention.

Fig. 8 is shown and according to an embodiment of the invention is circulated to distribute the base of handling capacity by different clients group Stand.

Fig. 9 shows according to an embodiment of the invention based on the custom partitioning for closing on.

Figure 10 shows the embodiments of the invention used in NVIS systems.

Figure 11 shows the embodiment of the DIDO emitters with I/Q compensation function units.

Figure 12 shows the DIDO receivers with I/Q compensation function units.

Figure 13 shows a kind of embodiment of the DIDO-OFDM systems with I/Q compensation.

Figure 14 shows of DIDO2 × 2 performance (performance) in the case where having and not there is I/Q to compensate Plant embodiment.

Figure 15 shows a kind of embodiment of DIDO2 × 2 performance in the case where having and not having I/Q compensation.

Figure 16 is shown in the case where having and not there is I/Q to compensate for SER (the symbol mistakes of different Q AM planisphere Code check) a kind of embodiment.

Figure 17 shows DIDO2 × 2 performance in the case where different user devices position has and not there is I/Q to compensate A kind of embodiment.

Figure 18 shows have in preferable (i.i.d. (independent and same distribution)) channel and do not have a case that I/Q is compensated A kind of embodiment of lower SER.

Figure 19 shows a kind of embodiment of the transmitter architecture of self adaptation DIDO system.

Figure 20 shows a kind of embodiment of the receiver architecture of self adaptation DIDO system.

Figure 21 shows a kind of embodiment of the method for self adaptation DIDO-OFDM.

Figure 22 shows a kind of embodiment of the antenna arrangement for DIDO measurements.

Figure 23 shows the embodiment of the array configuration for different stage (order) DIDO systems.

Figure 24 shows the performance of different stage DIDO systems.

Figure 25 shows a kind of embodiment of the aerial array for DIDO measurements.

Figure 26 shows one kind of DIDO2 × 2 performance of 4-QAM and 1/2FEC rates and the functional relation of location of user equipment Embodiment.

Figure 27 shows a kind of embodiment of the antenna arrangement for DIDO measurements.

Figure 28 show DIDO8 × 8 in one embodiment how producing ratio have the DIDO2 of low TX power demands × 2 bigger SE.

Figure 29 shows a kind of embodiment of DIDO2 × 2 performance in the case of with day line options.

Figure 30 shows average BER (BER) performance of the different DIDO pre-coding schemes in i.i.d. channels.

Figure 31 shows the function in the snr gain of ASel and i.i.d. channels between the quantity of extra transmitting antenna Relation.

Figure 32 shows that SNR threshold values are diagonal with for block in the case of having 1 and 2 exterior antenna in i.i.d. channels Change the functional relation between (BD) and the number of users (M) of ASel.

Figure 33 is shown for the BER positioned at equal angular direction and two users with different angle spreads (AS) value SNR average with every user.

Figure 34 shows the result similar with Figure 33, but there is higher angle to be spaced between user.

Figure 35 depicts the different value of the mean angle of arrival (AOA) for user, and the function between AS and SNR threshold values is closed System.

Figure 36 shows the SNR threshold values of the exemplary cases for 5 users.

Situations of the Figure 37 for 2 users, there is provided in the case of with 1 and 2 additional antenna, SNR threshold values BD with The comparison of ASel.

Figure 38 shows the result similar with Figure 37, but for the situation of 5 users.

Figure 39 shows the SNR threshold values for the BD schemes with difference AS values.

Figure 40 is shown for BD and ASel with 1 and 2 additional antenna, with AS=0.1 ° of space correlation letter SNR threshold values in road.

Figure 41 shows the calculating of the SNR threshold values of two other channel conditions for AS=5 °.

Figure 42 shows the calculating of the SNR threshold values of two other channel conditions for AS=10 °.

Figure 43-Figure 44 is respectively illustrated in the case of 1 and 2 additional antenna, SNR threshold values and number of users (M) and BD And the functional relation between the angle spread (AS) of ASel schemes.

Figure 45 shows the receiver equipped with frequency offset estimator/compensator ;

Figure 46 shows DIDO2 × 2 system model according to one embodiment of the present invention.

Figure 47 shows the method according to one embodiment of the present invention.

Figure 48 show with and have frequency shift (FS) in the case of, the SER results of DIDO2 × 2 system.

Figure 49 compares the SNR threshold performances of different DIDO schemes.

Figure 50 compares the amount of overhead needed for distinct methods embodiment.

Figure 51 is shown in f_{max}=2Hz small frequency skew and without integer offset correction in the case of emulation.

Figure 52 shows the result when integer offset estimator is closed.

Specific embodiment

In the following description, for purposes of explanation, it is multiple special to illustrate thoroughly is understood to the present invention in order to provide Details.It may be evident, however, that for one of ordinary skilled in the art, even if without some specific details, still can be with Realize the present invention.Additionally, known construction and device is shown as block diagram format, to avoid obscuring ultimata of the present invention Change.

Fig. 1 shows the prior art mimo system with transmitting antenna 104 and reception antenna 105.Such system Throughput can realize 3 times of the throughput generally realized in available channel.There are various different methods this to realize The details of mimo system, it had description in the published document with regard to the theme, and explanation below will be described as one Method.

Before data are transmitted in the mimo system of Fig. 1, channel is by " characterization ".This is by starting " training letter Number " be transferred to each receiver 105 to realize from each transmitting antenna 104.Training signal has coding and mod subsystem 102 Generate, and analog signal is converted into by D/A converter (not shown), then changed from baseband

signal by each transmitter 103 For RF signals.The reception antenna 105 for being each coupled to its RF receiver 106 receives each training signal and is converted into base band Signal.Baseband signal is converted to data signal by D/A converter (not shown), and then signal processing subsystem 107 is characterized The training signal.The feature of each signal can include several factors, and for example, it includes, relative to the reference inside receiver The phase place and amplitude of signal, absolute reference signal, relative reference signal, characteristic noise or other factors.The feature of each signal It is normally defined the vector that the phase place and amplitude variations of the several aspects of signal are showed when signal is transmitted by channel.For example, exist In the modulated signal of quadrature amplitude modulation (" QAM "), the feature is probably the phase place and amplitude of several multipath images of signal The vector of skew.Another example is that in the signal of OFDM (" OFDM ") modulation, it is probably OFDM frequency spectrums In several or all single component signals (sub-signal) phase place and amplitude excursion vector.

Signal processing subsystem 107 deposits the channel characteristics received by each reception antenna 105 and corresponding receiver 106 Storage is got up.All of three transmitting antennas 104 are completed after their training signal transmission, and signal processing subsystem 107 is by Stored three, for the channel characteristics of each in three reception antennas 105, this results in 3 × 3 matrix 108, its table It is shown as channel characteristics matrix " H ".Each single matrix element $H_{I, j}$It is transmission antenna 104i that reception antenna 105j is received The channel characteristics of training signal transmission.

At this point, signal processing subsystem 107 inverts matrix H 108 to produce H^{-1}, and wait from transmitting antenna The transmission of 104 real data.Note, various existing MIMO technologies described in available document can be used to guarantee H-matrix 108 is reversible.

In force, the content (payload) of the data to be transmitted is sent to data input subsystem 100.Then it is being sent to Before coding and mod subsystem 102, its allocated device (splitter) 101 is divided into three parts.For example, if content is The ASCII bits of " abcdef ", it can just be allocated device and be divided into three sub- contents " ad ", " be " and "

cf ".Then, each Sub- content is individually sent to coding and mod subsystem 102.

By using the coded system of the statistical independence and error correcting capability for being adapted to each signal, individually in every height Appearance is encoded.These include, and are not limited only to, Reed-Solomon codings, Viterbi coding (Viterbi coding) (Turbo Codes) is encoded with strengthening.Finally, use to the suitable modulator approach of channel in the sub- contents after these three codings Each be modulated.Exemplary modulator approach is differential phase keying (DPSK) modulation (" DPSK "), 64-QAM modulation and OFDM. It should be noted there that the diversity gain that MIMO is provided allows the modulation constellation of higher plate number, the modulation constellation to exist Using being also feasible in SISO (single-input single-output) system of same channel.Then, each is encoded and the letter after modulating Number transferred out by the antenna 104 of its own, the transmission follow D/A converting unit (not shown)s D/A conversions and After the RF of each transmitter 103 is generated.

Assume that there are enough space diversitys to be present in sends and receives between antenna, each reception antenna 105 will be from antenna 104 various combinations for receiving three transmission signals.Each signal is received and is converted them to base by each RF receiver 106 Band signal, then A/D converter (not shown) signal is digitized again.If y_nIt is to be connect by n-th reception antenna 105 The signal for receiving, x_nIt is the signal sent by n-th transmitting antenna 104, N is noise, then this just can be described with following equalities.

$y_1=x_1H_{11}+x_2H_{12}+x_3H_{13}+N$

$y_2=x_1H_{12}+x_2H_{22}+x_3H_{23}+N$

$y_3=x_1H_{13}+x_2H_{32}+x_3H_{33}+N$

Assume that this is the system of three equation with three unknown quantitys, then here it is signal processing subsystem 107 derive x_1、x_2And x_3Linear algebra problem (assuming N in sufficiently low level, it is allowed to decode to signal) :

$x_1=y_1H^{-1}{}_{11}+y_2H^{-1}{}_{12}+y_3H^{-1}{}_{13}$

$x_2=y_1H^{-1}{}_{21}+y_2H^{-1}{}_{22}+y_3H^{-1}{}_{23}$

$x_3=y_1H^{-1}{}_{31}+y_2H^{-1}{}_{32}+y_3H^{-1}{}_{33}$

Once derive the signal x of three transmission$_n$, they are just demodulated by signal processing subsystem 107, are decoded and error correction, To recover three bit streams for being separated by distributor 101 originally.These bit streams merge in combiner unit 108, and from Single data stream is output as in data output 109.Assume that system robust can overcome noise induced damage, then data output 109 is produced Raw bit stream is by as the bit stream being incorporated in data input 100.

Although described prior art systems are generally effectively until four antennas, perhaps until the antenna of as many as 10, Due to due to described in the background parts in the disclosure, it becomes very not during with a large amount of antennas (such as 25,100 or 1000) It is actual.

Generally, such prior art systems are two-way, and return path is realized in an identical manner, but anti-mistake Come, all have in every side of communication channel and send and receive subsystem.

Fig. 2 shows one embodiment of the present of invention, and wherein, base station (BS) 200 is configured with wide area network (WAN) interface (for example by T1 or other connect at a high speed) 201 and be provided with a number of (N number of) antenna 202.We use for the time being art Language " base station " carries out any wireless site of radio communication to refer to the one of fixed position group of client.The example of base station can be nothing Access point in line LAN (WLAN), or WAN antenna or aerial array.There is some customer set ups 203-207, each has single Antenna, base station 200 is wirelessly serviced them.Although for the purpose of this example, being very easy to expect position In the base station of office environment, in this environment, it is the user's set 203- of the personal computer for being equipped with wireless network 207 provide services, but this structure will apply to substantial amounts of applicable cases, indoor and outdoors, and here base station is served wirelessly Client.For example, the base station may be located on cellular tower, or in television broadcast towers.In one embodiment, base Stand and 200 be placed in ground, for (such as the frequency of 24MHz) up transmission of HF frequencies, by signal from ionospheric reflection Return, as on April 20th, 2004 proposes, Serial No. No.10/817,731, entitled SYSTEM

AND METHOD FOR ENHANCING NEAR VERTICAL INCIDENTCE SKYWAVE(“NVIS”)COMMUNICATION USING SPACE- While TIME CODING as pending application description, it is propped up the agent of dispensing the application, herein as ginseng Examine.

Some details being associated with base station 200 and illustrated customer set up just for the sake of illustration purpose, and not Necessary to be cardinal principle of the invention.For example, the base station can be connected to multiple different types via wan interface 201 Wide area network, it includes private wide area network, and for example those are used for the wide area network that digital video sends.Similarly, customer set up can To be any kind of wireless data processing and/or communicator, it includes, and is not limited solely to, cell phone, individual number Word assistant (" PDA "), receiver and wireless camera.

In one embodiment, the n antenna 202 of base station is spatially separated, so as to each sends and receives The related signal of non-space, just look like the base station be prior art MIMO transceiver it is the same.As retouched in the introduction State, antenna has been made with the experiment that λ/6 (i.e. 1/6 wavelength) is spaced apart, and it successfully realizes the handling capacity from MIMO Lifted, but in general, these antenna for base station are more spaced further apart placing, and the performance of system is better, and λ/2 are gratifying Minimum range.Certainly, cardinal principle of the invention is not limited to any specific separation between antenna.

Note, its antenna can be positioned over far distance by single base station 200 well.For example, in HF frequency spectrums, antenna Can have 10 meters or farther (for example during the NVIS, being generally noted above is realized).If using 100 such antennas, the base station Aerial array can just occupy several square kilometres of area.

In addition to space diversity technology, in order to improve the effective throughput of system, one embodiment of the present of invention will be believed Number polarization.It is a kind of known technology that channel capacity is improved by polarizing, and it has been used very by satellite television providers For many years.Using Polarization technique, multiple (such as three) base stations or user antenna can be made with each other closely, and be remained Non-space correlation.Although traditional RF

systems mostly just benefit from two dimension (such as x and y) diversity of polarization, retouch here The structure stated can further benefit from three-dimensional (x, the y and z) diversity of polarization.

In addition to space and polarity diversity, one embodiment of the present invention is using almost orthogonal antenna pattern (pattern), improving link performance via directional diagram diversity.Directional diagram diversity can improve the capacity and error code of mimo system Rate performance, and it can be found in following article compared to the advantage of other diversity antenna technologies :

[17] L.Dong, H.Ling, and R.W.Heath Jr., " Multiple-input multiple-output

wireless communication systems using antenna pattern diversity," Proc.IEEE Glob.Telecom.Conf., vol.1, pp.997-1001, Nov.2002.

[18] R.Vaughan, " Switched parasitic elements for antenna diversity, " IEEE

Trans.Antennas Propagat., vol.47, pp.399-405, Feb.1999.

[19] P.Mattheijssen, M.H.A.J.Herben, G.Dolmans, and L.Leyten, " Antenna-Pattern diversity versus space diversity for use at handhelds, " IEEE Trans.on Veh.Technol., vol.53, pp.1035-1042, July2004.

[20] C.B.Dietrich Jr, K.Dietze, J.R.Nealy, and W.L.Stutzman, " Spatial, Polarization, and pattern diversity for wireless handheld terminals, " Proc.IEEE Antennas and Prop.Symp., vol.49, pp.1271-1281, Sep.2001.

[21] A.Forenza and R.w.Heath, Jr., " Benefit of Pattern Diversity Via2-Element Array of Circular Patch Antennas in Indoor Clustered MIMO Channels ", IEEE Trans.on Communications, vol.54, no.5, pp.943-954, May2006.

By using directional diagram diversity, but cause multiple base stations or user antenna each other closely, although and such as This also will not spatially be associated.

Fig. 3 provides the additional detail of one embodiment of the base station 200 shown in Fig. 2 and customer set up 203-207.For The purpose for simplifying, the base station 300 is only displayed as three antennas 305 and three customer set up 306-308.However, it is necessary to note Meaning, embodiments of the invention described herein can be with the antenna 305

of virtually limitless quantity (that is, only by available sky Between and noise limiting) and customer set up 306-308 realizing.

Fig. 3 is similar with the prior art MIMO structure shown in Fig. 1, wherein, both have three in every one end of communication channel Antenna.Significant difference is, in the mimo system of prior art, three antennas 105 on the right side of Fig. 1 be from each other it is fixed away from From (for example, being integrated in single device), the signal received from each antenna 105 is together in signal processing subsystem 107 Processed.By contrast, in figure 3, three antennas 309 on the right side of figure are each coupled to different customer set up 306- On 308, each described customer set up can be distributed in the range of base station 305 Anywhere.So, each customer set up The signal for receiving can be in its coding, modulation, signal processing subsystem 311 independently of other two signals for receiving Processed.Therefore, compared with " MIMO " system of the multiple output (i.e. antenna 104) of multiple input (i.e. antenna 105), Fig. 3 Distributed output (the i.e. antenna 305) system of multiple input (i.e. antenna 305) is shown, below refers to " MIDO " system.

Note, the application uses the term usage different from application before, it is used to conform better to academia and industry Example.In entitled " the SYSTEM AND METHOD FOR ENHANCING NEAR that 20 days cited before April in 2004 submits to VERTICAL INCIDENCE SKYWAVE (" NVIS ") COMMUNICATION USING SPACE-TIME CODING's " is total to (the application is this to the application NO.10/902,978 submitted to copending application NO.10/817,731 and on July 30th, 2004 The continuation application of application) in, the meaning of " input " and " output " (in the environment of SIMO, MISO, DIMO and MIDO) with should Term expressing the meaning in this application is contrary.In application before, " input " refers to input to reception antenna (for example, in Fig. 3 Antenna 309) wireless signal, and " output " refer to transmitting antenna (for example, antenna 305) export wireless signal.In academia In wireless industry, the antisense of " input " and " output " is usually used, wherein " input " refers to the wireless signal being input into channel (that is, the wireless signal for sending from antenna 305), and " output " refers to that (that is, antenna 309 is received from the wireless signal of channel output Wireless signal).The application adopts this term usage, and the usage is contrary with the usage

in the application that this section is cited before.Cause This, below depicts the term usage equivalent form of value between several applications :

MIDO structures shown in Fig. 3 are real in SISO systems similar to MIMO for the transmitting antenna of given quantity is realized Existing capacity boost.However, it is to realize multiple base station days that a difference of the specific MIDO embodiments shown in MIMO and Fig. 3 is The capacity boost that line is provided, each MIDO customer set up 306-308 is only required to single reception antenna, and for MIMO, each Customer set up at least requires the as many reception antenna of capacity multiple realized with hope.Hypothesis generally has the restriction of an implementation, It is limited how many antennas (as explained in the introduction) can be placed in customer set up, and this just limits mimo system on typical case System is between 4 to 10 antennas (4 times to 10 times of capacity).Because base station 300 generally from fixation and fills dynamic position Many customer set ups are served, is expanded to and is exceeded well over 10 antennas, and with appropriately distance separate antenna realizing space Diversity is very actual.As described, each antenna arrangement has transceiver 304 and part coding, modulation and Signal Processing Element 303 disposal ability.It should be noted that in this embodiment, no matter base station 300 extends how many, each customer set up 306- 308 will be only required to an antenna 309, therefore the cost for single user customer set up 306-308 will be very low, and base station 300 cost can be shared in the user of large cardinal.

In Fig. 4 to Fig. 6, it is shown that how to complete from base station 300 to the example of the MIDO of customer set up 306-308 transmission Son.

In one embodiment of the invention, before MIDO transmission starts, channel is characterized.For mimo system, often Individual antenna 405 is transmitted one by one to training signal.Fig. 4 only show the transmission of first training signal, but right For three antennas 405, three separate transmission are had.Each training signal is by encoding, modulate and signal processing subsystem 403 generate, and by D/A converter analog signal is converted into, and are sent by each RF transceiver 404 as RF signals. Available a variety of codings, modulation and signal processing technology include, and are not limited to those technology described above

(examples Such as, Reed Solomon, Viterbi coding (Viterbi Coding) ; QAM, DPSK, QPSK modulation etc.).

Each customer set up 406-408 is received training signal and is believed the training by transceiver 410 by its antenna 409 Number it is converted into baseband signal.A/D converter (not shown) is encoded in the signal, modulate and signal processing subsystem 411 at Data signal is converted thereof into where reason.Then the feature of the identification of signal characteristic logical block 320 gained signal (for example, is known Not above-mentioned phase place and amplitude distortion) and this feature is stored in memory.This characteristic processing process is similar to existing skill The processing procedure of the mimo system of art a, significant difference is that each customer set up only calculates one antenna, rather than The characteristic vector of n antenna.For example, as it is known that the training signal of pattern is by the coding of customer set up 406, modulation and signal Reason subsystem 420 initialize (produce when by send information in receive it, or by other initialization process).The same day When line 405 sends in a known pattern the training signal, coding, modulation and signal processing subsystem 420 are using correlation method Find most strong training signal reception pattern, it saves phase place and amplitude excursion, then its by the pattern from receiving Signal in the middle of cut.Next, it finds the last the second reception pattern related to the training signal, phase place and amplitude is inclined Shifting is saved, and then it cuts the second strong mode from the signal for receiving.The process is carried out always, until preserving The phase place of certain fixed qty and amplitude excursion (for example, 8) or detectable training signal pattern drop to given background Under noise.The vector of the phase/amplitude skew becomes the element H of vector 413_{11}.At the same time, customer set up 407 and 408 Coding, modulation and signal processing subsystem perform same process, produce their vector element H_{21}And H_{31}。

The memory of channel characteristics storage can be nonvolatile memory, such as flash memory, or hard disk, and/or volatibility Memory, such as random access memory (for example, SDRAM, RDAM).Additionally, different user's sets can simultaneously using not The memory of same type is storing characteristic information (such as PDA can be and use flash memory, and notebook computer can be and use hard disk). On various

customer set ups or base station, ultimata of the present invention is not limited to any certain types of storing mechanism.

As described above, according to the scheme for being used, because each customer set up 406-408 only has an antenna, each is only Only store 1 × 3 row 413-415 of H-matrix.Fig. 4 shows the stage after the transmission of the first training signal, here, 1 × 3 row 413- 415 first row stores the channel characteristic information of first antenna of three antenna for base station 405.Remaining two row is stored from it The channel characteristics that ensuing two training signals of two antenna for base station of remaininging are transmitted.Note, for the purpose of illustration, described three Individual training signal pattern is in separate time tranfer.If selected for three training signal patterns so as to orthogonal, then it Can with simultaneous transmission, therefore reduce the training time.

As shown in figure 5, after all three pilot transmission is completed, each customer set up 506-508 will have been stored 1 × 3 row 513-515 of matrix H send back base station 500.For purposes of simplicity, client's dress is only shown in Figure 5 Put 506 transmission its characteristic informations.With reference to appropriate Error Correction of Coding (such as Reed Solomon, Viterbi coding (Viterbi Coding) and/or strengthen coding (Turbo Codes)), it is possible to use suitable modulator approach (such as DPSK, 64QAM, OFDM) guaranteeing the data that base station 500 is accurately received in row 513-515.

In Fig. 5, although all three antenna 505 shows reception signal, for reception is per 1 × 3 row 513-515's Transmission, the single antenna and list transceiver of base station 500 is enough.However, under certain condition, using many or all antennas 505 and transceiver 504 come receive each transmission (that is, coding, modulation and signal processing subsystem 503 used in prior art The multiple output of single input (" SIMO ") treatment technology) can realize than single antenna 505 and the more preferable signal to noise ratio of transceiver 504 (SNR)。

Coding, modulation and signal processing subsystem 503 when base station 500 receives described 1 from each customer set up 507-508 When × 3 row 513-515, it is stored in 1 × 3 row 513-515 in 3 × 3 H-matrix 516.For customer set up, base Standing can carry out storage matrix 516 using many different memory technologies, and it includes, but are not limited to, non-volatile

mass storage Device (such as hard disk) and/or volatile memory (such as SDRAM).Fig. 5 shows that base station is received and stores from visitor The stage of 1 × 3 row 513 of family device 509.When 1 × 3 row 514 and 515 is transmitted from remaining customer set up, they can be with In being transmitted and being stored in H-matrix 516, until whole H-matrix 516 it is stored.

With reference to Fig. 6, will now be described from base station 600 to the embodiment of the MIDO of customer set up 606-608 transmission.Because every Individual customer set up 606-608 is independent device, so each device receives different data transfers.So, the reality of base station 600 Applying example includes being located at wan interface 601 and encodes, carries out liaison to them between modulation and signal processing subsystem 603 Router 602, it receives multiple data flows (form is bit stream), corresponds respectively to each customer set up from wan interface 601 The data flow is pressed separate data flow u by 606-608$_1$-u$_3$Send.For this purpose, the router 602 can using it is various The route technology known.

As shown in fig. 6, by three bit streams, u$_1$-u$_3$, it route into the coding, modulation and signal processing subsystem In 603, the error correction stream (for example, using Reed Solomon, Viterbi or enhancing coding) of statistical iteration is encoded that as, and With they are modulated to the suitable modulator approach of channel (such as DPSK, 64QAM or OFDM).Additionally, the embodiment bag that Fig. 6 shows Signal precoding logical block 630 is included, based on signal characteristic matrix 616, the signal precoding logical block 630 is used for from every The signal that individual antenna 605 sends carries out unique encodings.Especially, in this embodiment, precoding logical block 630 is by Fig. 6 In three bit stream u$_1$-u$_3$It is multiplied to generate three new bit stream u ' with the inverse matrix of H-matrix 616$_1$-u′$_3$, rather than will Each coding and modulated bit stream are routed to separate antenna (as done in Fig. 1).Then, D/A converter (does not show Go out) described three precoding bits circulation is changed to into analog signal, transceiver 604 and antenna 605 are sent out as RF signals See off.

Before explaining how customer set up 606-608 receives the not special stream, description precoding module 630 is performed Operation.Similar to

the example of MIMO in Fig. 1 above, the coding and modulated letter of each bit stream in three original bit streams Number will be indicated as u_n.In the embodiment shown in fig. 6, each u_iComprising the number that three bit streams that router 602 is route come According to, each such bit stream will become three user's set 606-608 one of those.

However, not as the MIMO examples in Fig. 1, there, each x_iThere is each antenna 104 to send, in the present invention shown in Fig. 6 Embodiment in, receive each u in each customer set up antenna 609_i(plus noise N any in upper signal channel).It is such to realize As a result, (we are denoted as v to the output of each in three antennas 605_i) it is u_iWith the H for characterizing each customer set up The function of matrix.In embodiment, the precoding logical block 630 in coding, modulation and signal processing subsystem is by performing Following equalities are calculating each v_i :

$v_1=u_1H^{-1}{}_{11}+u_2H^{-1}{}_{12}+u_3H^{-1}{}_{13}$

$v_2=u_1H^{-1}{}_{21}+u_2H^{-1}{}_{22}+u_3H^{-1}{}_{23}$

$v_3=u_1H^{-1}{}_{31}+u_2H^{-1}{}_{32}+u_3H^{-1}{}_{33}$

Therefore, unlike MIMO, there, channel calculates each x after translating the signals in $receiver_i$, and described herein Inventive embodiment solved each v before channel is translated the signals in $transmitter_i$.Each antenna 609 is received from other For the u of other antennas 609_{n-1}The u separated in bit $stream_i$.Each transceiver 610 is converted into the signal for respectively receiving Baseband signal, here A/D converter (not shown) it is digitized, each coding, modulation and signal processing subsystem 611 To its x_iBit stream is demodulated and decodes, and its bit is streamed to into (for example, the visitor of data-interface 612 that customer set up is used Application program on the device of family).

Embodiments of the invention described herein can be realized using various different codings and modulator approach.For example, In OFDM is realized, its intermediate frequency spectrum is divided into multiple split-band, and technology described herein can be used to characterize each single point Frequency band.However, as described above, the cardinal principle of the present invention is not limited to any specific modulator approach.

If customer set up 606-608 is portable data processing device, such as PDA, notebook computer and/or radio If words, then because customer set up may

move to another from a position, then channel characteristics can frequently change Become.So, in one embodiment of the invention, the channel characteristics matrix 616 of base station is constantly updated.In an enforcement Example in, base station 600 sends new training signal to each customer set up (per 250 milliseconds) periodically, each customer set up by its Channel eigenvectors constantly send back base station 600 to guarantee that channel characteristics are kept accurately (for example, if environment changes or client Device movement is so as to having influence on channel).In one embodiment, it is right in the actual data signal for be sent to each customer set up Training signal is interleaved.Typically, the handling capacity of the training signal is far below the handling capacity of the data-signal, therefore this The throughput total to system will have little to no effect.Correspondingly, in this embodiment, channel characteristics matrix 616 is in base station active Renewal can be continuously available when being communicated with each customer set up, next position is moved to from a position so as to work as customer set up Put, or environment to change and keep accurate channel characteristics so as to have influence on when channel.

One embodiment of the present of invention shown in Fig. 7 improves uplink communication channel (that is, from client using MIMO technology Channels of the device 706-708 to base station 700).In this embodiment, the uplink channel characteristics logical block 741 in base station is continuous Channel to coming from each customer set up is analyzed and characterizes.Especially, each customer set up 706-708 sends training letter Number to base station 700, there channel characteristics logical block 741 is analyzed to produce the channel characteristics matrix 741 of N × M, and here N is visitor The quantity of family device, M is the quantity of the antenna that base station is used.Embodiment shown in Fig. 7 uses three Hes of antenna 705 in base station Three customer set up 706-708, which results in 3 × 3 channel characteristics matrixes 741 for depositing in base station 700.Customer set up can be by MIMO uplink shown in Fig. 7 is used to send data back to base station 700 and channel eigenvectors are sent back into base station 700, such as Shown in Fig. 5.But with unlike the embodiment shown in Fig. 5, in Figure 5, the channel eigenvectors of each customer set up with point The time opened is transmitted, and the method shown in Fig. 7 allows simultaneously channel eigenvectors to be transmitted back to into

base from multiple customer set ups 700 are stood, so as to substantially reduce impact of the channel eigenvectors to Return Channel throughput.

As described above, the feature of each signal can include several factors, for example, it is included relative to inside receiver Reference signal, absolute reference signal, the phase place of relative reference signal, characteristic noise or other factors and amplitude.For example, orthogonal In the signal that amplitude modulation(PAM) is modulated, the feature can be the phase place of several multipath images of signal and amplitude excursion vector. Another example is, in the signal that OFDM is modulated, the feature can be several or all in OFDM frequency spectrums The phase place of single component signal and amplitude excursion vector.The training signal can be by the coding of each customer set up and modulation subsystem System 711 is generated, and the training signal is converted into analog signal by D/A converter (not shown), then the transmitter of each customer set up It is converted into RF signals by 709 from baseband signal.In one embodiment, in order to ensure the synchronization of training signal, customer set up Training signal (for example, in the case of circulation (round robin)) is only transmitted when base station requests.Furthermore, it is possible to Training signal is interleaved in the actual data signal sent from each customer set up, or training signal can with it is described Actual data signal is transmitted together.Therefore, even if customer set up 706-708 is mobile, uplink channel characteristics logical block 741 The training signal can also be continuously transmitted and analyze, so that it is guaranteed that channel characteristics matrix 741 keeps updating.

Total channel capacity that previously described embodiments of the present invention is supported can be defined as min (N, M), and here, M is visitor The quantity of family device, and N is the quantity of antenna for base station.That is, capacity is limited by the antenna amount of base station side or customer side It is fixed.Thus, one embodiment of the present of invention guarantees to be less than the individual antennas of min (N, M) within preset time using simultaneous techniques Sending/receiving.

In the typical case, the quantity of the antenna 705 of base station 700 will be less than the quantity of customer set up 706-708.Fig. 8 An exemplary situation is shown, it allows 5 customer set up 804-808 to be led to the base station with three antennas 802 Letter.In this embodiment, it is determined

that the quantity of total customer set up 804-808 and detecting necessary channel characteristic information After (for example, above description), base station 800 selects first crowd of three client 810 being in communication with (because min (N, M) =3, so being in this instance three clients).After the specified time that communicated with first crowd of client 810, base station just selects separately Three clients 811 that a group is communicated with.In order to evenly distribute communication channel, base station 800 selects to be not comprised in first group Two customer set ups 807,808.Further, since extra antenna is available, base station 800 just selects to be included in first group Extra customer set up 806.In one embodiment, base station 800 is circulated by this way in the client masses such that it is able to have Distribute to effect the handling capacity of each client equal number in time.For example, in order to evenly distribute handling capacity, base station can connect Any combinations of three customer set up of the selection in addition to customer set up 806 (that is, because customer set up 806 is used to start Two circulation in communicated with base station).

In one embodiment, in addition to the data communication of standard, base station can transmit training using aforementioned techniques Signal receives training signal and signal characteristic data to each customer set up and from each customer set up.

In one embodiment, some customer set ups or customer set up group can be assigned to the handling capacity of varying level, example Such as, customer set up can be distinguished order of priority, such that it is able to guarantee that the customer set up of of a relatively high priority must be relatively low preferential There are more communication cycles (that is, more handling capacities) at client's dress family of level.Based on a number of variable, can be to client's " priority " is selected, and the variable includes, for example, the subscription fee to wireless bandwidth of user is (for example, for being ready for volume Outer handling capacity pays more), and/or communication is to/from data type (for example, real-time Communication for Power, for example call voice of customer set up And video, obtain the priority higher than non-realtime traffic, such as Email).

In the present load required based on each customer set up, in the embodiment of base station dynamically distributes handling capacity.For

example, if The live video stream of customer set up 804, and other devices 805-808 is in the non real-time function of performing such as Email, then base Stand 800 can distribute relatively large number of handling capacity to the client 804.It is to be noted, however, that the cardinal principle of the present invention is not It is limited to any specific throughput distribution technology.

As shown in figure 9, two customer set ups 907,908 can be closely so that the channel characteristics of the client are in reality It is the same on border.As a result, base station by receive and store two customer set ups 907,908 actually equal channel characteristics to Amount, therefore this will can not produce, signal of spatial distribution unique for each client.Correspondingly, in one embodiment, base station Will ensure that mutual distance any two closely or more customer set ups are assigned to different groups.For example, in fig .9, Base station 900 communicates first with first group 910 of customer set up 904,905 and 908, then with customer set up 905,906,907 Second group 911 communication, which ensure that customer set up 907 and 908 in different groups.

Selectively, in one embodiment, base station 900 is communicated with customer set up 907 and 908 simultaneously, but is made Communication channel is multiplexed with known multi-channel Technology.For example, base station can be using time division multiplex (" TDM "), frequency Multiplexing (" FDM ") or CDMA (" CDMA ") technology is divided to separate single, space correlation between customer set up 907 and 908 Signal.

Although above-mentioned each customer set up is equipped with single antenna, can be come by using the customer set up with multiple antennas Realize the cardinal principle of the present invention to improve handling capacity.For example, when on the above-mentioned wireless system, with 2 antennas Client will realize 2 times of throughput hoisting, the client with 3 antennas by realize 3 times throughput hoisting, etc. (i.e., it is assumed that Space and angular separation between antenna is enough).When being circulated by the customer set up with multiple antennas, base station Same general rule can be applied.For example, it can regard each antenna as separate client, and give that by throughput distribution Individual " client ", is that any other client is the same (for example, it is ensured that each client is provided with enough or suitable communication weeks just as it Phase).

As described above, one embodiment of the present of invention uses above-mentioned MIDO and/or MIMO signal transmission technology intimate Signal to noise ratio and handling capacity are improved in vertical incidence sky wave (" NVIS ").With reference to Figure 10, in one embodiment of the invention, equipment The NVIS base stations 1001 for having the matrix of N number of antenna 1002 are used for and M customer set up 1004 is communicated.Described NVIS days The antenna 1004 of line 1002 and various user's sets with vertical direction about into the angle within 15 degree by signal uplink transmit with The NVIS that wants of acquisition and surface wave disturbing effect is preferably minimized.In one embodiment, antenna 1002 and customer set up 1004 assigned frequencies using above-mentioned various MIDO and MIMO technology in NVIS frequency spectrums (such as in carrier frequency or are less than The frequency of 23MHz, but usually less than in the frequency of 10MHz) multiple independent data flows 1006 are supported, so as to significantly improve The handling capacity (that is, being directly proportional with the quantity of the data flow with statistical iteration) of assigned frequency.

Serving the NVIS antennas of given base station can have each other far physical distance.Hypothesis is less than The long range (300 miles of round distances) that the long wavelength of 10MHz and signal are propagated, hundreds of code, even several miles of antenna Physical separation can provide benefit in diversity.In such a situa-tion, single aerial signal can be withdrawn into centre bit Put, it is processed with traditional wired or wireless communication system.Selectively, each antenna can have local device Its signal is processed, then the data transfer center is gone back to into using traditional wired or wireless communication system.In the present invention One embodiment in, NVIS base stations 1001 have to internet 1010 (or other wide area networks) wideband link 1015, so as to It is supplied to that customer set up 1003 is long-range, high speed, wireless network access.

In one embodiment, base station and/or user can utilize polarization/direction figure diversity (pattern Diversity) technology, while diversity is provided with handling capacity is lifted, to reduce array size and/or user distance.For example, In the DIMO systems transmitted with HF, due to polarization/direction figure diversity, user can be located at same position and their signal not Can be associated.Especially, by using directional diagram diversity, a user can be communicated via earthwave with base

station, and other are used Family can be communicated via NVIS with base station.

The Additional embodiments of the present invention

I、DIDO-OFDM precodings are carried out using I/Q imbalances

One embodiment of the present invention is using for the distribution of the distributed input with OFDM (OFDM) The system and method that inphase quadrature (I/Q) imbalance in formula output (DIDO) system is compensated.In short, according to this enforcement Mode, user equipment is estimated channel, and the information is fed back to into base station ; Base station calculates pre-coding matrix, to eliminate Interference between the caused carrier wave of the uneven institutes of I/Q and user between ; And parallel data stream is sent out via DIDO precodings Deliver to multiple user equipmenies ; The user equipment forces (ZF), least mean-square error (MMSE) or maximum likelihood (ML) to connect via zero Receipts machine being demodulated to data, to suppress residual interference.

As detailed below, some notable features of the embodiment of the invention include, but are not limited to :

Precoding for eliminate ofdm system in from mirror image adjust (mirror tone) inter-carrier interference (ICI) (because I/Q is mismatched to be caused) ;

Precoding is for the inter-user interference and ICI (being caused because I/Q is mismatched) in elimination DIDO-OFDM systems ;

For ICI being eliminated via the ZF receivers in the DIDO-OFDM systems using block diagonalization (BD) (because of I/Q not Matching caused) technology ;

For via the precoding (at emitter) in DIDO-OFDM systems and ZF or MMSE wave filters (in receiver Place) eliminating the technology of inter-user interference and ICI (being caused because I/Q is mismatched) ;

For via the precoding (at emitter) in DIDO-OFDM systems and similar to maximum likelihood (ML) detector Nonlinear detector (at receiver) eliminating the technology of inter-user interference and ICI (being caused because I/Q is mismatched) ;

Done for the intercarrier adjusted from mirror image in elimination ofdm system using the precoding based on channel condition information Disturb (ICI) (being caused because I/Q is mismatched) ;

Using the precoding based on channel condition information for the carrier wave in elimination DIDO-OFDM systems from mirror image tune Between interference (ICI) (because I/Q mismatch caused) ;

Known DIDO precoder (I/Q mismatch aware DIDO are mismatched in base station using I/Q Precoder DIDO receivers known to I/Q), and at user terminal are used ;

Known DIDO precoder (I/Q mismatch aware DIDO are mismatched in base station using I/Q Precoder), DIDO receivers known to I/Q are used at user terminal, and using I/Q known channel estimators ;

Known DIDO precoders are mismatched using I/Q in base station, uses DIDO known to I/Q to receive at user terminal Machine, and using DIDO feedback generators known to I/Q known channels estimator and I/Q (maker by channel condition information from User terminal is sent to website) ;

Known DIDO precoders are mismatched using I/Q in base station, and (this is matched somebody with somebody using DIDO configurators known to I/Q Put device and various functions performed using I/Q channel informations, including user select, adaptive coding and modulation, the mapping of space-time frequency or Precoder is selected) ;

Using DIDO receivers known to I/Q, (receiver is via the DIDO-OFDM using block diagonalization (BD) precoder ZF receivers in system come eliminate ICI (because I/Q mismatch caused)) ;

Using DIDO receivers known to I/Q, (receiver is via the precoding in DIDO-OFDM systems (at emitter) With similar to the nonlinear detector (at receiver) of maximum likelihood (ML) detector come eliminate inter-user interference and ICI (because I/Q is mismatched to be caused)) ; And

Using DIDO receivers known to I/Q, (receiver disappears via ZF the or MMSE wave filters in DIDO-OFDM systems Except ICI (is caused) because I/Q is mismatched).

a、Background

Exemplary radio communication system sends and receives signal comprising inphase quadrature (I/Q) component.In actual system, The inphase quadrature component may due to mixing and baseband operations in defect and distortion.These distortion (distortion) performances Mismatch for I/Q phase places, gain and delay.Unbalance in phase is by the sine (sine) and cosine in modulator/demodulator (cosine) it is incorrect orthogonal and caused.Gain imbalance is caused by the different amplification between inphase quadrature component 's.Due to the delay difference between I the and Q tracks (rail) in analog circuit, it is also possible to there is additional distortion, the distortion is referred to as It is uneven to postpone.

In OFDM (OFDM) system, I/Q imbalances can cause the intercarrier for carrying out spontaneous emission tune uneven (ICI).The impact is studied in some data, and in following information, it has been proposed that for defeated to single input list The I/Q gone out in SISO-OFDM systems mismatches the method for compensating : M.D.Benedetto and P.Mandarini, “Analysis of the effect of the I/Q baseband filter mismatch in an OFDM Modem, " Wireless personal communications, pp.175-186,2000 ; S.Schuchert and R.Hasholzner, " A novel I/Q imbalance compensation scheme for the reception of OFDM signals, " IEEE Transaction on Consumer Electronics, Aug.2001 ; M.Valkama, M.Renfors and V.Koivunen, " Advanced methods for I/Q imbalance compensation in Communication receivers, " IEEE Trans.Sig.Proc, Oct.2001 ; R.Rao and B.Daneshrad, " Analysis of I/Q mismatch and a cancellation scheme for OFDM systems, " IST Mobile Communication Summit, June2004 ; A.Tarighat, R.Bagheri and A.H.Sayed, “Compensation schemes and performance analysis of IQ imbalances in OFDM Receivers, " Signal Processing, IEEE TransactiohS on [referring also to Acoustics, Speech, And Signal Processing, IEEE TransactiohS on], vol.53, pp.3257-3268, Aug.2005.

Show the work to the extension of multiple-input and multiple-output MIMO-OFDM systems in following information : R.Rao and B.Daneshrad, " I/Q mismatch cancellation for MIMO OFDM systems, " in Personal, Indoor and Mobile Radio CommunicatiohS, 2004 ; PIMRC2004.15th IEEE International Symposium on, vol.4,2004, pp.2710-2714.For spatial reuse (SM), R.M.Rao, W.Zhu are referred to, S.Lang, C.Oberli, D.Browne, J.Bhatia, J.F.Frigon, J.Wang, P ; Gupta, H.Lee, D.N.Liu, S.G. Wong, M.Fitz, B.Daneshrad, and

O.Takeshita, " Multiantenna testbeds for Research and education in wireless communications, " IEEE Communications Magazine, vol.42, no.12, pp.72-81, Dec.2004 ; S.Lang, M.R.Rao and B.Daneshrad, " Design and development of a5.25GHz software defined wireless OF DM communication Platform, " IEEE Communications Magazine, vol.42, no.6, pp.6-12, June2004 ; For orthogonal Space-Time Block Coding (OSTBC), refers to A.Tarighat and A.H.Sayed, " MIMO OFDM receivers for Systems with IQ imbalances, " IEEE Trans.Sig.Proc, vol.53, pp.3583-3596, Sep.2005.

Unfortunately, how introduction is not currently existed in distributed input distributed output (DIDO) communication system The data that I/Q gains and unbalance in phase error are corrected.Invention described below embodiment provides a kind of solution The scheme of these problems.

DIDO systems include a base station with spaced antenna, and the base station is using the nothing for being same as traditional SIO systems Line resource (that is, identical time-slot duration and frequency band), sends parallel data stream (Jing precodings) to multiple users, to increase Strong downlink throughput.S.G. entitled " the System and that Perlman and T.Cotter was submitted in July 30 in 2004 The Shen of Method for Distributed Input-Distributed Output Wireless Communications " Please No.10/902,978 (" earlier applications ") give the detailed description of DIDO systems, and this application is transferred and gives the application's Assignee, and this application is incorporated herein as reference.

There are various ways and realize DIDO precoders.A kind of scheme is the block diagonalization described in following information (BD) : Q.H.Spencer, A.L.Swindlehurst and M.Haardt, " Zero forcing methods for Downlink spatial multiplexing in multiuser MIMO channels, " IEEE Trans.Sig.Proc, vol.52, pp.461-471, Feb.2004 ; K.K.Wong, R.D.Murch, and K.B.Letaief, " A Joint channel diagonalization for multiuser MIMO antenna systems, " IEEE Trans.Wireless Comm., vol.2, pp.773-786, Jul2003 ; L.U.Choi and R.D.Murch, " A transmit preprocessing technique for multiuser MIMO systems using a decomposition Approach, " IEEE Trans.Wireless Comm., vol.3, pp.20-24, Jan2004 ; Z.Shen, J.G.Andrews, R.W.Heath and B.L.Evans, " Low complexity user selection algorithms For multiuser MIMO systems with block diagonalization, " IEEE is published in by acceptance Trans.Sig.Proc, Sep.2005 ; Z.Shen, R.Chen, J.G.Andrews, R.W.Heath and

B.L.Evans, " Sum Capacity of multiuser MIMO broadcast channels with block diagonalization, " quilt It is committed to IEEE Trans.Wireless Comm., Oct.2005 ; R.Chen, R.W. Heath, and J.G. Andrews, “Transmit selection diversity for unitary precoded multiuser spatial Multiplexing systems with linear receivers, " received to IEEE Trans, on Signal Processing, 2005.The method for I/Q compensation gone out given in these materials contemplates BD precoders, and this prelists Code device can be scaled up to any types of DIDO precodings.

In DIDO-OFDM systems, I/Q is mismatched can cause two kinds of impacts ：ICI and inter-user interference.With SISO-OFDM Similar in system, the former is that the interference due to adjusting from mirror image is caused.The latter is due to the fact that what is caused, i.e. I/Q Mismatch can destroy the orthogonal of DIDO precoders, so as to produce interference between users.Can by method described herein, This two classes interference is eliminated at transmitter and receiver.There is described herein three kinds of sides for the I/Q compensation in DIDO-OFDM systems Method, and for having and not with I/Q mismatches, comparing their performance.Based on using performed by DIDO-OFDM prototypes Emulation and actual measurement, illustrate result.

Present embodiment is the extension of earlier application.Especially, these embodiments have with the following characteristics of earlier application Close ：

System described in earlier application, wherein I/Q tracks can be affected by gain and unbalance in phase ；

At emitter, the DIDO calculated using the training signal adopted for channel estimation with I/Q compensation is pre- Encoder ； And

Signal characteristic data take into account due to distortion caused by the uneven institutes of I/Q, and at emitter, according to this material The method for being proposed, using the signal characteristic data DIDO precoders are calculated.

b、Embodiments of the present invention

First, by the Mathematical Modeling and framework of the description present invention.

Before this programme is shown, explain that core mathematics concept is highly useful.We are by assuming that I/Q gains and phase Position is uneven, and (comprising phase delay in this description, but the phase delay will be automatic in the algorithm of DIDO-OFDM forms Process) explaining to it.To explain basic thought, it is assumed that we are want two plural number $s=s_I+js_Q$And $h=h_I+jh_Q$It is multiplied, And so that x=h*s.We represent inphase quadrature component using subscript.Call below equation：

$X_I=S_Ih_I-S_Qh_Q$

And

$x_Q=S_Ih_Q+S_Qh_I$

Its matrix form is rewritable to be：

Normalization conversion is marked by channel matrix (H).It is now assumed that s is sent symbol, and h is channel.Can lead to Cross the following non-normalized change of establishment and bring and I/Q gains and the presence of unbalance in phase are modeled：

The effect of the skill is to confirm to be written as：

Now (A) is written over：

We carry out defined below：

And

The two matrixes have normalization structure, therefore may be expressed as plural form：

$h_e=h_{11}+h_{22}+j(h_{21}-h_{12})$

And

$h_c=h_{11}-h_{22}+j(h_{21}+h_{12})$

By using all these knowledge, we can derive back effective equation with two channel (equivalent channels h_eWith Conjugate channel h_c) scalar form.Therefore, the efficient transformation in (5) is changed into：

$x=h_es+h_cs^*$

First channel is referred to as equivalent channels by us, and second channel is referred to as conjugate channel.If there is no I/Q gains and phase Position is uneven, then the equivalent channels are our channels to be observed.

By using similar argument, discrete time MIMON with I/Q gains and unbalance in phase × M systems Input output Relationship can be shown as (their matrix corresponding form is set up by using the scalar equivalent form of value) :

Wherein, t is discrete time index, h_e, $h_c \in C^{M \times N}$, $s=[s_1..., s_N]$, $x=[x_1..., x_M]$ and L is channel Tap (channel tap) number.

In DIDO-OFDM systems, received signal in frequency domain is illustrated.If meeting below equation, from signal Re-call with system :

$FFT_K\{ s [t] \}=S [k]$ then $FFT_K\{s^*[t] \}=S^*[(- k)]=S^*[K-k]$ for k=0,1 ..., K-1

Using OFDM, for subcarrier k, the Input output Relationship of equal value of MIMO-OFDM systems is :

Wherein, k=0,1 ..., K-1 be OFDM subcarrier indexes, H_eAnd H_cOf equal value and conjugate channel matrix is represented respectively, It is defined as follows :

And

(1) the second base value in is the interference adjusted from mirror image.Can be by building following stacked (stacked) matrix system (please carefully noting conjugate) is processing it :

WhereinWithRespectively send and receive symbol vector in a frequency domain.

By using the method, active matrix can be set up, for DIDO operations.For example, using DIDO2 × 2 input-it is defeated Go out relation (assuming that each user has single reception antenna), first user equipment is contemplated that below equation (does not have noise When) :

And second user notes below equation :

Wherein,Matrix H is represented respectively$_e$And H_cM rows, and $w \in C^{4x4}$For DIDO precoding squares Battle array.According to (2) and (3), it may

be noted that the symbol that user m is receivedBy two interference sources caused by the uneven institutes of I/Q (that is, from mirror image adjust inter-carrier interference (i.e.,And inter-user interference is (i.e.,And , p ≠ m)) impact.(3) the DIDO pre-coding matrix W in are designed to eliminate the two distracters.

There are multiple different embodiments in the DIDO precoders that can be used for herein, this depends on being applied at receiver Joint-detection.In one embodiment, can adopt according to composite channel(rather than) calculated Block diagonalization (BD) (refer to for example, Q.H.Spencer, A.L.Swindlehurst, and M.Haardt, " Zeroforcing Methods for downlink spatial multiplexing in multiuser MIMO channels, " IEEE Trans.Sig.Proc, vol.52, pp.461-471, Feb.2004.K.K ; Wong, R.D.Murch, and K.B.Letaief, " A Joint channel diagonalization for multiuser MIMO antenna systems, " IEEE Trans.Wireless Comm., vol.2, pp.773-786, Jul2003 ; L. U.Choi and R.D.Murch, " A transmit preprocessing technique for multiuser MIMO systems using a Decomposition approach, " IEEE Trans.Wireless Comm., vol.3, pp.20-24, Jan2004 ; Z.Shen, J.G. Andrews, R.W. Heath, and B.L Evans, " Low complexity user selection Algorithms for multiuser MIMO systems with block diagonalization, " delivered by acceptance In IEEE Trans.Sig.Proc, Sep.2005 ; Z.Shen, R.Chen, J.G. Andrews, R.W. Heath, and B.L Evans, " Sum capacity of multiuser MIMO broadcast channels with block Diagonalization, " it is submitted to IEEE Trans.Wireless Comm., Oct.2005).Therefore, current DIDO systems System selects precoder, so that :

Wherein, $\alpha_{I,j}$For constant, andThe method is highly profitable, because being prelisted by using this Code device, because the impact that I/Q gains and unbalance in phase are completely eliminated at emitter can make other of DIDO precoders Aspect is kept intact.

Also DIDO precoders can be designed as eliminating inter-user interference in advance, and not eliminated in advance because IQ imbalances are led The ICI of cause.Using the method, receiver (rather than emitter) can be by one of receiving filter using described below come to IQ Imbalance is compensated.Therefore, the Precoding Design standard in (4) can be modified to :

And

Wherein for m-th sending signal,And The symbolic vector received by user m.

In receiving side, in order to sending symbolic vectorEstimated, user m adopts ZF wave filters, and estimated Symbolic vector is given as：

Although ZF wave filters are easiest to understand that receiver can also be applied any number of known to one of skill in the art Other wave filters.A kind of popular selection is MMSE wave filters, wherein：

And ρ is signal to noise ratio.Alternatively, the executable maximum likelihood symbol detection of user (or change of Sphere decoder device, iteration). For example, first user can use ML receivers, and solve following optimization：

Wherein, S is the set of all possible vectorial s, and depending on constellation sizes.The ML receivers are given can be preferable Performance, but higher complexity is required at receiver.One group of similar equation can be applicable to second user.

Note, in (6) and (7)WithIt is assumed to be with zero.The hypothesis is only in transmitting precoder Can be completely eliminated in the case of the inter-user interference for the standard in (4) effectively.Similar,WithOnly exist Transmitting precoder is diagonal matrix in the case of inter-carrier interference (that is, adjusting from mirror image) can be completely eliminated.

Figure 13 shows a kind of embodiment of the framework of the DIDO-OFDM systems with I/Q compensation, the DIDO- Ofdm system include be located at base station (BS) in IQ-DIDO precoders 1302, send channel 1304, in user equipment Channel estimation logic 1306 and ZF, MMSE or ML receiver 1308.The channel estimation logic 1306 is via training signal To channelWithEstimated, and these estimations are fed back to the precoder in AP.BS calculates DIDO precoders Weight (matrix W), to eliminate in advance because of interference and user's interference caused by I/Q gains and unbalance in phase institute, and by data Sent to user by wireless channel 1304.User equipment m adopts ZF, MMSE or ML receiver 1308, by

using unit 1304 channel estimations for being provided are demodulated eliminating residual interference to data.

This I/Q backoff algorithm can be realized using three below embodiment.

Method 1-TX is compensated : In this embodiment, standard of the emitter in (4) is calculating pre-coding matrix. At receiver, user equipment adopts " simplification " ZF receivers, whereinWithIt is assumed to be diagonal matrix.Therefore, Formula (8) is reduced to :

Method 2-RX is compensated : In this embodiment, emitter is based on R.Chen, R.W.Heath, and J.G. Andrews, " Transmit selection diversity for unitary precoded multiuser spatial Multiplexing systems with linear receivers, " accepted to IEEE Trans, on Signal Processing, the traditional BD methods described in 2005 calculate pre-coding matrix, and are not for the standard in (4) eliminating load Between ripple and inter-user interference.Using the method, the pre-coding matrix in (2) and (3) is reduced to :

At receiver, user equipment adopts ZF wave filters as in (8).Note, the method is not such as above-mentioned method 1 Like that, interference is eliminated in advance at emitter.Therefore, it eliminates inter-carrier interference at receiver, but can not eliminate user Between disturb.Additionally, requiring to feed back compared to method 1WithIn method 2, user only needs feedback for emitter VectorTo calculate DIDO precoders.Therefore, method 2 is particularly suitable for the DIDO systems with low-rate feedback channel. On the other hand, method 2 needs have somewhat higher computation complexity at user equipment, to calculate in (8) (rather than (11)) ZF receivers.

Method 3-TX-RX is compensated : In one embodiment, above-mentioned two method is merged.Emitter is counted as (4) Pre-coding matrix is calculated, and receiver is estimated sending symbol according to (8).

I/Q uneven (either unbalance in phase, gain are uneven, or are to postpone imbalance) can be to radio communication system Signal quality in system causes harmful degradation.For that reason, conventional circuit is both designed to relatively low imbalance pin. However, as described above, can correct this and ask by using the Digital Signal Processing of transmitting precoded form and/or specific receiver Topic.One embodiment of the

present invention includes the system with multiple New function units, and each unit is for realizing ofdm communication I/Q corrections in system or DIDO-OFDM communication systems are critically important.

One embodiment of the present invention using based on channel condition information precoding, to eliminate ofdm system in from The inter-carrier interference (ICI) (causing because I/Q is mismatched) that mirror image is adjusted.As shown in figure 11, launched according to the DIDO of present embodiment Machine includes subscriber selector unit 1102, multiple coded modulation units 1104, corresponding multiple map units 1106, DIDO IQ Known precoding unit 1108, multiple RF transmitter units 1114, user feedback unit 1112 and DIDO configurator units 1110。

The subscriber selector unit 1102 is selected and multiple users based on the feedback information acquired in feedback unit 1112 U_1-U_MAssociated data, and provide this information to each coded modulation unit in multiple coded modulation units 1104 1104.Each coded modulation unit 1104 is encoded and demodulated to the information bit of each user, and sends them to reflect Penetrate unit 1106.Input bit is mapped to complex symbol by the map unit 1106, and result is sent to pre- known to DIDO IQ Coding unit 1108.The channel conditions that the precoding unit 1108 known to DIDO IQ is obtained using feedback unit 1112 from user Information, calculates precoding weight known to DIDO IQ, and the incoming symbol to obtaining from map unit 1106 carries out precoding.Often One pre-code data stream is sent to OFDM units 1115 by precoding unit 1108 known to DIDO IQ, the OFDM units 1115 calculate IFFT, and add Cyclic Prefix.The information is sent to D/A units 1116, and the D/A units 1116 carry out digital-to-analogue and turn Change, and send it to RF units 1114.The RF units 1114 to intermediate frequency/radio frequency, and send it to baseband signal raising frequency Transmitting antenna.

The precoder is operated together to the conventional mirror image tune that reconciles, uneven to compensate I/Q.Arbitrary Digit can be used The precoder design standard of amount, including ZF, MMSE or weighting MMSE designs.In a preferred embodiment, precoder can be complete Remove entirely because I/Q mismatches ICI caused by institute, so that receiver need not perform any ancillary relief.

In one embodiment, the precoder uses block diagonalization standard, not exclusively to eliminate each user I/Q affect (this needs accessory receiver to process) in the case of, inter-user interference is completely eliminated.In another embodiment, The precoder is completely eliminated because of inter-user interference and ICI interference caused by the uneven institutes of I/Q using zero pressure standard. The embodiment can use traditional DIDO-OFDM processors at receiver.

One embodiment of the present invention is used based on the precoding of channel condition information, to eliminate DIDO-OFDM systems In from mirror image adjust inter-carrier interference (ICI) (because I/Q mismatch caused), and each user adopt known to IQ DIDO reception Machine.As shown in figure 12, in one embodiment of the invention, system (including receiver 1202) includes multiple RF units 1208th, correspondingly multiple A/D units 1210, IQ known channels estimator 1204 and DIDO feedback generator units 1206。

The RF units 1208 receive the signal sent from DIDO transmitter units 1114, by the signal down to base band, And the signal after the frequency reducing is supplied to into A/D units 1210.Afterwards, 1210 pairs of signals of the A/D units carry out analog-to-digital conversion, and Send it to OFDM units 1213.The OFDM units 1213 remove Cyclic Prefix, and carry out FFT, by the signal reports extremely Frequency domain.During cycle of training, OFDM units 1213 transmit output to IQ known channels estimation unit 1204, the known letters of the IQ Road estimation unit 1204 calculates in a frequency domain channel estimation.Alternatively, the channel estimation can in the time domain be calculated.In data week During phase (data period), OFDM units 1213 transmit output to receiver unit 1202 known to IQ.The IQ is known to be received Machine unit calculates IQ receivers, and the signal is demodulated/is decoded, to obtain data 1214.The IQ known channels are estimated Meter unit 1204 sends the channel estimation to DIDO feedback generators unit 1206, and the feedback generator unit 1204 can be to institute State channel estimation to be quantified, and emitter is sent it back to via feedback control channel 1112.

Receiver 1202 shown in Figure 12 can any number of standard known in those skilled in the art (including ZF, MMSE, maximum likelihood or MAP receivers) under work.In a preferred embodiment, receiver is eliminated

using MMSE wave filters ICI caused by the uneven institutes of IQ adjusted because of mirror image.In another preferred embodiment, receiver is used similar to maximum likelihood The nonlinear detector of search carrys out the symbol that joint-detection mirror image is adjusted.The method has good performance, but with higher Complexity.

In one embodiment, receiver coefficient is determined using IQ known channels estimator 1204, to remove ICI. Therefore, we require that DIDO-OFDM systems (using the load eliminated based on the precoding of channel condition information from mirror image tune Disturb between ripple (ICI) (being caused because I/Q is mismatched)), the rights and interests of DIDO receivers known to IQ and IQ known channel estimators. The channel estimator can use traditional training signal, or the instruction of the special structure for being usable in being sent on inphase quadrature signal Practice signal.Any number of algorithm for estimating, including least square method, MMSE or maximum likelihood can be implemented.The IQ known channels Estimator receiver known to IQ provides input.

Channel condition information can be provided to website by channel reciprocity or by feedback channel.The one of the present invention is real Mode is applied including DIDO-OFDM systems, the system has precoder known to I/Q, and for the letter of user terminal in future Feedback channel known to the I/Q that road condition information transmits to website.The feedback channel can be physically or logically control channel.It can be By special or shared in RACH.Can be by using user terminal (we have also required that the rights and interests of the user terminal) place DIDO feedback generators generating feedback information.The DIDO feedback generators are defeated by the I/Q known channels estimator Go out as input.It can quantized channel coefficient, or any amount Limited Feedback algorithm known in the field can be used.

The distribution of user, modulation and encoding rate, the mapping to space-time frequency coding time slot can be generated according to DIDO feedbacks The result of device and change.Therefore, an embodiment includes DIDO configurators known to IQ, and the configurator is used from one or more The IQ known channels of user are estimated to configure precoder known to DIDO IQ, selection modulation rate, encoding rate, the

use for allowing transmission The subset at family and their mapping to space-time frequency coding time slot.

In order to evaluate the performance of proposed compensation method, three DIDO2 × 2 systems will be compared：

1st, mismatch with I/Q：Be transmitted by all of tune (adjusting except DC reconciles edge), and not to I/Q not With compensating；

2nd, compensate with I/Q：It is transmitted by all of tune, and comes to mismatch I/Q by using above-mentioned " method 1 " Compensate；

3rd, preferably：Only adjusted by odd number and be transmitted, to avoid inter-user interference and be caused because I/Q is mismatched Intercarrier (that is, from mirror image adjust) interference.

After this, illustrate in true propagation situation and measure acquired result using DIDO-OFDM prototypes.Figure 14 depict from the 64-QAM planispheres acquired in above three system.These planispheres are in same customer location and fixation Obtain in the case of average signal-to-noise ratio (～45dB).First planisphere 1401 is very noisy (because I/Q imbalances are led Cause the interference adjusted from mirror image).Second planisphere 1402 shows that some are improved (due to I/Q compensation).Note, the second planisphere 1402 ideal situations not shown in planisphere 1403 are pure like that (may to produce inter-carrier interference (ICI) due to existing Phase noise).

Figure 15 show with and have I/Q it is unmatched in the case of, DIDO2 × 2 of 64-QAM and 3/4 encoding rate The average SER (SER) 1501 of system and every user's goodput (goodput) 1502.OFDM is with a width of 250KHZ, adjusts and circulating prefix-length L with 64_{cp}=4.Because in the ideal case, we are only sent by the subset adjusted Data, therefore according to the transmission power (rather than total transmission power) averagely often adjusted evaluating SER and goodput performance, with Ensure the fair comparison between different situations.Additionally, in following result, we (are divided using the normalized value of transmission power Shellfish indicates) because the target of ours is to compare relative (rather than absolute) performance of different schemes herein.Figure 15 shows exist In the case of I/Q is unbalanced, SER saturations and miss the mark SER (～

10^{-2}), this and A.Tarighat and A.H.Sayed, " MIMO OFDM receivers for systems with IQ imbalances, " IEEE The result reported in Trans.Sig.Proc, vol.53, pp.3583-3596, Sep.2005 is consistent.The saturation effect be by On the fact that caused, i.e., signal power and jamming power (adjusting from mirror image) increase with the increase of TX power.So And, pass through proposed I/Q compensation methodes, interference can be eliminated, and obtain preferably SER performances.Note, due to 64-QAM modulation Larger transmission power is needed, therefore, because of the amplitude saturation effect in DAC SER can be caused to have at high SNR trickle Increase.

Additionally, can be observed, existing in the case that I/Q compensates, SER performances closely ideal situation.This two kinds of feelings Between condition, the 2dB gaps of TX power are that (phase noise may produce additional between adjacent OFDM is adjusted due to phase noise Interference) cause.Finally, goodput curve 1502 shows that it can send out compared to ideal situation when using I/Q methods The data of twice are sent, because we used, all of data are adjusted rather than only odd number is adjusted (for ideal situation).

Figure 16 is illustrated in the case of with I/Q compensation or not having I/Q compensation, the SER of different Q AM planisphere Energy.We can be observed, and in this embodiment, the method for being proposed is particularly advantageous for 64-QAM planispheres. For 4-QAM and 16-QAM, I/Q compensation methodes meeting producing ratio has the worse performance of the unmatched situations of I/Q, and this may It is because that the interference that proposed method requires bigger power to carry out data is activation and adjust from mirror image is eliminated.Additionally, The unmatched shadows of I/Q are subject to due to the larger minimum range between constellation point, 4-QAM and 16-QAM and not as 64-QAM Ring.Referring to A.Tarighat, R.Bagheri, and A.H.Sayed, " Compensation schemes and performance Analysis of IQ imbalances in OFDM receivers, " Signal Processing, IEEE Transactions on are [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.53, pp.3257-3268, Aug.2005.Also observable Figure 16 and by by I/Q mismatch with It is compared for the ideal situation of 4-QAM and 16-QAM and draws the conclusion.When therefore, for 4-QAM and 16-QAM For, the secondary power required for the DIDO precoders eliminated (adjust from mirror image)

with interference can not be compensated for I/Q Slight interests go bail for.Note, can be by solving the problem using above-mentioned I/Q compensation methodes 2 and 3.

Finally, under different propagation conditions, the relative SER performances of above three method are measured.Also describe and depositing It is for reference in the SER performances of the unmatched situations of I/Q.It is 450.5MHZ and with a width of that Figure 17 is depicted for carrier frequency 64-QAM DIDO2 × 2 system of 250KHz, in the SER measured by two different customer locations.At position 1, user and place BS in not chummery and in NLOS (without sighting distance) state is at a distance of∼6 λ.In position 2, user and the BS with LOS (sighting distance) At a distance of∼λ.

Figure 17 shows a case that all three compensation method is had outstanding performance than not compensating.It should be noted, however, that , under any channel conditions, method 3 surpasses other two kinds of compensation methodes.The relative performance of method 1 and 2 depends on passing Broadcast situation.By actual measurement activity, can show that method 1 generally surpasses method 2, because it is eliminated in advance (in emitter Place) the caused inter-user interference of the uneven institutes of I/O.When the inter-user interference very little, as shown in the curve map 1702 of Figure 17, side Method 2 can surpass method 1, because it is not exposed to because of power attenuation caused by I/Q compensation precoder institute.

Up to the present, by only considering limited group of propagation situation (as shown in figure 17) distinct methods have been compared Compared with.After this, the relative performance of these methods is measured in preferable i.i.d. (independent and tool is with distribution) channel.Using send out Penetrate I/Q phase places and gain imbalance with receiving side to emulate DIDO-OFDM systems.Figure 18 shows and have in only transmitting pusher side (that is, there is gain 0.8 on the I rails of the first transmitting chain, there is gain 1 on other rails) in the case of gain balance, institute The performance of the method for proposition.Can be seen that, method 3 has surpassed every other method.Additionally, with position in the curve map 1702 of Figure 17 Obtain result at 2 to compare, in i.i.d. channels, the comparable method 2 of method 1 is better carried out.

Therefore, the I/Q in giving three kinds of novel methods to compensate above-mentioned DIDO-OFDM systems is uneven, and method 3 surpasses Other compensation methodes for being proposed.In the system with low-rate feedback channel, can using method 2 reducing DIDO precodings Required feedback quantity, but poor SER performances can be caused.

II、Self adaptation DIDO delivery plan

Description is used to strengthen the another of the system and method for the performance of distributed input distributed output (DIDO) system Embodiment.The channel status that the method passes through tracking change, dynamically gives different user equipmenies by radio resource allocation, with Increase handling capacity while some target error rates are met.The user equipment estimated channel quality, and its is anti- It is fed to base station (BS) ; The base station is processed the channel quality for being obtained from user equipment, to select to be used for what is sent next time Optimal user cluster tool, DIDO schemes, modulation/coding scheme (MCS) and array configuration ; The base station is via precoding Parallel data is sent to multiple user equipmenies, and signal is demodulated at receiver.

One system for effectively distributing resource for DIDO Radio Links is also described.The system is included with DIDO configurators DIDO base stations, the base station is processed the feedback for receiving personal family, with select optimal user set for sending next time, DIDO schemes, modulation/coding scheme (MCS) and array configuration ; Receiver in DIDO systems, the receiver to channel and its He measures relevant parameter, to generate DIDO feedback signals ; And DIDO feedback control channels, for by from the anti-of user Feedforward information is transferred to base station.

As detailed in the following, some notable features of the embodiment of the invention may include, but be not limited to :

For based on channel quality information, be adaptive selected number of users, DIDO delivery plans (that is, day line options or Multiplexing), modulation/coding scheme (MCS) and array configuration, to minimize SER, or maximize spectrum efficiency per user or under The technology of line link spectrum efficiency ;

For defining multigroup DIDO sending modes using the technology as DIDO schemes and the combination of MCS ;

For giving different time slots, the OFDM skill for reconciling DIDO subflows by different DIDO mode assignments according to channel status Art ;

For different DIDO patterns to be dynamically assigned to the technology of different user based on the channel quality of different user ;

For being swashed to the switching of self adaptation DIDO based on the link quality metric calculated in time domain, frequency domain and spatial domain Standard living ;

For the standard of line activating to be switched into self adaptation DIDO based on look-up table.

The DIDO systems in base station with DIDO configurators as shown in figure 19, the system can be believed based on channel quality Breath, be adaptive selected number of users, DIDO delivery plans (that is, day line options or multiplexing), modulation/coding scheme (MCS) with And array configuration, to minimize SER, or maximize the spectrum efficiency or downlink tone spectrum efficiency per user ;

As shown in figure 20 having DIDO configurators and at each user equipment there is DIDO to feed back in base station is generated The DIDO systems of device, the system is using the other specification at estimated channel conditions and/or receiver (similar to estimated SNR), with generate be input into DIDO configurators feedback message.

DIDO systems, the system has DIDO configurators (in base station), DIDO feedback generators and DIDO feedback control Channel processed (the DIDO feedback channels are used to DIDO specific configuration informations are transmitted to base station from user).

a、Background

In multiple-input and multiple-output (MIMO) system, it is contemplated that diversity scheme (for example, orthogonal space time packet (OSTBC) (referring to V.Tarokh, H.Jafarkhani, and A.R.Calderbank, " Spacetime block codes from Orthogonal designs, " IEEE Trans.Info.Th., vol.45, pp.1456-467, Jul.1999) or day line selection Select (referring to R.W.Heath Jr., S.Sandhu,

and A.J.Paulraj, " Antenna selection for spatial Multiplexing systems with linear receivers, " IEEE Trans.Comm., vol.5, pp.142- 144, Apr.2001), to prevent fading channel, improve link reliability (reliability can be exchanged into more preferably coverage rate).Separately On the one hand, spatial reuse (SM) can be sent using multiple parallel datas and carry out strengthening system handling capacity as means.Referring to G.J.Foschini, G.D.Golden, R.A.Valenzuela, and P.W.Wolniansky, " Simplified processing for high spectral effciency wireless communication employing Multielement arrays, " IEEE Jour.Select.Areas in Comm., vol.17, no.11, pp.1841- 1852, Nov.1999.According to from L.Zheng and D.N.C.Tse, " Diversity and multiplexing: a Fundamental tradeoff in multiple antenna channels, " IEEE Trans.Info.Th., The theoretical diversity of vol.49, no.5, pp.1073-1096, May2003/multiplexing compromise, these benefits can be same in mimo systems Shi Shixian.One actual form of implementation is the channel status by tracking change, is carried out certainly between diversity and multiplexing delivery plan Adapt to switching.

It has been proposed that a large amount of adaptive MIMO transmission technologies.R.W.Heath and A.J.Paulraj, " Switching Between diversity and multiplexing in MIMO systems, " IEEE Trans.Comm., vol.53, Diversity in no.6, pp.962-968, Jun.2005/multiplexing changing method is designed to based on momentary channel quality information, changes Enter the BER (bit error rate) sent for fixed rate.Alternatively, can such as S.Catreux, V.Erceg, D.Gesbert, and R.W.Heath.Jr., " Adaptive modulation and MIMO coding for broadband wireless Data networks, " IEEE Comm.Mag., adopt in vol.2, pp.108-115, June2002 (" Catreux ") like that Statistic channel information entering line activating to self adaptation, so as to reduce the quantity of feedback overhead and control message.In Catreux Self adaptation transmission algorithm be designed to based on channel when/frequency selection designator, in OFDM (OFDM) system The predeterminated target bit error rate, strengthens spectrum efficiency.It is also directed to narrowband systems, it is proposed that similar low feedback adaptive method, the party Method is selectively come to be switched between diversity scheme and spatial reuse using channel space.See, for example, A.Forenza, M.R.McKay, A.Pandharipande, R.W.Heath.Jr., and I.B.Collings, " Adaptive MIMO Transmission for exploiting the capacity of spatially correlated channels, " Accepted to the IEEE Trans, on Veh.Tech.,

Mar.2007 ; M.R.McKay, I.B.Collings, A.Forenza, and R.W.Heath.Jr., " Multiplexing/beamforming switching for coded MIMO in spatially correlated Rayleigh channels, " received to IEEE Trans, on Veh.Tech., Dec.2007 ; A.Forenza, M.R.McKay, R.W.Heath.Jr., and I.B.Collings, “Switching between OSTBC and spatial multiplexing with linear receivers in Spatially correlated MIMO channels, " Proc.IEEE Veh.Technol.Conf., vol.3, Pp.1387-1391, May2006 ; M.R.McKay, I.B.Collings, A.Forenza, and R.W.Heath Jr., " A throughput-based adaptive MIMO BICM approach for spatially correlated Channels, " occur in Proc.IEEE ICC, June2006.

In the data, the working range represented in various previous disclosures is extended to DIDO-OFDM systems by us.Ginseng See such as R.W.Heath and A.J.Paulraj, " Switching between diversity and multiplexing in MIMO systems, " IEEE Trans.Comm., vol.53, no.6, pp.962-968, Jun.2005 ; S.Catreux, V.Erceg, D.Gesbert, and R.W.Heath Jr., " Adaptive modulation and MIMO coding for Broadband wireless data networks, " IEEE Comm.Mag., vol.2, pp.108-115, June2002 ; A.Forenza, M.R.McKay, A.Pandharipande, R.W.Heath Jr., and I.B.Collings, " Adaptive MIMO transmission for exploiting the capacity of spatially correlated Channels, " IEEE Trans, on Veh.Tech., vol.56, n.2, pp.619-630, Mar.2007 ; M.R.McKay, I.B.Collings, A.Forenza, and R.W.Heath Jr., " Multiplexing/beamforming switching For coded MIMO in spatially correlated Rayleigh channels, " received to IEEE Trans, On Veh.Tech., Dec.2007 ; A.Forenza, M.R.McKay, R.W.Heath Jr., and I.B.Collings, “Switching between OSTBC and spatial multiplexing with linear receivers in Spatially correlated MIMO channels, " Proc.IEEE Veh.Technol.Conf., vol.3, Pp.1387-1391, May2006 ; M.R.McKay, I.B.Collings, A.Forenza, and R.W.Heath Jr., " A throughput-based adaptive MIMO BICM approach for spatially correlated Channels, " occur in Proc.IEEE ICC, June2006.

There is described herein NEW ADAPTIVE DIDO sending strategy, the strategy with based on channel quality information in varying number Switch over as a kind of means to improve systematic function between user, the transmitting antenna of varying number and delivery plan.Note Meaning, M.Sharif and B.Hassibi, " On the capacity of MIMO broadcast channel with Partial side information, " IEEE Trans.Info.Th., vol.51, p.506522, Feb.2005 and W.Choi, A.Forenza, J.G.Andrews, and R.W.Heath Jr., " Opportunistic

space division Multiple access with beam selection, " IEEE Trans are occurred in, carry on Communications The scheme of the adaptively selected user in multi-user MIMO system is gone out.However, opportunistic during these are open (opportunistic) space division multiplexing access (OSDMA) scheme be designed to be maximized by using multi-user diversity it is total Capacity, and they can only realize the part of theory capacity of dirty paper (dirty paper) code, because incomplete at emitter Interference is eliminated in advance.In DIDO transmission algorithms described herein, inter-user interference is eliminated in advance using block diagonalization.So And, propose that self adaptation sending strategy can apply to any DIDO systems, the type without the need for considering precoding technique.

This patent application describes the extension of the embodiment of the invention described above and earlier application, including but not limited to Lower supplementary features：

1st, can by wireless client device using in earlier application be used for channel estimation training symbol come to self adaptation DIDO Link quality metric in scheme is evaluated.

2nd, as described in earlier application, base station receives the signal characteristic data from client device.Current real In applying mode, signal characteristic data are defined for entering self adaptation the link quality metric of line activating.

3rd, one is prior application described for selecting the mechanism of antenna and number of users, and define throughput distribution.This Outward, the handling capacity of different stage can be dynamically assigned to into different clients as earlier application.The current enforcement of the present invention Mode defines the novel standard related to the selection and throughput distribution.

b、Embodiments of the present invention

The target of the self adaptation DIDO technology for being proposed is by the way that the Radio Resource in time, frequency and space is dynamic Distribute to the different user in system to strengthen the spectrum efficiency or downlink tone spectrum efficiency of every user.The overall self adaptation mark It is mutatis mutandis in while target error rate is met,

improve handling capacity.According to spread state, it is also possible to use the adaptive algorithm via Diversity scheme is improving the link-quality (or coverage rate) of user.The flow chart that Figure 21 shows describes self adaptation DIDO scheme Step.

2102, base station (BS) collects the channel condition information from all users.2104, according to the CSI for being received, Base station calculates link quality metric in time domain/frequency domain/spatial domain.2106, being selected using these link quality metrics will be under The user of being serviced in one transmission, and for the sending mode of each user.Note, sending mode include modulation/coding with And the various combination of DIDO schemes.Finally, user is sent data to via DIDO precodings in 2108, BS.

2102, the base station selected channel condition information (CSI) from all user equipmenies.2104, base station uses should CSI is determining the instantaneous or statistical channel quality of all user equipmenies.In DIDO-OFDM systems, can be in time domain, frequency domain and sky Domain is estimated by channel quality (or link quality metric).Afterwards, 2106, base station is determined most using link quality metric Good user's subset and the sending mode for current propagation state.DIDO sending mode set is combined into DIDO schemes (i.e., Its line options or multiplexing), the combination of modulation/coding scheme (MCS) and array configuration.2108, by using selected user Quantity and sending mode, send data to user equipment.

Can by look-up table (LUT) (look-up table be based on DIDO systems difference communication environments in bit error rate performance and It is pre-calculated) carrying out model selection.Channel quality information is mapped to bit error rate performance by these LUT.In order to build LUT, can evaluate DIDO systems bit error rate performance in different propagation situations according to SNR.Can be seen that from ber curve, can calculate Realize the minimum SNR needed for a certain predeterminated target bit error rate.The SNR requirement definitions are SNR threshold values by we.Afterwards, in difference Propagation situation and for different DIDO sending modes evaluating SNR threshold values, and store it in LUT.For example, can make LUT is built with SER results in Figure 24 and Figure 26.Afterwards, according to the LUT, base station may be

selected the transmission mould for active user Formula, the pattern can improve handling capacity while the predeterminated target bit error rate is met.Finally, base station via DIDO precodings by number According to transmission to selected user.Note, different DIDO mode assignments can be adjusted and DIDO subflows to different time slots, OFDM, with So that self adaptation can be carried out in time domain, frequency domain and spatial domain.

Figure 19-Figure 20 shows a kind of embodiment using the adaptive systems of DIDO.Introduce some new functions Unit is implementing proposed DIDO adaptive algorithms.Specifically, in one embodiment, DIDO configurators 1910 can base In the channel quality information 1912 that user equipment is provided, several functions are performed, including select number of users, DIDO delivery plans (that is, day line options and multiplexing), modulation/coding scheme (MCS) and array configuration.

Subscriber selector unit 1902 is selected and multiple users based on the feedback information by acquired in DIDO configurators 1910 U_1-U_MAssociated data, and each coded modulation unit during the information is provided per multiple coded modulation units 1904.Often Individual coded modulation unit 1904 is encoded and modulated to the information bit of each user, and they are sent to map unit 1906.Input bit is mapped to complex symbol by the map unit 1906, and sends it to precoding unit 1908.Coding is adjusted Unit processed 1904 and map unit 1906 are selected as each user institute using the information for being obtained from DIDO configurators unit 1910 Using modulation/coding scheme type.Described information can be provided by configurator unit 1910 by using feedback unit 1912 Each user channel quality information calculating.DIDO precoding units 1908 are utilized and obtained by DIDO configurators unit 1910 The information for taking is calculating DIDO precoding weights, and the incoming symbol to being obtained from map unit 1906 carries out precoding.By DIDO precoding units 1906 send the data flow after each precoding to OFDM units 1915, and the OFDM units 1915 are calculated IFFT simultaneously adds Cyclic Prefix.D/A units 1916 are sent this information to, the D/A units 1916 carry out digital-to-analogue conversion, and will most Whole analog signal is sent to RF units 1914.The RF units 1914 to intermediate frequency/radio frequency, and are sent to baseband signal raising frequency To transmitting antenna.

The RF units 2008 of each client device receive the signal sent from DIDO transmitter units 1914, by the signal Base band is downconverted to, and the signal after frequency reducing is supplied to into A/D units 2010.Afterwards, the A/D units 2010 by the signal from mould Plan is converted to numeral, and sends it to OFDM units 2013.The OFDM units 2013 remove Cyclic Prefix, and perform FFT, with By signal reports to frequency domain.In cycle of training, OFDM units 2013 transmit output to channel estimating unit 2004, and the channel is estimated Meter unit 2004 calculates in a frequency domain channel estimation.Alternatively, can be in time-domain calculation channel estimation.During the data cycle, OFDM units 2013 transmit output to receiver unit 2002, and the receiver unit 2002 is demodulated to signal/decodes, with Obtain data 2014.The channel estimating unit 2004 sends channel estimation to DIDO feedback generators unit 2006, should DIDO feedback generators unit 2006 can quantify to channel estimation, and send it back to send out via feedback control channel 1912 Penetrate machine.

The DIDO configurators 1910 are usable in the information that base station is obtained, or in a preferred embodiment, additionally make With the output for working in the DIDO feedback generators 2006 (referring to Figure 20) at each user equipment.The DIDO feedback generators 2006 are generated using the other specification similar to estimated SNR at estimated channel conditions 2004 and/or receiver The feedback message of DIDO configurators 1910 will be input to.The DIDO feedback generators 2006 can enter at receiver to information Row compresses, quantifies and/or using some Limited Feedback strategies known in the field.

The DIDO configurators 1910 can use the information recovered from DIDO feedback control channels 1912.DIDO feedback controls Channel is logic or physical control channel, and the channel can be used to the output of DIDO feedback generators 2006 is sent to base from user Stand.Control channel 1912 can be implemented in any number of mode known in the field, and can be that logic or physical control are believed Road.Used as physical channel, it may include the dedicated time slot/frequency band for being assigned to user.It can also be by all users to share with Machine access channel.The control channel can be pre-assigned, or can occupy bit by predetermined way in existing control channel (stealing bits) is creating.

In the following discussion, will measure by using DIDO-OFDM prototypes described in true propagation environment and be obtained The result for taking.These results show that the realizability of potential gain in self adaptation DIDO system.Represent different stage first The performance of DIDO systems, shows that antenna/user quantity can be increased, to realize bigger downlink throughput.Afterwards, description with The relevant DIDO performances in the position of user equipment, demonstrate the need for tracking the channel status of change.Finally, to using diversity technique The performance of DIDO systems is described.

The performance of i, different stage DIDO systems

The performance of different DIDO systems is evaluated using increasing transmitting antenna (N=M, wherein M are number of users). The performance of following system is compared：SISO, DIDO2 × 2, DIDO4 × 4, DIDO6 × 6 and DIDO8 × 8.DIDON×M Refer to that there is the DIDO of N number of transmitting antenna and M user at BS.

Figure 22 shows transmit/receive antenna layout.Transmitting antenna 2201, and user position are arranged with square array configuration Around emission array.In Figure 22, T refers to " transmitting " antenna, and U refers to " user equipment " 2202.

Different antennae subset in 8 yuan of emission arrays is active, and this depends on the N selected by different measuring Value.For every DIDO ranks (N), the true area of maximum that can be targeted to the constraint of the fixed size of 8 element array is selected to carry out The antenna subset of covering.The standard is expected to that the space diversity of given N values can be strengthened.

Figure 23 shows the array configuration of the different DIDO ranks for being adapted to available true area (that is, dotted line).It is square Dashed box have 24 " × 24 " size, corresponding to 450MHz carrier frequencies～λ × λ.

Based on the commentary related to Figure 23 and refer to Figure 22, now define and in relatively following system each system property Energy：

SISO (2301) with T1 and U1

With T1,2 and U1,2 DIDO2 × 2 (2302)

With T1,2,3,4 and U1,2,3,4 DIDO4 × 4 (2303)

With T1,2,3,4,5,6 and U1,2,3,4,5,6 DIDO6 × 6 (2304)

With T1,2,3,4,5,6,7,8 and U1,2,3,4,5,6,7,8 DIDO8 × 8 (2305)

Figure 24 shown in the case of 4-QAM and 1/2FEC (forward error correction) rate, SER, BER, SE in above-mentioned DIDO systems (spectrum efficiency) and goodput performance and the functional relation for launching (TX) power.Observation show that SER and BER performances can be because of N Value increases and declines.The impact is caused by following two phenomenons：For fixed TX power, the input power of DIDO arrays Can be divided between increasing user (or data flow)；Space diversity can be with the number of users in actual DIDO channels Increase and reduce.

As shown in figure 24, in order to compare the relative performance of different stage DIDO systems, target BER is fixed as into 10^{-4}(the value Can be changed according to system), the value corresponds roughly to SER=10^{-2}.We will be referred to as TX corresponding to the TX performance numbers of the target Power threshold (TPT).For any N, if TX power is less than TPT, it will be assumed that can not possibly carry out sending out under DIDO level ns Send, and we need to switch to the other DIDO of even lower level.Additionally, in Figure 24, observable draws, when TX power exceedes for any During the TPT of N values, SE and goodput performance can reach saturation.According to these results, self adaptation sending strategy can be designed to Switch between different stage DIDO, to strengthen SE or goodput for the fixed predeterminated target bit error rate.

Performance under ii, user variable situation

Aiming at for the test, via being emulated in space correlation channel, evaluates the DIDO of different user position Performance.DIDO2 × 2 system is considered with 4QAM and 1/2FEC rates.As shown in figure 25, user 1 is located at the side of emission array (broadside) direction is penetrated, and the position of user 2 penetrates direction and is changed into end-fire (endfire) direction from side.Transmitting antenna interval- λ/2, and it is separated by -2.5 λ with user.

Figure 26 shows the diverse location for user equipment 2, SER and the SE results per user.From the side of emission array Orientation measurement is penetrated, the angle of arrival (AOA) of user equipment is 0 ° to 90 °.Observation draws, as the angular distance of user equipment increases, DIDO performances will be lifted, because DIDO channel memories are in bigger diversity.Additionally, in target SER=10^{-2}Place, AOA2=0 ° There is the gap of 10dB between AOA2=90 ° of both of these case.Imitated for 10 ° of angle spread in the result and Figure 35 True result is consistent.Furthermore, it is noted that in the case of AOA1=AOA2=0 °, coupling effect is there may be between two users Answer (being caused because their antenna is adjoining), this may be such that their performance is different from the simulation result in Figure 35.

Iii, for the preferred situation of DIDO8 × 8

Figure 24 shows the bigger SE of DIDO8 × 8 producing ratio even lower level DIDO, but with higher TX power demands.This point Aiming at for analysis illustrates the presence situation, i.e. DIDO8 × 8 not only at peaks spectrum efficiency (SE) aspect, but also in TX work(Rate demand (or TPT) aspect, surpasses DIDO2 × 2, to realize the peak value SE.

Note, in i.i.d. (ideal) channel, TX power presence～6dB between DIDO8 × 8 and the SE of DIDO2 × 2 Gap.The gap be because of the fact caused by, i.e., DIDO8 × 8 are split TX power between 8 data flows, and Only split between two stream DIDO2 × 2.The result is illustrated via the emulation in Figure 32.

However, in space correlation channel, TPT is that (for example, array direction, customer location, angle expand communication environments characteristic Exhibition) function.For example, Figure 35 show between two different user devices positions low angle extension～15dB between Gap.Similar result is illustrated in the application Figure 26.

Similar to mimo system, when user is located at the end-on direction of TX arrays, the performance of DIDO systems can decline (because lacking Few diversity is caused).The impact can measure by using current DIDO prototypes and observe and draw.Therefore, one kind is illustrated It is that user is placed in the end-on direction relative to DIDO2 × 2 array that DIDO8

× 8 surpass the mode of DIDO2 × 2.In this respect, DIDO8 × 8 have surpassed DIDO2 × 2, because 8- aerial arrays provide higher diversity.

In this analysis, it is contemplated that following system：

System 1：DIDO8 × 8 (8 parallel data streams are sent per time slot) of 4-QAM；

System 2：DIDO2 × 2 (per 4 time slots, once being sent to sending user X and Y) of 64-QAM.For this is System, it is contemplated that four kinds of combinations of TX and RX aerial positions：A) T1, T2U1,2 (end-on direction)；B) T3, T4U3,4 (end-fire side To)；C) T5, T6U5,6 (being separated by～30 ° with end-on direction)；D) T7, T8U7,8 (NLOS (without sighting distance))；

System 3：DIDO8 × 8 of 64-QAM；And

System 4：MISO8 × 1 (per 8 time slots, once being sent to sending user X) of 64-QAM.

For all these situations, using 3/4 FEC rates.

Figure 27 depicts the position of user.

In Figure 28, SER results are shown because different array direction and customer location are between system 2a and 2c A～15dB gap (similar to the simulation result in Figure 35).The first subgraph in a second row shows that SE curves are satisfied The value (that is, corresponding to BER1e-4) of the TX power of sum.It is observed that system 1 than system 2 for relatively low TX power demands (less than～5dB) generates the SE of each bigger user.Spatial multiplexing gain of DIDO8 × 8 on DIDO2 × 2 is additionally, since, DIDO8 × 8 become apparent from compared to the benefit of DIDO2 × 2 for DL (downlink) SE and DL goodputs.Due to The array gain (that is, the MRC with MISO8 × 1) of beam forming, system 4 has lower TX power demands (little than system 1 In 8dB).But system 4 only generates the 1/3 of the SE of each user compared to system 1.System 2 than system 1 poor performance (i.e., For larger TX power demands generate relatively low SE).Finally, system 3 than system 1 for larger TX power demands (～15dB) generate much bigger SE (because larger exponent number (larger order) is modulated).

According to these results, can be inferred that to draw a conclusion：

A kind of channel conditions are confirmed to be DIDO8 × 8 surpasses DIDO2 × 2 (i.e. for relatively low TX power demands are generated Larger SE)；

In the channel conditions, DIDO8 × 8 than DIDO2 × 2 and MISO8 × 1 generate each bigger user SE and DL SE；And

Can by with larger TX power demands (more than～15dB) use for cost high order modulation (i.e. 64-QAM, and It is not 4-QAM) further increasing the performance of DIDO8 × 8.

Iv. there is the DIDO of day line options

Below, we assess received by IEEE journals on Signal Processing 2005 by R.Chen, " the Transmit selection diversity for unitary that R.W.Heath and J.G.Andrews are delivered Described in precoded multiuser spatial multiplexing systems with linear receivers " Antenna Selection Algorithem benefit.We with two users, 4-QAM and 1/2 FEC rates presenting for a specific DIDO system The result of system.Following system is compared in figure 27：

With T1,2 and U1,2 DIDO2 × 2；And

With T1, DIDO3 × 2 of 2,3 and U1,2 use day line options.

Position of transmitting antenna and user device location are identical with Figure 27.

Figure 29 shows that DIDO3 × 2 with day line options can carry compared with DIDO2 × 2 system (not with selection) For the gain of～5dB.Notice that channel is almost static (i.e. without Doppler effect), so selection algorithm is damaged suitable for path Consumption is related to channel space, rather than rapid decay.We should see difference in the situation with high Doppler effect Gain.And, in the particular experiment, it was observed that Antenna Selection Algorithem selects antenna 2 and 3 to be used to send.

Iv. it is used for the SNR threshold values of LUT

[0171] is being selected, we are stated model selection and are realized by LUT.LUT can be by assessing SNR threshold values come quilt

Precomputation with realize in different communication environments be used for DIDO sending modes specific predefined target error rate performance.Under Face, we provides with and does not have the performance of the DIDO systems of day line options and transformable number of users, the performance Can serve as constructing the guidance of LUT.Although Figure 24, Figure 26, Figure 28, Figure 29 with the actual measurement of DIDO prototypes by being obtained, under The figure in face is obtained by emulation.Following BER results are assumed without FEC.

Figure 30 shows the average BER performances of different DIDO pre-coding schemes in independent same distribution channel.Indicate and " do not have Have selection " curve refer to using the situation of BD.In same figure, the performance of day line options (ASel) is for varying number Additional antenna (for the user of varying number) and be illustrated.As can be seen that as the quantity of additional antenna increases, ASel is provided Preferably diversity gain (slope of the BER curve in Yi Gao SNR areas is characterized), generates preferably covering.For example, if we Target BER is fixed to into 10^{-2}(for the actual value of uncoded system), then by the SNR gain of ASel offers with antenna Quantity increases.

Figure 31 shows the number as the extra transmitting antenna in independent same distribution channel for different targets BER The SNR gain of the ASel of the function of amount.As can be seen that only by adding 1 or 2 antenna, ASel generates huge compared with BD SNR gain.In following part, we by only for the situation of 1 or 2 additional antenna by the way that target BER is fixed to into 10^{-2} (for uncoded system) is assessing the performance of ASel.

Figure 32 shows that the conduct for the BD and ASel in independent same distribution channel with 1 and 2 additional antenna is used The SNR threshold values of the function of amount amount (M).It is observed that due to the larger reception SNR demands for large number of user, SNR threshold values increase with M.Note, it will be assumed that for any number of user (uses varying number for fixed total transmission power Transmitting antenna).Additionally, Figure 32 is shown because the gain of day line options is for any amount in independent same distribution channel User for be constant.

Below, the performance of the DIDO systems in spatial correlation channel is we illustrated.We by X.Zhuang, F.W. Vook, K.L. Baum, T.A.Thomas and M.Cudak in September, 2004 in IEEE802.16Broadband " the Channel models for link and system delivered on Wireless Access Working Group COST-259 spatial Channel Models described in level simulations " emulate the channel of each user.We generate is used for The single group of each user.As a kind of case study, it will be assumed that NLOS channels, there is uniform linear array in emitter (ULA), element spacing is 0.5 λ.When 2 custom systems, we are for the first and second users are respectively with arrival The average angle of AOA1 and AOA2 carrys out emulation group.AOA is measured relative to the side surface direction of ULA.When having in systems more than two Individual user, we generate has in scope $[-\varphi_m, \varphi_m]$ in evenly spaced average AOA user group, wherein we Definition

K is the quantity of user, and $\Delta\varphi$ is the angular distance between the average AOA of user.Note angular range $[-\varphi_m, \varphi_m]$ center For 0 °, corresponding to the side of ULA direction is penetrated.Below, we study work with BD and ASel delivery plans and different numbers of users The BER performances of the DIDO systems of the function of the angular distance of (AS) and between user are distributed for channel angle.

Figure 33 is shown for positioned at same angle direction (side i.e. relative to ULA penetrate direction, AOA1=AOA2=0 °) The BER of the average SNR relative to each user of two users with different AS values.As can be seen that as AS increases, BER Performance improvement and close independent same distribution situation.In fact, higher AS statistically generates the character modules in two users The better performance of less covering and BD precoders between formula.

Figure 34 shows the result similar to Figure 33, but has higher angular distance between users.It is contemplated that AOA1= 0 °, AOA2=90 ° (i.e. 90 ° angular distances).Best performance is realized in the case of low AS.In fact, for high angular distance Situation, when angular distance is low, there is less crossover between the feature mode of user.It is interesting that it is observed that for just Mention it is identical the reasons why, the BER performances in low AS are better than independent same distribution channel.

Next, for 10 in different relevant situations^{-2}Target BER, we calculate SNR threshold values.Figure 35 is depicted The SNR threshold values of the function as AS of the different value for the average AOA of user.For low user's angular distance, with rational The reliable transmission of SNR demands (i.e. 18dB) is possible only for the channel being characterized with high AS.On the other hand, when user exists When being spatially separated, less SNR is needed to meet identical target BER.

Figure 36 is shown a case that for the SNR threshold values of 5 users.Definition in (13) is generated has different angles The average AOA of user of the value away from Δ φ.It is observed that for Δ φ=0 ° and AS<15 °, due to the little angular distance between user, BD poor performances, do not meet target BER.For the AS of increase, the SNR demands for meeting fixed target BER reduce.The opposing party Face, for Δ φ=30 °, in low AS minimum SNR demands is obtained, consistent with the result in Figure 35.As AS increases, SNR thresholds Value is saturated to one in independent same distribution channel.Note, Δ φ=30 ° with 5 users are corresponding to [- 60 °, 60 °] AOA scopes, this is typical for the base station in the cellular system with 120 ° of sector elements.

Next, we have studied the performance of the ASel delivery plans in spatial correlation channel.Figure 37 compare for The SNR threshold values of the BD and ASel with 1 and 2 additional antenna of two user situations.We consider the angular distance between user Two kinds of different situations : { AOA1=0 °, AOA2=0 ° } and { AOA1=0 °, AOA2=90 ° }.For BD schemes (i.e. without antenna Select) curve it is identical with Figure 35.It is observed that ASel is generated with 1 and 2 additionally respectively for high AS The SNR gain of the 8dB and 10dB of antenna.With AS reduce, on BD because the gain of ASel is due in MIMO broadcast channels In the free degree quantity reduce and become less.It is interesting that for AS=0 ° (i.e. close to LOS channels) and situation { AOA1=0 °, AOA2=90 ° }, ASel does not provide any gain due to difference in the spatial domain.Figure 38 shows and figure 37 similar results, but during for 5 users.

We calculate the SNR thresholds of the function as number of users (M) in systems for BD and ASel delivery plans Value (assumes 10^{-2}General objectives BER).SNR threshold values correspond to average SNR, so that total transmission power is permanent for any M Fixed.We assume that in

azimuth coverage [- φ_m, φ_m- 60 ° of]=[, 60 °] in each customer group average AOA between maximum Interval.Then, the angular distance between user is Δ φ=120 °/(M-1).

Figure 39 is shown for the SNR threshold values of the BD schemes with difference AS values.It is observed that due between user Big angular distance, for AS=0.1 ° (the i.e. low angular spread) of the user (i.e. K≤20) with relatively small number, obtains most Low SNR demands.However, for M>50, because Δ φ is very little and BD can not be carried out, SNR demands are far longer than 40dB.This Outward, for AS>10 °, SNR threshold values are held nearly constant for any M, and the DIDO systems in spatial correlation channel are close to independent With the performance of distribution channel.

In order to reduce the value of SNR threshold values and improve the performance of DIDO systems, we apply ASel delivery plans.Figure 40 is illustrated For the SNR threshold values in the spatial correlation channel with AS=0.1 ° of BD and ASel with 1 and 2 additional antenna. In order to refer to, we there is also reports that the curve of the independent same distribution situation for figure 32 illustrates.Can see, for less User (i.e. M≤10), due to lacking diversity in DIDO broadcast channels, day line options do not help reduce SNR demands.With Number of users increases, and ASel is benefited from multi-user diversity, generates SNR gain (i.e. for M=20 is 4dB).Additionally, for M ≤ 20, the performance of the ASel with 1 or 2 additional antenna in high spatial correlation channel is identical.

Then we calculate the SNR threshold values of the channel conditions other for two kinds：The AS in AS=5 ° and Figure 42 in Figure 41 =10°.Figure 41 shows due to larger angular spread that ASel is generated and is also used for relatively few number of use compared with Figure 40 The SNR gain at family (i.e. M≤10).As reported in Figure 42, for AS=10 °, SNR threshold values are further reduced, due to ASel's Gain becomes higher.

Finally, we summarize the result for proposing for correlated channels at present.Figure 43 and Figure 44 show and have respectively 1 With 2 additional antennas as the number of users (M) for BD and ASel schemes and the SNR thresholds of the function of angular spread (AS) Value.Note, AS=30 ° of situation actually corresponds to independent same distribution channel, this value of our AS used in figure is used to scheme Shape is

represented.It is observed that, although BD is affected by channel space correlation, and ASel generates almost identical for any AS Performance.Additionally, for AS=0.1 °, due to multi-user diversity, ASel is similar to BD performances for low M, and for big M (i.e. M >=20) more than BD.

Figure 49 compares the performance of different DIDO schemes in terms of SNR threshold values.The DIDO schemes for being considered are：BD、ASel, BD and maximum-ratio combing (MRC) that (BD-ESel) is selected with feature mode.Notice that MRC is sending out without elimination in advance The interference (unlike other methods) penetrated at machine, but larger gain is provided in the case where user is spatially separated.In Figure 49 In, we depict when two users respectively be located at penetrate direction into -30 ° and 30 ° with the side of emission array when, for DIDO N × 2 systems for target BER=10^{-2}SNR threshold values.It is observed that for low AS, MRC schemes are carried compared with other schemes The gain of 3dB has been supplied, because the space channel of user is separated well, the impact very little of the interference between user.Note, The gain of the MRC on DIDO N × 2 is due to array gain.For the AS more than 20 °, QR-ASel schemes exceed other schemes And generate the gain of about 10dB compared with not having selectable BD2 × 2.Arbitrary values of the QR-ASel and BD-ESel for AS There is provided about the same performance.

Described above is the new self adaptation transmission technology for DIDO systems.The method is between DIDO sending modes Dynamic translation strengthens for the handling capacity of fixed target error rate to different users.The property of the DIDO systems of different stage Can be measured under different propagation conditions, it was observed that the huge gain in handling capacity can be used as propagation feelings by dynamic select The DIDO patterns and number of users of the function of condition is realizing.

III. the precompensation of frequency and phase difference

A. background

As described above, wireless communication system transmits information using carrier wave.These carrier waves are typically sine wave, its amplitude And/or phase response is modulated in the information for being sent.The

nominal frequency of sine wave is known as carrier frequency.In order to create this Waveform, emitter synthesizes one or two sine wave, and is created using up-conversion and overlap with designated carrier frequency The signal after modulation on sine wave.This can be realized by directly conversion, wherein, signal is on carrier wave or by multiple liters The frequency conversion stage is from directly modulated.In order to process the waveform, receiver must demodulate received RF signals, and effectively move Except modulation carrier wave.This needs the modulated process that receiver synthesizes one or more sinusoidal signals to be reversed at emitter, it is known that For frequency reducing conversion.Regrettably, the sine wave signal for generating in transmitter and receiver is obtained from different reference oscillators.Not yet There is reference oscillator to create the frequency reference of perfect (perfect) ; In fact, generally there is some deviations with actual frequency.

In a wireless communication system, the difference of the output of the reference oscillator at transmitter and receiver is at receiver Create the phenomenon for being known as carrier frequency shift or simple frequency shift (FS).Substantially, after down conversion, received Signal in have some residue modulation (corresponding to the difference that sends and receives in carrier wave).This create in received signal Distortion, result in higher bit error rate and relatively low handling capacity.

There are the different technologies for processing carrier frequency shift.Most methods estimate the carrier frequency at receiver Skew, then using offset correction of carrier frequency algorithm.Carrier frequency offset estimation algorithm is blindness using following methods (blind) : Offset-QAM (T.Fusco and M.Tanda, " Blind Frequency-offset Estimation for OFDM/ OQAM Systems, " Signal Processing, IEEE Transactions on [referring also to Acoustics, Speech, And Signal Processing, IEEE Transactions on], vol.55, pp.1828-1838,2007) ; Cycle is special Property (E.Serpedin, A.Chevreuil, G. B.Giannakis and P. Loubaton, " Blind channel and carrier frequency offset estimation using periodic modulation precoders," Signal Processing, IEEE Transactions on are [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.48, no.8, pp.2389-2405, Aug.2000) ; Or it is orthogonal Cyclic Prefix (J.J.van de Beek, M.Sandell and P. in frequency division multiplexing (OFDM) structural approach O.Borjesson, "

ML estimation of time and frequency offset in OFDM systems, " Signal Processing, IEEE Transactions on are [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.45, no.7, pp.1800-1805, July1997 ; U.Tureli, H.Liu and M.D.Zoltowski, " OFDM blind carrier offset estimation : ESPRIT,"IEEE Trans.Commun., vol.48, no.9, pp.1459-1461, Sept.2000 ; M.Luise, M.Marselli and R.Reggiannini, " Low-complexity blind carrier frequency recovery for OFDM Signals over frequency-selective radio channels, " IEEE Trans.Commun., vol.50, No.7, pp.1182-1188, July2002).

Alternatively, special training signal can be utilized, including data symbol (P. the H.Moose, " A for repeating technique for orthogonal frequency division multiplexing frequency offset Correction, " IEEE Trans.Commun., vol.42, no.10, pp.2908-2914, Oct.1994) ; Two different Symbol (T.M.Schmidl and D.C.Cox, " Robust frequency and timing synchronization for OFDM, " IEEE Trans.Commun., vol.45, no.12, pp.1613-1621, Dec.1997) ; Or periodically insert Known symbol sebolic addressing (M.Luise and R.Reggiannini, " Carrier frequency acquisition and Tracking for OFDM systems, " IEEE Trans.Commun., vol.44, no.11, pp.1590-1598, Nov.1996).Correction can occur in analog or digital mode.Receiver can also come pre- school using Carrier frequency offset estimation The signal for just being sent is eliminating skew.Due to the sensitivity of multicarrier and ofdm system to frequency shift (FS), carrier frequency shift Correction is extensively studied (J.J.van de Beek, M.Sandell and P. for multicarrier and ofdm system O.Borjesson, " ML estimation of time and frequency offset in OFDM systems, " Signal Processing, IEEE Transactions on are [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.45, no.7, pp.1800-1805, July1997 ; U.Tureli, H.Liu and M.D.Zoltowski, " OFDM blind carrier offset estimation : ESPRIT, " IEEE Trans.Commun., vol.48, no.9, pp.1459-1461, Sept.2000 ; T.M.Schmidl and D.C.Cox, " Robust Frequency and timing synchronization for OFDM, " IEEE Trans.Commun., vol.45, No.12, pp.1613-1621, Dec.1997 ; M.Luise, M.Marselli and R.Reggiannini, " Low-complexity blind carrier frequency recovery for OFDM signals over frequency-selective Radio channels, " IEEE Trans.Commun., vol.50, no.7, pp.1182-1188, July2002).

Frequency offset estimation and correction are for multiple antenna communication or more generally MIMO (multiple-input and multiple-output)

system It is important problem.In mimo systems, transmitting antenna is locked into a frequency reference, and receiver is locked into another frequency Rate benchmark, there is between the transmitter and receiver single skew.Propose several algorithms and ask processing this using training signal Topic (K.Lee and J.Chun, " Frequency-offset estimation for MIMO and OFDM systems using Orthogonal training sequences, " IEEE Trans.Veh.Technol., vol.56, no.1, pp.146- 156, Jan.2007 ; M.Ghogho and A.Swami, " Training design for multipath channel and Frequency offset estimation in MIMO systems, " Signal Processing, IEEE Transactions on are [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.54, no.10, pp.3957-3965, Oct.2006 ; And adaptive tracking C.Oberli and B.Daneshrad, " Maximum likelihood tracking algorithms for MIMOOFDM, " In Communications, 2004IEEE International Conference on, vol.4, June20-24,2004, pp.2468-2472).Prior problem is encountered in mimo systems, wherein, transmitting antenna is not locked into same frequency Rate benchmark, but reception antenna is locked into together.This actually occurs at the up-link that space division multiple access accesses (SDMA) system In, SDMA systems are considered mimo system, and wherein different user corresponds to different transmitting antennas.In this case, frequency The compensation of skew is more complicated.Specifically, frequency shift (FS) creates the interference in the different MIMO streams for being sent.Can make It is corrected (A.Kannan, T.P. Krauss and M.D.Zoltowski, " with complicated Combined estimator and equalization algorithm Separation of cochannel signals under imperfect timing and carrier Synchronization, " IEEE Trans.Veh.Technol., vol.50, no.1, pp.79-96, Jan.2001), and Equilibrium (T.Tang and R.W.Heath, " Joint frequency offset estimation after Frequency offset estimation And interference cancellation for MIMO-OFDM systems [mobile radio], " 2004.VTC2004-Fall.2004IEEE60thVehicular Technology Conference, vol.3, pp.1553- 1557, Sept.26-29,2004 ; X.Dai, " Carrier frequency offset estimation for OFDM/SDMA Systems using consecutive pilots, " IEEE Proceedings-Communications, vol.152, Pp.624-632, Oct.7,2005).A few thing has processed the relevant issues of excess phase shift and tracking error, wherein remaining Remaining phase offset is estimated and compensates after Frequency offset estimation, but this work only accounts for the upper of SDMAOFDMA systems Line link (L Haring, S.Bieder and

A.Czylwik, " Residual carrier and sampling frequency synchronization in multiuser OFDM systems,"2006.VTC2006-Spring.IEEE63rd Vehicular Technology Conference, vol.4, pp.1937-1941,2006).When all transmittings and reception antenna During with different frequency references, there is the situation of most serious in mimo systems.With regard to the only available work of this topic Asymptotic analysis (O.Besson and P.Stoica, " On parameter of the evaluated error in flat fading channel are only processed Estimation of MIMO flat-fading channels with frequency offsets, " Signal Processing, IEEE Transactions on [referring also to Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.51, no.3, pp.602-613, Mar.2003).

When the different transmitting antennas of mimo system do not have identical frequency reference, and reception antenna is independently processed from signal When, situation about having been furtherd investigate occurs.This generation is being known as distributed input and output (DIDO) communication system (in document In also referred to as MIMO broadcast channels) in occur.DIDO systems include with spaced antenna access point, and the antenna is sent out Data streams in parallel (via precoding) is sent to multiple users to strengthen the handling capacity of downlink, is now wirelessly provided using identical Source (i.e. identical time-slot duration and frequency band) is used as conventional SISO systems.DIDO systems are described in detail in S.G.Perlman and T.Cotter, entitled " the System and method for distributed that in July, 2004 submits to In the U.S. Patent application 20060023803 of input-distributed output wireless communications " Propose.There is the mode of many enforcement DIDO precoders.One solution is block diagonalization (BD), in such as documents below Description : Q.H.Spencer, A.L.Swindlehurst and M.Haardt, " Zero-forcing methods for Downlink spatial multiplexing in multiuser MIMO channels, " IEEE Trans.Sig.Proc, vol.52, pp.461-471, Feb.2004 ; K.K.Wong, R.D.Murch and K.B.Letaief, " A Joint-channel diagonalization for multiuser MIMO antenna systems, " IEEE Trans.Wireless Comm., vol.2, pp.773-786, Jul2003 ; L.U.Choi and R.D.Murch, " A transmit preprocessing technique for multiuser MIMO systems using a decomposition Approach, " IEEE Trans.Wireless Comm., vol.3, pp.20-24, Jan2004 ; Z.Shen, J.G.Andrews, R.W.Heath and B.L Evans, " Low complexity user selection algorithms For multiuser MIMO systems with block diagonalization,

" are published in IEEE by acceptance Trans.Sig.Proc, Sep.2005 ; Z.Shen, R.Chen, J.G.Andrews, R.W.Heath and B.L Evans, " Sum Capacity of multiuser MIMO broadcast channels with block diagonalization, " quilt It is committed to IEEE Trans.Wireless Comm., Oct.2005 ; R.Chen, R.W.Heath and J.G.Andrews, " Transmit selection diversity for unitary precoded multiuser spatial Multiplexing systems with linear receivers, " are received to IEEE Trans, on Signal Processing, 2005.

In DIDO systems, send precoding and be used to separate the data flow for different user.When transmitting antenna radio frequency When chain does not share same frequency benchmark, carrier frequency shift result in the Railway Project related to system enforcement.When this generation When, each antenna is effectively sent with slightly different carrier frequency.The integrality of DIDO precoders is this destroys, causes each User is subjected to extra interference.Forth below are several solutions to this problem.One in solution implements In mode, DIDO transmitting antennas share a frequency reference by wired, optical or wireless network.In solution Another embodiment in, one or more users estimate frequency offset difference (relative mistakes in skew between antenna pair It is different) and the information is sent back into emitter.Then emitter precorrection frequency shift (FS) and proceed for DIDO training and Precoder estimates phase place.The embodiment is problematic when feedback channel has delay.Reason is that possible have by trimming process The residual phase error of establishment, the trimming process does not consider subsequent channel estimation.In order to solve this problem, one other Embodiment solves this problem using new frequency shift (FS) and phase estimating device by estimation delay.Based on emulation and logical The actual measurement of DIDO-OFDM prototypes execution is crossed providing result.

The frequency for proposing in this document and phase offset compensation method may be to due to the estimations of the noise at receiver Application condition is sensitive.Therefore, another embodiment proposes the method estimated for the time and frequency shift, in low SNR bars It is also very strong under part.

There are the different methods for performing time and frequency shift estimation.Due to its sensitivity to synchronous error, this Many methods in a little methods are proposed exclusively for OFDM waveforms.

These algorithms are therefore general for single carrier wave and overloading waveform without the structure for typically using OFDM waveforms It is enough.Algorithm described below using known fiducial mark (for example, training data) assisting synchronous technology Among class.Many methods are extension (see P.H.Moose, " the A technique for of the frequency offset estimator of Moose Orthogonal frequency division multiplexing frequency offset correction, " IEEE Trans.Commun., vol.42, no.10, pp.2908-2914, Oct.1994).Moose proposes the instruction using two repetitions Practice signal and obtain frequency shift (FS) using the phase difference between received signal.The method of Moose only can be corrected Fraction (fractional) frequency shift (FS).The extension of the method for Moose proposes (T.M.Schmidl and by Schmidl and Cox D.C.Cox, " Robust frequency and timing synchronization for OFDM, " IEEE Trans.Commun., vol.45, no.12, pp.1613-1621, Dec.1997).Their main innovation is to use one The training symbol of periodic OFDM symbol and other differential coding.The integer that differential coding in second symbol is realized Offset correction.Coulson is considered in T.M.Schmidl and D.C.Cox, " Robust frequency and timing Synchronization for OFDM, " IEEE Trans.Commun., vol.45, no.12, pp.1613-1621, Similar setting described in Dec.1997, and in A.J.Coulson, " Maximum likelihood synchronization for OFDM using a pilot symbol : Analysis, " IEEE J.Select.Areas Commun., vol.19, no.12, pp.2495-2503, Dec.2001 and A.J.Coulson, " Maximum likelihood synchronization for OFDM using a pilot symbol : Algorithms, " IEEE J.Select.Areas Commun., vol.19, no.12, pp.2486-2494, provide algorithm with being discussed in detail for analyzing in Dec.2001.One master The difference wanted is the correlation properties that Coulson is provided using the maximal-length sequence for repeating.He also advises using linear Frequency modulation (chirp) signal, because its constant envelope properties in time domain and frequency domain.Coulson considers actual details But do not estimate including integer.The training signal of multiple repetitions by Minn et.al.in H.Minn, V.K.Bhargava and K.B.Letaief, " A robust timing and frequency synchronization for OFDM systems, " IEEE Trans.Wireless Commun., vol.2, no.4, pp.822-839,

July2003 are considered, but the structure trained It is not optimised.Shi and Serpedin propose some optimalitys (K.Shi that training structure has the idea for forming frame synchronization And E.Serpedin, " Coarse frame and carrier synchronization of OFDM systems：a new Metric and comparison, " IEEE Trans.Wireless Commun., vol.3, no.4, pp.1271-1284, July2004).An embodiment of the invention performs frame synchronization and fractional frequency using the method for Shi and Serpedin Bias estimation.

Many methods in the literature are concentrated in frame synchronization and fractional frequency offset correction.Integer offset correction is used T.M.Schmidl and D.C.Cox, " Robust frequency and timing synchronization for OFDM, " IEEE Trans.Commun., vol.45, no.12, pp.1613-1621, the other training symbol in Dec.1997 is being solved Certainly.For example, Morrelli etc. is in M.Morelli, A.N.D ' Andrea and U.Mengali, " Frequency ambiguity Resolution in OFDM systems, " IEEE Commun.Lett., vol.4, no.4, pp.134-136, Apr.2000 In obtained T.M.Schmidl and D.C.Cox, " Robust frequency and timing synchronization for The OFDM, " modified version of IEEE Trans.Commun., vol.45, no.12, pp.1613-1621, Dec.1997.Using not The interchangeable method of same preamble structure proposes (M.Morelli and U.Mengali, " An by Morelli and Mengali Improved frequency offset estimator for OFDM applications, " IEEE Commun.Lett., Vol.3, no.3, pp.75-77, Mar.1999).This method has used the correlation between the M identical training symbol for repeating Property come by the M factors increase fractional frequency offset estimator scope.This is best linear unbiased estimator and receives most Big skew (there is suitable design), but the timing synchronization not provided.

System describe

An embodiment of the invention uses the frequency eliminated based on the precoding of channel condition information in DIDO systems Rate and phase offset.See Figure 11 and for the associated description above the description of the embodiment.

In an embodiment of the invention, each user is using being equipped with connecing for frequency offset estimator/compensator Receipts machine.Go out as shown in Figure 45, in an embodiment of the invention, including the system of receiver includes multiple RF units 4508th, corresponding multiple

A/D units 4510, be equipped with frequency offset estimator/compensator 4512 receiver and DIDO it is anti- Feedback maker unit 4506.

RF units 4508 receive the signal sent from DIDO transmitter units, and signal down is transformed into into base band, and will drop Signal after frequency is changed is provided to A/D units 4510.Then signal from analog is converted to numeral by A/D units 4510, and by it It is sent to frequency offset estimator/compensator units 4512.Frequency offset estimator/compensator units 4512 estimate frequency shift (FS) And compensating frequency deviation, as described here, then the signal after compensation is sent to into OFDM units 4513.OFDM units 4513 remove Cyclic Prefix and run FFT (FFT) by signal reports to frequency domain.During the training period, OFDM is mono- Unit 4513 sends output to channel estimating unit 4504 to calculate channel estimation in a frequency domain.Alternatively, channel estimation Can be calculated in the time domain.During the data cycle, OFDM units 4513 send output to DIDO receiver units 4502, The DIDO receiver units 4502 are demodulated/decode to obtain data to signal.Channel estimating unit 4504 is by channel estimation DIDO feedback generators unit 4506 is sent to, the DIDO feedback generators unit 4506 can be with quantized channel estimation and via anti- They are sent back emitter by feedback control channel, as shown.

Description to an embodiment of the algorithm for DIDO2 × 2 situation

The embodiment of the algorithm for the frequency/phase migration in DIDO systems is described below.DIDO systems System model start with and have frequency/phase skew in the case of be described.For simplicity, there is provided DIDO2 × 2 are The particular implementation of system.However, the general principle of the present invention can be being carried out in high-order DIDO systems.

With/there is no the DIDO system models of frequency and phase offset

The received signal of DIDO2 × 2 can be write as first user：

$$r_1[t]=h_{11}(w_{11}x_1[t]+w_{21}x_2[t])+h_{12}(w_{12}x_1[t]+w_{22}x_2[t]) \quad (1)$$

And for second user is write as：

$$r_2[t]=h_{21}(w_{11}x_1[t]+w_{21}x_2[t])+h_{22}(w_{12}x_1[t]+w_{22}x_2[t]) \quad (2)$$

Wherein t is discrete time index, h_{mn}And W_{mn}It is respectively the channel between m-th user and n-th transmitting antenna With DIDO precoding weights, x_mIt is for the sending signal of user m.Note, h_{mn}And w_{mn}It is not the function of t, as we assume that Channel is constant on the cycle between training and data is activation.

In the presence of frequency and phase offset, the signal for receiving is represented as

And

Wherein, T_sIt is symbol period ; For n-th transmitting antenna, $\omega_{Tn}=2\ \Pi\ f_{Tn}$; For m-th user, $\omega_{Um}=2\ \Pi f_{Um}$; And f_{Tn}And f_{Um}It is respectively for the practical carrier frequency of n-th transmitting antenna and m-th user (by bias effect). Value t_{mn}Represent in channel h_{mn}On cause the random delay of phase offset.Figure 46 depicts DIDO2 × 2 system model.

For the time, we use defined below :

$\Delta\omega_{mn}=\omega_{Um}-\omega_{Tn}$ (5)

For representing the frequency shift (FS) between m-th user and n-th transmitting antenna.

The description of an embodiment of the invention

The method of an embodiment of the invention is illustrated in Figure 47.The method includes following general step (including sub-step, as shown) : For the cycle of training 4701 of Frequency offset estimation ; For the cycle of training of channel estimation 4702 ; Via the data is activation 4703 for having balanced DIDO precodings.These steps are described in detail in the following.

A () is used for the cycle of training (4701) of Frequency offset estimation

During the first cycle of training, base station will be sent to use from one or more training signals of each transmitting antenna One (4701a) in family.As described here, " user " is wireless client device.When DIDO2 × 2, by The signal that m-th user receives is given by :

Wherein, p_1And p_2It is respectively the training sequence sent from the first and second antennas.

M-th user using any type of frequency offset estimator (i.e. by the convolution of training sequence) and can estimate Shifted by delta ω_{m1}With $\Delta\ \omega_{m2}$.Then, according to these values, user calculates the frequency shift (FS) between two emitting antennas：

$\Delta\omega_T=\Delta\ \omega_{m2}-\Delta\omega_{m1}=\omega_{T1}-\omega_{T2}$ (7)

Finally, the value in (7) is fed back to base station (4701b).

Note, the p in $(6)_1$And p_2Be designed to it is orthogonal, so as to user can estimate $\Delta\ \omega_{m1}$With $\Delta\ \omega_{m2}$.Can Alternatively, in one embodiment, identical training sequence is used in two continuous time slots, and user therefrom estimates skew. Additionally, the estimation of the skew in order to improve (7), the calculating of same as described above for DIDO systems all users (no Only for m-th user) for can be done, last estimation can be (after the weighting) of the value obtained from all users Mean value.However, this solution needs more calculating times and feedback quantity.Finally, the renewal of Frequency offset estimation There are when frequency shift (FS) is changed over just needs.Therefore, according to the step of the stability of the clock at emitter, algorithm 4701 Can be performed for a long time (i.e. for each data is activation) so that above-mentioned feedback is reduced.

B () is used for the cycle of training (4702) of channel estimation

During the second cycle of training, base station first from m-th user or from multiple users or must have the value in (7) Frequency shift (FS) feedback.Value in (7) be used to pre-compensate for the frequency shift (FS) in transmitting terminal.Then, base station is by training data It is sent to all users to come for channel estimation (4702a).

For DIDO2 × 2 system, the signal received at first user is given by：

And at second user：

Wherein,And Δ t is random or known between first transmission and the second transmission of base station Postpone.Additionally, p_1And p_2It is respectively the training sequence sent from the first and second antennas of user's frequency shift (FS) and channel estimation.

Note, precompensation is only applied in this embodiment the second antenna.

Launch (8), we obtain

It is similarly for second user：

Wherein,

In receiving terminal, user is by using training sequence p_1And p_2Carry out compensating frequency offset residue.Then, user is by instruction Practice vector channel to be estimated (4702b)：

These channels or channel condition information (CSI) in (12) are fed back to base station (4702b), base station such as lower face DIDO precoders are calculated described in point.

C () has the DIDO precodings (4703) of precompensation

Base station receives the channel condition information (CSI) in (12) and by block diagonalization (BD) calculates precoding from user Weight (4703a) so that

Wherein, vector h_1It is defined in (12), and $w_m=[w_{m1}, w_{m2}]$.Note, the present invention for proposing in the disclosure Can be used in any other DIDO method for precoding in addition to BD.Also mended in advance by using the estimation in (7) base station Frequency shift (FS) is repaid, and delay (the Δ t between sending and being currently transmitted is trained by estimating $second_0$) pre-compensating for phase offset (4703a).Finally, base station transmits data to user (4703b) via DIDO precoders.

After transmission process, the signal received at user 1 is given by：

Wherein,Using attribute (13), we obtain

Similarly, for user 2, we obtain：

Launch (16)：

Wherein,

Finally, user calculates frequency offset residue and channel estimation carrys out demodulated data stream $x_1[t]$ and $x_2[t]$(4703c)。

It is generalized to DIDON × M

In the portion, the technology for describing before is generalized to the DIDO systems with N number of transmitting antenna and M user System.

I. the cycle of training of user's Frequency offset estimation

During the first cycle of training, due to the letter that the training sequence that sends from N number of antenna is received by m-th user Number it is given by :

Wherein, p_nIt is the training sequence from n-th antenna transmission.

Estimating shifted by delta ω_{mn}Afterwards,M-th user calculate first and n-th transmitting antenna it Between frequency shift (FS) :

$\Delta\omega_{T,1n}=\Delta\omega_{mn}-\Delta\omega_{m1}=\omega_{T1}-\omega_{Tn}$ (19)

Finally, the value in (19) is fed back to base station.

Ii. it is used for the cycle of training of channel estimation

During the second cycle of training, base station is obtained with (19) first from m-th user or from multiple users The frequency shift (FS) feedback of value.Value in (19) be used to pre-compensate for the frequency shift (FS) in transmitting terminal.Then, number will be trained in base station Come for channel estimation according to all users are sent to.

For DIDO N × M systems, the signal received at m-th user is given by :

Wherein,And Δ t is the first and second transmissions of base station Between random or known delay.Additionally, P_nIt is the training sequence from n-th antenna transmission for frequency shift (FS) and channel estimation Row.

In receiving side, user is by using training sequence P_nCarry out compensating frequency offset residue.Then, each user m passes through Trained vector channel is being estimated :

And base station is fed back to, base station calculates DIDO precoders as described in following part.

Iii. there is the DIDO precodings of precompensation

Base station receives the channel condition information (CSI) in (12) and by block diagonalization (BD) calculates precoding from user Weight so that

Wherein, vector h_mIt is defined in (21), and $w_m=[w_{m1}, w_{m2}\ldots, w_{mN}]$.Base station is also by using in (19) Estimation pre-compensating for frequency shift (FS), and by estimating that the second training sends and be currently transmitted between delay (Δt_0) mending in advance Repay phase offset.Finally, base station transmits data to user via DIDO precoders.

After transmission process, the signal received at user i is given by：

Wherein,Using attribute (22), we obtain：

Finally, user calculates frequency offset residue and channel estimation carrys out demodulated data stream $x_i[t]$。

As a result

Figure 48 is shown with the SER results with DIDO2 × 2 system with frequency shift (FS).Can see, be proposed Method completely eliminate frequency/phase skew, generate and have skew system identical SER.

Next, we assess fluctuation of the proposed compensation method for frequency offset error and/or real time offset Sensitivity.Therefore, we are rewritten as (14)：

Wherein, ε represents evaluated error and/or the change of the frequency shift (FS) between training and data is activation.Note, the effect of ε Fruit is the orthogonal property destroyed in (13), so that the distracter in (14) and (16) is no at emitter being disappeared in advance completely Remove.Because so, SER performances are reduced with the ε values of increase.

Figure 48 shows the SER performances of the frequency offset compensation method for different ∈ values.These results assume T_s= 0.3ms (i.e. with the signal of 3KHz bandwidth).It is observed that for ε=0.001Hz (or less), SER performances are inclined with no The situation of shifting is similar.

F. it is used for the description of an embodiment of the algorithm that the time and frequency shift is estimated

Below, we describe and perform the other embodiment (4701b in Figure 47) that the time and frequency shift is estimated.Examine The transmission signal structure of worry in H.Minn, V.K.Bhargava and K.B.Letaief, " A robust

timing and Frequency synchronization for OFDM systems, " IEEE Trans.Wireless Commun., Propose in vol.2, no.4, pp.822-839, July2003, in K.Shi and E.Serpedin, " Coarse frame and carrier synchronization of OFDM systems：A new metric and comparison, " IEEE Trans.Wireless Commun., vol.3, no.4, pp.1271-1284, are studied in detail in July2004.Generally have The sequence of good association attributes is used for training.For example, for our system, Chu sequences are used, and Chu sequences such as exist D.Chu, " Polyphase codes with good periodic correlation properties (corresp.), " IEEE Trans.Inform.Theory, vol.18, no.4, pp.531-532, are described in July1972.These sequences have Interesting attribute, i.e., they are with perfect circulation correlation.Allow L_{cp}Represent the length of Cyclic Prefix, N_tRepresent component training sequence Length.So that $N_t=M_t$, wherein M_tIt is the length of training sequence.Under these assumptions, the symbol sequence for the beginning for being sent Row can be written to：

$s[n]=t[n-N_t]$ for $n=-1 \ldots ,-L_{cp}$

S [n]=t [n] for n=0 ..., N_t-1

$s[n]=t[n-N_t]$ for $n=N_t \ldots, 2N_t-1$

$s[n]=-t[n-2N_t]$ for $n=2N_t \ldots, 3N_t-1$

S [n]=t [n-3N_t] for $n=3N_t \ldots, 4N_t-1$

Noting the structure of the training signal can be scaled up to other length, but block structure is repeated.For example, in order that with 16 training signals, it is contemplated that a kind of structure, for example：

[CP, B, B ,-B, B, B, B ,-B, B ,-B ,-B, B ,-B, B, B ,-B, B ,].

By using the structure, and make $N_t=4M_t$, all algorithms that will be described can without modification in the case of be made With.Effectively, our repetition training sequences.This is particularly useful in the case of suitable training signal possibility is disabled.

After the filtering and down sample matched to symbol rate, it is considered to received signal below：

Wherein ε is unknown discrete-time frequency skew, and Δ is unknown vertical shift, and h [1] is unknown discrete time letter Road coefficient, and

v [n] is additional noise.Key idea in order to explain following part, ignores the presence of additional noise.

I. rough frame synchronization

The purpose of rough frame synchronization is to solve unknown vertical shift Δ.Let us is made defined below：

$r_1[n]$ ：=[r[n],r[n+1],...,r[n+N_t-1]]^T,

$r_2[n]$ ：=[r[n+N_t],r[n+1+N_t] ..., r [n+2N_t-1]]^T,

$r_3[n]$ ：=[r[n+2N_t], r [n+1+2N_t] ..., r [n+3N_t-1]]^T,

$r_4[n]$ ：=[r[n+3N_t], r [n+1+3N_t] ..., r [n+4N_t-1]]^T,

The rough frame synchronization algorithm for being proposed from K.Shi and E.Serpedin, " Coarse frame and carrier synchronization of OFDM systems：A new metric and comparison, " IEEE Trans.Wireless Commun., vol.3, no.4, pp.1271-1284, the algorithm in July2004 takes a hint, according to Maximum-likelihood criterion is obtained.

The improved rough frame synchronization of method 1-：Rough frame synchronization estimator solves following optimization：

Wherein,

So that the signal being corrected is defined as：

Little inceptive impulse that other correction term is used to compensate in channel simultaneously can be conditioned based on application.This volume Outer delay will be included in the channel afterwards.

Ii. fractional frequency offset correction

Fractional frequency offset is corrected after rough frame synchronization block.

The improved fractional frequency offset correction of method 2-：Fractional frequency offset is following solution：

This is known as fractional frequency offset, because algorithm only can be with correcting offset

This problem will be solved in next part.Fine frequency offset correction signal is allowed to be defined as :

Note, method 1 and 2 is for preferable K.Shi, the E.Serpedin of working in frequency selective channel, " Coarse frame and carrier synchronization of OFDM systems : a new metric and Comparison, " IEEE Trans.Wireless Commun., vol.3's, no.4, PP.1271-1284, July2004 changes Enter.Here one especially innovation be the use of r recited above andUse improve former estimator because it Have ignored the sampling being affected because internal symbol is disturbed.

Iii. integer frequency deviation correction

In order to correct integer frequency deviation, it is necessary to write one for received after fine frequency offset correction The equivalent system model of signal.The timing error of reservation is absorbed in channel, does not have noisy received signal to have Following structure :

Wherein n=0,1 ..., $4N_t$-1.Integer frequency deviation is k, and unknown equivalent channels are g [l].

The improved integer frequency deviation correction of method 3- : Integer frequency deviation is following solution :

Wherein :

r=D[k]Sg

These give the estimation of total frequency shift (FS) :

In fact, method 3 has very high complexity.In order to reduce complexity, following observation can be made.First, take advantage of Product S $(S^*S)^{-1}$S can be expected to calculate.Regrettably, this still leaves sizable matrix multiplication.Alternatively using tool The observation of the training sequence for having proposed, $S^*S \approx I$.The method that this generates the support type of following reduction.

The improved integer frequency deviation correction of method 4- low-complexity :

The integer frequency deviation estimator of low-complexity is solved

Iv. result

In the portion, we compare the performance of the different estimators for being proposed.

First, in Figure 50, we compare the amount of the expense required for every kind of method.Notice that two kinds of new methods will be opened Pin reduces 10 times to 20 times.For the performance of relatively more different estimators, MonteCarlo experiments are performed.What is considered sets It is to send waveform from our common NVIS of the construction of the linear modulation with 3K symbols symbol rate per second to put, and is corresponded to The pass band width of 3KHz, and the cosine impulse shaping for rising.For each Monte Carlo is realized, frequency shift (FS) from [- f_{max}, f_{max}] on be uniformly distributed and generate.

With f_{max}The little frequency shift (FS) and the emulation without integer offset correction of=2Hz is illustrated in Figure 51.Can be with Find out from the Performance comparision, with N_t/ M_t=1 performance slightly degrades from original estimator, although substantially reduce expense. With N_t/M_t=4 performance is more preferable, almost 10dB.Due to the error in integer bias estimation, all curves are in low SNR Point experienced complications.Little error in integer skew can create big frequency error and big splicing square error.It is whole Number offset correction can be turned off to improvement performance in little skew.

In the presence of multi-path channel, the performance of frequency offset estimator is typically reduced.However, in Figure 52, Close integer offset estimator and present extraordinary performance.Therefore, in multi-path channel, in the rough correction for performing robust Improved fine correction algorithm afterwards is prior.Note, with N_t/M_t=4 offset behavior is good in the case of multipath Much.

Embodiments of the present invention can include various steps suggested above.The step can be executable with machine The mode of instruction realizes that the instruction causes universal or special computing device particular step.For example, base described above Stand/AP and customer set up in various assemblies may be implemented as on universal or special processor perform software.In order to keep away Exempt from the various known personal computer of the parties concerned for obscuring the present

invention, computer storage, hard disk, input unit etc. Component is saved from figure.

Alternatively, in one embodiment, various functions module and correlation step herein shown can pass through Specialized hardware components (such as special IC (ASIC)) of the hardwired logic comprising execution step or by program calculation Any combination of thermomechanical components and custom hardware components and be performed.

In one embodiment, the particular module of coding as described above, modulation and signal processing logic 903 can To be carried out on programmable digital signal processor (DSP) (or DSP groups), the DSP for example uses Texas Instrument (Texas The DSP (for example, TMS320C6000, TMS320C5000 etc.) of TMS320x frameworks Instruments).In this embodiment DSP can be embedded in the package card of personal computer, for example, pci card.Certainly, in the general principle for meeting the present invention In the case of, a variety of DSP architectures can be used.

The various parts of the present invention can also be provided for storing the machine readable media of machine-executable instruction.Machine Device computer-readable recording medium can include but is not limited to flash memory, CD, CD-ROM, DVD ROM, RAM, EPROM, EEPROM, magnetic card or light Learn card, communication media or be suitable to store the other types of machine readable media of e-command.For example, the present invention can be downloaded For computer program, the computer program being included in the data-signal of carrier wave or other communication medias by way of via Communication link (for example, modem or network connection) is sent to requesting computer from remote computer (such as server) (such as client).

Throughout described above, for purposes of explanation, many specific details are proposed to provide the complete of system and method Foliation solution.It will be apparent, however, to one skilled in the art that system and method can be without these certain details It is implemented in the case of some in section.Therefore, the scope of the present invention and essence should be judged according to claims.

Additionally, in the foregoing written description, many documents are cited to provide being more completely understood for the present invention.It is

all these to draw During bibliography is by reference to being integrated into the application.

Claims (24)

Hide Dependent

1. a kind of for compensating frequency and phase offset in the multiaerial system MAS MU-MAS sent with multi-user MU Method, the method includes ：

At each transceiver in multiple transceivers of base station, send or receive more than first training signal and analyze every Individual training signal, to obtain frequency and phase offset compensation at the transceiver ；

Between one in each transceiver and multiple wireless client devices of the base station or each wireless client device Link on send more than second training signal, and analyze each training signal with the base station obtain channel characteristics number According to ；

Multiple MU-MAS precoders weights are calculated based on the channel characteristics data and usage frequency and phase offset compensation, The MU-MAS precoder weights are calculated to the interference between pre- elimination user ；

Data are carried out using the MU-MAS precoders weight precoding generate for the base station each receive and dispatch Data-signal after the precoding of letter machine ； And

Data-signal after the precoding is sent to into each respective client's dress by each transceiver of the base station Put.

2. method according to claim 1, the method also includes ：

More than second training signal is sent to institute by each wireless client device from the plurality of wireless client device Each transceiver of base station is stated, the base station analyzes each training signal to generate channel characteristics data.

3. method according to claim 1, the method also includes ：

Precoding is carried out using the MU-MAS precoders weight to the 3rd many training signals to generate for the base Training signal after the precoding of each transceiver stood ;

The training signal after the precoding is sent to into the plurality of wireless client dress from each transceiver of the base station Each wireless client device in putting, it is special to generate channel that each customer set up in the customer set up analyzes each training signal Data and frequency offset compensation data are levied, and the data-signal after the precoding is demodulated using these data.

4. method according to claim 1, wherein the base station is that the wireless client device is coupled to into wide area network Access point.

5. method according to claim 1, one of them centralized transceiver unit is estimated between all transceivers Frequency shift (FS) and pre-compensate for the skew, or the transceiver of the base station by wired, optical or wireless network come Shared frequencies benchmark.

6. method according to claim 1, wherein more than first training signal is sent to reduce expense for a long time.

7. method according to claim 1, wherein the transceiver estimates the rate of change of frequency shift (FS) and determines training Turnover rate.

8. method according to claim 1, wherein data precoding is performed using block diagonalization BD technologies.

9. method according to claim 1, wherein the MU-MAS systems are distributed input distributed output DIDO leading to Letter system, and wherein described MU-MAS precoders weight is DIDO precoder weights.

10. a kind of uneven for compensating inphase quadrature I/Q in the multiaerial system MAS MU-MAS sent with multi-user MU The method of weighing apparatus, the method includes :

Chain between one in each transceiver and multiple wireless client devices of base station or each wireless client device Multiple training signals are sent on road, and analyzes each training signal to obtain channel characteristics data in the base station ;

Multiple MU-MAS precoders weights are calculated based on the channel characteristics data, the MU-MAS precoders weight is counted Calculate to eliminate the interference brought due to I/Q gains and unbalance in phase and/or inter-user interference in advance ;

Data are carried out using the MU-MAS precoders weight precoding generate for the base station each receive and dispatch Data-signal after the precoding of letter machine ; And

Data-signal after the precoding is sent to into each respective client's dress by each transceiver of the base station Put.

11. methods according to claim 10, the method also includes :

The plurality of training signal is sent to the base station by each wireless client device from multiple wireless client devices Analyze each training signal to generate channel characteristics data in each transceiver, the base station.

12. methods according to claim 10, the method also includes :

Each wireless client device being sent to training signal from each transceiver of base station in multiple wireless client devices, Each customer set up in the customer set up analyzes each training signal to generate channel characteristics data, and receives in the base station The channel characteristics data.

13. methods according to claim 10, wherein the base station is that the wireless client device is coupled to into wide area network Access point.

14. methods according to claim 10, the method also includes :

Using zero ZF receivers, least mean-square error MMSE receivers or maximum likelihood ML receivers are forced come in each user's dress The place's of putting demodulated data stream is suppressing residual interference.

15. methods according to claim 10, wherein performing precoding using block diagonalization BD technologies.

16. methods according to claim 10, wherein the MU-MAS systems are distributed input distributed output DIDO Communication system, and wherein described MU-MAS precoders weight is DIDO precoder weights.

17. methods according to claim 16, wherein the precoder weight be calculated to eliminate inter-user interference and It is not intercarrier interference, and wherein described wireless client device includes having the wave filter for eliminating the ICI to connect Receipts machine.

A kind of 18. methods of the communication characteristic of multiaerial system MASMU-MAS that for dynamically adapting there is multi-user MU to send, The method includes：

Chain between one in each transceiver and multiple wireless client devices of base station or each wireless client device Multiple training signals are sent on road, and analyzes each training signal to obtain channel characteristics data in the base station；

Instantaneous or statistical channel quality, the i.e. link for the wireless client device is determined using the channel characteristics data Quality metric；

The subset and MU-MAS sending modes of user are determined based on the link quality metric；

Multiple MU-MAS precoders weights are calculated based on the channel characteristics data；

Data are carried out using the MU-MAS precoders weight precoding generate for the base station each receive and dispatch Data-signal after the precoding of letter machine；And

Data-signal after the precoding is sent in selected subset by each transceiver of the base station Each respective customer set up.

19. methods according to claim 18, the method includes：

The plurality of training signal of each wireless client device in the plurality of wireless client device is sent to into institute Each transceiver of base station is stated, the base station analyzes each training signal to generate channel characteristics data.

20. methods according to claim 18, the method includes：

Each wireless client device being sent to training signal from each transceiver of base station in multiple wireless client devices, Each customer set up in the customer set up analyzes each training signal to generate channel characteristics data, and connects in the base station Receive the channel characteristics data.

21. methods according to claim 18, wherein the MU-MAS sending modes include transceiver selection/diversity or The various combination of multiplexing, modulation/coding scheme MCS and array configuration/geometry.

22. methods according to claim 18, wherein the link quality metric is in time domain, frequency domain and/or spatial domain It is estimated.

23. methods according to claim 18, wherein the link quality metric is included at the customer set up being connect The signal to noise ratio snr of the signal for receiving.

24. methods according to claim 18, wherein the MU-MAS systems are distributed input distributed output DIDO Communication system, wherein the MU-MAS sending modes are based on the DIDO sending modes of the link quality metric, and wherein The MU-MAS precoders weight is DIDO precoder weights.

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Priority And Related Applications

Parent Applications (1)

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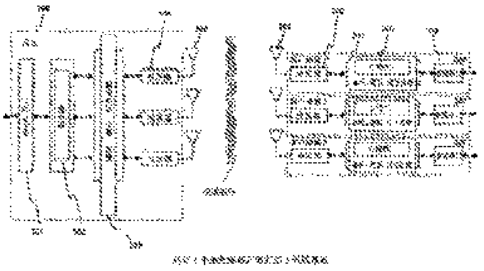
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(54)发明名称

补偿MU-MAS通信、动态适应MU-MAS的方法

(57)摘要

描述了补偿MU-MAS通信、动态适应MU-MAS的方法。该补偿MU-MAS通信的方法包括:从基站的每个天线将训练信号发送到一个或每个无线客户装置,该一个或每个客户装置分析训练信号以生成频率偏移补偿数据并在基站接收频率偏移补偿数据;基于频率偏移补偿数据计算MU-MAS预编码器权重以预消除发射机处的频率偏移;在每个天线与一个或每个无线客户装置之间的链路上发送训练信号并分析训练信号以在基站处获得信道特征数据;基于信道特征数据来计算MU-MAS预编码器权重来预消除频率和相位偏移和/或用户间的干扰;使用MU-MAS预编码器权重来对预编码数据以生成预编码后的数据信号;将预编码后的数据信号通过天线发送到各自的客户装置。



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1.一种用于在具有多用户MU发送的多天线系统MAS MU-MAS中补偿频率和相位偏移的方法，该方法包括：

在基站的多个收发信机中的每个收发信机处，发送或接收第一多个训练信号并分析每个训练信号，以在所述收发信机处获得频率和相位偏移补偿；

在所述基站的每个收发信机与多个无线客户装置中的一个或每个无线客户装置之间的链路上发送第二多个训练信号，并分析每个训练信号以在所述基站处获得信道特征数据；

基于所述信道特征数据并使用频率和相位偏移补偿来计算多个MU-MAS预编码器权重，该MU-MAS预编码器权重被计算来预消除用户间的干扰；

使用所述MU-MAS预编码器权重来对数据进行预编码以生成针对所述基站的每个收发信机的预编码后的数据信号；以及

将所述预编码后的数据信号通过所述基站的每个收发信机发送到每个各自的客户装置。

2.根据权利要求1所述的方法，该方法还包括：

从所述多个无线客户装置中的每个无线客户装置将所述第二多个训练信号发送到所述基站的每个收发信机，该基站分析每个训练信号以生成信道特征数据。

3.根据权利要求1所述的方法，该方法还包括：

使用所述MU-MAS预编码器权重来对第三多个训练信号进行预编码以生成针对所述基站的每个收发信机的预编码后的训练信号；

从所述基站的每个收发信机将所述预编码后的训练信号发送到所述多个无线客户装置中的每个无线客户装置，该客户装置中的每个客户装置分析每个训练信号以生成信道特征数据和频率偏移补偿数据，并使用这些数据来解调所述预编码后的数据信号。

4.根据权利要求1所述的方法，其中所述基站是将所述无线客户装置耦合到广域网的接入点。

5.根据权利要求1所述的方法，其中一个集中式收发信机单元估计所有收发信机之间的频率偏移并预补偿所述偏移，或者所述基站的收发信机通过有线、光学的或无线网络来共享频率基准。

6.根据权利要求1所述的方法，其中所述第一多个训练信号长期被发送以降低开销。

7.根据权利要求1所述的方法，其中所述收发信机估计频率偏移的改变率并决定训练的更新率。

8.根据权利要求1所述的方法，其中数据预编码使用块对角化BD技术而被执行。

9.根据权利要求1所述的方法，其中所述MU-MAS系统是分布式输入分布式输出DIDO通信系统，以及其中所述MU-MAS预编码器权重是DIDO预编码器权重。

10.一种用于在具有多用户MU发送的多天线系统MAS MU-MAS中补偿同相正交I/Q不平衡的方法，该方法包括：

在基站的每个收发信机与多个无线客户装置中的一个或每个无线客户装置之间的链路上发送多个训练信号，并分析每个训练信号以在所述基站获得信道特征数据；

基于所述信道特征数据来计算多个MU-MAS预编码器权重，该MU-MAS预编码器权重被计算来预消除由于I/Q增益和相位不平衡和/或用户间干扰带来的干扰；

使用所述MU-MAS预编码器权重来对数据进行预编码以生成针对所述基站的每个收发信机的预编码后的数据信号；以及

将所述预编码后的数据信号通过所述基站的每个收发信机发送到每个各自的客户装置。

11.根据权利要求10所述的方法，该方法还包括：

从多个无线客户装置中的每个无线客户装置将所述多个训练信号发送到所述基站的每个收发信机，所述基站分析每个训练信号以生成信道特征数据。

12.根据权利要求10所述的方法，该方法还包括：

从基站的每个收发信机将训练信号发送到多个无线客户装置中的每个无线客户装置，该客户装置中的每个客户装置分析每个训练信号以生成信道特征数据，并在所述基站接收所述信道特征数据。

13.根据权利要求10所述的方法，其中所述基站是将所述无线客户装置耦合到广域网的接入点。

14.根据权利要求10所述的方法，该方法还包括：

使用零强制ZF接收机、最小均方误差MMSE接收机或最大似然ML接收机来在每个用户装置处解调数据流以抑制剩余干扰。

15.根据权利要求10所述的方法，其中使用块对角化BD技术来执行预编码。

16.根据权利要求10所述的方法，其中所述MU-MAS系统是分布式输入分布式输出DIDO通信系统，以及其中所述MU-MAS预编码器权重是DIDO预编码器权重。

17.根据权利要求16所述的方法，其中所述预编码器权重被计算来消除用户间干扰而不是载波间干扰ICI，以及其中所述无线客户装置包括具有用于消除所述ICI的滤波器的接收机。

18.一种用于动态适应具有多用户MU发送的多天线系统MASMU-MAS的通信特性的方法，该方法包括：

在基站的每个收发信机与多个无线客户装置中的一个或每个无线客户装置之间的链路上发送多个训练信号，并分析每个训练信号以在所述基站处获得信道特征数据；

使用所述信道特征数据来确定针对所述无线客户装置的瞬时或统计信道质量，即链路质量度量；

基于所述链路质量度量来确定用户的子集和MU-MAS发送模式；

基于所述信道特征数据来计算多个MU-MAS预编码器权重；

使用所述MU-MAS预编码器权重来对数据进行预编码以生成针对所述基站的每个收发信机的预编码后的数据信号；以及

将所述预编码后的数据信号通过所述基站的每个收发信机发送到所选择的子集内的每个各自的客户装置。

19.根据权利要求18所述的方法，该方法包括：

将来自所述多个无线客户装置中的每个无线客户装置的所述多个训练信号发送到所述基站的每个收发信机，所述基站分析每个训练信号以生成信道特征数据。

20.根据权利要求18所述的方法，该方法包括：

从基站的每个收发信机将训练信号发送到多个无线客户装置中的每个无线客户装置，

该客户装置中的每个客户装置分析每个训练信号以生成信道特征数据，并在所述基站处接收所述信道特征数据。

21. 根据权利要求18所述的方法，其中所述MU-MAS发送模式包括收发信机选择/分集或复用、调制/编码方案MCS以及阵列配置/几何结构的不同组合。

22. 根据权利要求18所述的方法，其中所述链路质量度量在时域、频域和/或空间域中被估计。

23. 根据权利要求18所述的方法，其中所述链路质量度量包括在所述客户装置处所接收到的信号的信噪比SNR。

24. 根据权利要求18所述的方法，其中所述MU-MAS系统是分布式输入分布式输出DIDO通信系统，其中所述MU-MAS发送模式是基于所述链路质量度量的DIDO发送模式，以及其中所述MU-MAS预编码器权重是DIDO预编码器权重。

补偿MU-MAS通信、动态适应MU-MAS的方法

[0001] 本申请是申请日为2008年08月20日、申请号为200880102933.4、名称为"分布式输入分布式输出无线通信的系统和方法"的发明专利申请的分案申请。

[0002] 优先权要求

[0003] 本申请为2004年7月30提交的申请NO.10/902,978的继续申请。

技术领域

[0004] 本发明通常涉及通信系统领域。特别地，本发明涉及用于使用时空编码技术的分布式输入分布式输出的无线通信的系统和方法。

背景技术

[0005] 通信信号的时空编码

[0006] 空间多工和时空编码是无线技术中已知的较新的发展。由于有几个天线用在每个终端，所以一种特殊类型的时空编码被称为"多重输入多重输出"(MIMO)。通过使用多个天线来发送和接收，多个独立的无线电波可以在相同的频率范围内同时传送。下面的文章提供了MIMO的概述。

[0007] IEEE成员David Gesbert，IEEE会员Mansoor Shafi、IEEE成员Da-shan Shiu，、IEEE成员Peter J.Smith和IEEE高级会员Ayman Naguib的IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS,VOL.21,NO.3,APRIL2003:"From theory to Practice:An Overview of MIMO Space-Time Coded Wireless Systems"。

[0008] IEEE成员David Gesbert、IEEE成员Helmut Bolcskei、Dhananjay A.Gore和IEEE会员Arogyaswami J.Paulraj的IEEE TRANSCTIONS ON COMMUNICATIONS,VOL.50,NO.12,DECEMBER2000:"Outdoor MIMO Wireless Channels:Models and Performance Prediction"。

[0009] 基本上，MIMO技术是基于用于在公共频带内产生并列空间数据流的空间分布式天线的应用。无线电波以这样的方式传播，从而可以在接收器分离和解调单个信号，即使它们在相同的频带内传输，这能造成多个统计学意义上独立(也就是有效地分离)的通信信道。因此，和努力抑制多径信号的标准无线通信系统相比(即，同一频率的多个时延信号，且振幅和相位存在修改)，MIMO可以依赖于非相关或弱相关的多径信号，在给定的频带内实现较高的吞吐率和改善的信噪比。实例表明，在功率与信噪比(SNR)相当的条件下，MIMO技术实现了更高的吞吐量(throughput)，而传统的非MIMO系统仅能够实现较低的吞吐量。高通公司(高通是最大的无线技术供应商)网站http://www.cdmatech.com/products/what_mimo_delivers.jsp上标题为"What MIMO Delivers"的页面上描述了这一功能："MIMO is the only multiple antenna technique that increases spectral capacity by delivering two or more times the peak data rate of a system per channel or per Mhz of spectrum.To be more specific,for wireless LAN or Wi-Fi® applications QUALCOMM's fourth generation MIMO technology delivers speeds

of315Mbps in36Mhz of spectrum or8.8Mbps/Mhz.Compare this to the peak capacity of802.11a/g(even with beam-forming or diversity techniques)which delivers only54Mbps in17Mhz of spectrum or3.18Mbps/Mhz"。

[0010] 通常上,由于几个原因,MIMO系统面对着每个装置少于10个天线的实际性限制(因此网络中的改善少于10×吞吐率):

[0011] 1.物理限制:给定装置上的MIMO天线之间必须具有足够的间隔,从而每个都接收统计独立的信号。尽管即使在波长分数的天线间隔时仍然可以看到MIMO吞吐量的改善,但当天线更加接近时效率迅速恶化,这导致了较低的MIMO吞吐量的增器。

[0012] 参见例如以下参考文献:

[0013] [1]D.-S.Shiu,G.J.Foschini,M.J.Gans,and J.M.Kahn,"Fading correlation and its effect on the capacity of multielement antenna systems,"IEEE Trans.Comm.,vol.48,no.3,pp.502-513,Mar.2000.

[0014] [2]V.Pohl,V.Jungnickel,T.Haustein,and C.von Helmolt,"Antenna spacing in MIMO indoor channels,"Proc.IEEE Veh.Technol.Conf.,vol.2,pp.749-753,May2002.

[0015] [3]M.Stoytchev,H.Safar,A.L.Moustakas,and S.Simon,"Compact antenna arrays for MIMO applications,"Proc.IEEE Antennas and Prop.Symp.,vol.3,pp.708-711,July2001.

[0016] [4]A.Forenza and R.W.Heath Jr.,"Impact of antenna geometry on MIMO communication in indoor clustered channels,"Proc.IEEE Antennas and Prop.Symp.,vol.2,pp.1700-1703,June2004.

[0017] 此外,对于小的天线间隔,彼此之间的耦合效应可能会降低MIMO系统的性能。

[0018] 参见例如以下参考文献:

[0019] [5]M.J.Fakhereddin and K.R.Dandekar,"Combined effect of polarization diversity and mutual coupling on MIMO capacity,"Proc.IEEE Antennas and Prop.Symp.,vol.2,pp.495-498,June2003.

[0020] [7]P.N.Fletcher,M.Dean,and A.R.Nix,"Mutual coupling in multi-element array antennas and its influence on MIMO channel capacity,"IEEE Electronics Letters,vol.39,pp.342-344,Feb.2003.

[0021] [8]V.Jungnickel,V.Pohl,and C.Von Helmolt,"Capacity of MIMO systems with closely spaced antennas,"IEEE Comm.Lett.,vol.7,pp.361-363,Aug.2003.

[0022] [10]J.W.Wallace and M.A.Jensen,"Termination-dependent diversity performance of coupled antennas:Network theory analysis,"IEEE Trans.Antennas Propagat.,vol.52,pp.98-105,Jan.2004.

[0023] [13]C.Waldschmidt,S.Schulteis,and W.Wiesbeck,"Complete RF system model for analysis of compact MIMO arrays,"IEEE Trans.on Veh.Technol.,vol.53,pp.579-586,May2004.

[0024] [14]M.L.Morris and M.A.Jensen,"Network model for MIMO systems with coupled antennas and noisy amplifiers,"IEEE Trans.Antennas Propagat.,vol.53,

pp.545-552,Jan.2005.

[0025] 而且，当天线拥挤到一起的时候，天线通常必须做得更小，这也能够影响天线效率。

[0026] 参见例如以下参考文献：

[0027] [15]H.A.Wheeler,"Small antennas,"IEEE Trans.Antennas Propagat., vol.AP-23,n.4,pp.462-469,July1975.

[0028] [16]J.S.McLean,"Are-examination of the fundamental limits on the radiation Q of electrically small antennas,"IEEE Trans.Antennas Propagat., vol.44,n.5,pp.672-676,May1996.

[0029] 最后，用较低频率和较长波长的话，MIMO装置的物理尺寸就变得难以处理。一个极端的例子是在HF波段，这里MIMO装置天线必须互相分开10米或更大距离。

[0030] 2.噪声限制。每个MIMO的接收器/发送器子系统产生一定水平的噪声。当越来越多的这种子系统互相临近放置时，背景噪声就会上升。同时，当需要从多天线MIMO系统中识别出更多不同信号的时候，就要求更低的背景噪声。

[0031] 3.成本和功率限制。尽管有些MIMO应用中成本和功耗不是焦点，但在典型的无线产品中，开发一种成功的产品时，成本和功耗都是至关重要的制约因素。对于每个MIMO天线，需要分离的RF子系统，包括分离的模-数(A/D)和数-模(D/A)转换器。不像以摩尔定律来衡量规模的数字系统的很多方面(英特尔的共同创立者戈登摩尔所作出的经验层面的观察结果，微型器件的集成电路上的晶体管数目大约每隔24个月便会翻两倍；来源：http://www.intel.com/technology/mooreslaw/)，这样密集的模拟子系统通常具有一定的物理结构尺寸和功率要求，其尺寸与成本和功率线性成比例。因此，和单天线装置相比，多天线MIMO装置将变得极其昂贵并且具有惊人的能耗。

[0032] 作为上面的结果，今天预期的大多数MIMO系统是在2至4个天线的等级上，导致吞吐量2至4倍的上升和由于多天线系统的分集益处而引起的一些SNR(信噪比)的上升。已经预期到10个天线的MIMO系统(特别是由于较短的波长和较近的天线间隔的在较高的微波频率上)，但是除了对于一些特殊的和对成本不敏感的应用以外，超过10个天线是很不实际的。

[0033] 虚拟天线阵列

[0034] MIMO类型的技术的一种特殊应用是虚拟天线阵列。欧洲科学技术领域研究协作组织提出的研究文件中建议了这种系统，EURO,Barcelona,Spain,2003年1月15-17日；Center for Telecommunication Research,King's College London,UK;"A step towards MIMO: Virtual Antenna Arrays",Mischa Dohler & Hamid Aghvami。

[0035] 如文件中所述，虚拟天线阵列是协作无线装置系统(例如蜂窝电话)，其在分离的通信信道上互相通信(假如当它们相互足够临近)，而不是在它们主要的通信信道上与它们的基站通信，使得协作性地工作(例如，如果它们是UHF波段中的GSM蜂窝电话，那么这可以是5GHz的工业科学医学(ISM)无线波段)。通过在相互的中继范围(除了在基站范围内)内的几个装置之间转发信息，就好像他们是在物理上具有多个天线的一个装置工作一样，使得单天线装置潜在地实现象MIMO一样的吞吐量提升。

[0036] 然而，实际上，这样的系统极难实现并且用处有限。首先，必须保持每个装置现在

最少有两个不同的通信路径以实现吞吐率提升，其第二中继链路的可用性经常是不确定的。而且，由于它们最少具有第二通信子系统并且有更大的计算需求，因此该装置更昂贵，物理尺寸更大，并且消耗更多的功率。此外，潜在地通过多个通信链路，该系统依赖于非常复杂的所有系统的实时协作。最后，由于同时发生的信道利用增加(例如，使用MIMO技术的同时发生的电话呼叫传输)，对于各装置的计算负担也就增加了(通常随信道利用的线性增加而成指数增加)，这对具有严格的功率和尺寸限制的便携装置是很不实际的。

发明内容

[0037] 描述了一种用于对具有多用户(MU)发送("MU-MAS")的多天线系统(MAS)中的频率和相位偏移进行补偿的系统和方法。例如，根据本发明一种实施方式的方法包括：将来自基站每一天线的训练信号发送至多个无线客户装置中的一个或每个无线客户装置，该客户装置中的一个或每个客户装置分析每个训练信号以生成频率偏移补偿数据，并在基站处接收频率偏移补偿数据；基于所述频率偏移补偿数据来计算MU-MAS预编码器权重以预消除发射机处的频率偏移；使用所述MU-MAS预编码器权重对训练信号进行预编码，以生成针对基站每一天线的预编码训练信号；将来自所述基站的每个天线的预编码后的训练信号发送到所述多个无线客户装置中的每一个无线客户装置，每个客户装置分析每个训练信号以生成信道特征数据，并在所述基站接收所述信道特征数据；基于该信道特征数据来计算多个MU-MAS预编码权重，该MU-MAS预编码器权重被计算来用于预消除频率和相位偏移和/或用户之间的干扰；使用MU-MAS预编码器权重来对数据进行预编码，以生成针对基站每一天线的预编码后的数据信号；以及将所述预编码后的预编码数据信号通过基站的每个天线发送至其每个的客户端设备。

附图说明

[0038] 结合附图，下面详尽的描述可以获得对本发明更好的理解，其中：

[0039] 图1显示了现有技术的MIMO系统。

[0040] 图2显示了与多个单天线客户装置进行通信的N天线基站。

[0041] 图3显示了与三个单天线客户装置进行通信的三个天线的基站。

[0042] 图4显示了本发明的一个实施例中使用的训练信号技术。

[0043] 图5显示了根据本发明一个实施例的从客户装置传输到基站的信道特征数据。

[0044] 图6显示了根据本发明一个实施例的多重输入分布式输出("MIDO")下行传输。

[0045] 图7显示了根据本发明一个实施例的多重输入多重输出("MIMO")上行传输。

[0046] 图8显示了根据本发明的一个实施例的通过不同客户群循环以分配吞吐量的基站。

[0047] 图9显示了根据本发明的一个实施例的基于临近的客户分组。

[0048] 图10显示了在NWIS系统中使用的本发明的实施例。

[0049] 图11显示了具有I/Q补偿功能单元的DIDO发射机的实施方式。

[0050] 图12显示了具有I/Q补偿功能单元的DIDO接收机。

[0051] 图13显示了具有I/Q补偿的DIDO-OFDM系统的一种实施方式。

[0052] 图14显示了在具有和不具有I/Q补偿的情况下DIDO2×2性能(performance)的一

种实施方式。

[0053] 图15显示了在具有和不具有I/Q补偿的情况下DIDO2×2性能的一种实施方式。

[0054] 图16显示了在具有和不具有I/Q补偿的情况下针对不同QAM星座图的SER(符号误码率)的一种实施方式。

[0055] 图17显示了在不同用户设备位置具有和不具有I/Q补偿的情况下DIDO2×2性能的一种实施方式。

[0056] 图18显示了在理想(i.i.d.(独立且同分布))信道中具有和不具有I/Q补偿的情况下SER的一种实施方式。

[0057] 图19显示了自适应DIDO系统的发射机架构的一种实施方式。

[0058] 图20显示了自适应DIDO系统的接收机架构的一种实施方式。

[0059] 图21显示了自适应DIDO-OFDM的方法的一种实施方式。

[0060] 图22显示了用于DIDO测量的天线布局的一种实施方式。

[0061] 图23显示了用于不同级别(order)DIDO系统的阵列配置的实施方式。

[0062] 图24显示了不同级别DIDO系统的性能。

[0063] 图25显示了用于DIDO测量的天线阵列的一种实施方式。

[0064] 图26显示了4-QAM且1/2FEC率的DIDO2×2性能与用户设备位置的函数关系的一种实施方式。

[0065] 图27显示了用于DIDO测量的天线布局的一种实施方式。

[0066] 图28显示了在一种实施方式中DIDO8×8如何产生比具有低TX功率需求的DIDO2×2更大的SE。

[0067] 图29显示了在具有天线选择情况下的DIDO2×2性能的一种实施方式。

[0068] 图30显示了不同DIDO预编码方案在i.i.d.信道中的平均比特误码率(BER)性能。

[0069] 图31显示了ASel的信噪比增益与i.i.d.信道中额外发射天线的数量之间的函数关系。

[0070] 图32显示了在i.i.d.信道中具有1和2个外部天线的情况下SNR阈值与用于块对角化(BD)和ASel的用户数量(M)之间的函数关系。

[0071] 图33显示了针对位于相同角度方向且具有不同角度扩展(AS)值的两个用户的BER与每用户平均SNR。

[0072] 图34显示了与图33相类似的结果,但用户之间具有更高的角度间隔。

[0073] 图35绘制了针对用户的平均到达角度(AOA)的不同值,AS与SNR阈值之间的函数关系。

[0074] 图36显示了针对5个用户的示例性情况的SNR阈值。

[0075] 图37针对2个用户的情况,提供了在具有1和2个额外天线的情况下,SNR阈值BD与ASel的比较。

[0076] 图38显示了与图37相类似的结果,但是针对5个用户的情况。

[0077] 图39显示了针对具有不同AS值的BD方案的SNR阈值。

[0078] 图40显示了对于具有1和2个额外天线的BD和ASel,在具有AS=0.1°的空间关联信道中的SNR阈值。

[0079] 图41显示了针对AS=5°的另外两个信道情形的SNR阈值的计算。

[0080] 图42显示了针对AS=10°的另外两个信道情形的SNR阈值的计算。

[0081] 图43-图44分别显示了在1和2个额外天线的情况下，SNR阈值与用户数量(M)与BD和ASel方案的角度扩展(AS)之间的函数关系。

[0082] 图45显示了配备有频率偏移估计器/补偿器的接收机；

[0083] 图46显示了根据本发明一种实施方式的DIDO2×2系统模型。

[0084] 图47显示了根据本发明一种实施方式的方法。

[0085] 图48显示了在具有和不具有频率偏移的情况下，DIDO2×2系统的SER结果。

[0086] 图49将不同DIDO方案的SNR阈值性能进行了比较。

[0087] 图50将不同方法实施方式所需的开销量进行了比较。

[0088] 图51显示了在f_{offset}=2Hz的小频率偏移且没有整数偏移校正的情况下的仿真。

[0089] 图52显示了当关闭整数偏移估计器时的结果。

具体实施方式

[0090] 在下列描述中，为了解释的目的，为了提供对本发明彻底的理解，阐明了多个特殊细节。然而，很明显的是，对于本领域内的普通技术人员，即使没有一些特殊细节，仍然可以实现本发明。此外，公知的结构和装置显示为框图形式，以避免将本发明根本的原理模糊化。

[0091] 图1显示了具有发射天线104和接收天线105的现有技术MIMO系统。这样的系统的吞吐率可以实现通常在可用信道中实现的吞吐率的3倍。有多种不同的方法来实现这种MIMO系统的细节，其在关于该主题的出版文献中有过描述，下面的解释将描述一个这样的方法。

[0092] 数据在图1的MIMO系统中传输之前，信道被"特征化"。这是通过在开始将"训练信号"从每个发射天线104传输到每个接收器105来实现的。训练信号有编码和调制子系统102生成，并被D/A转换器(没有示出)转换成模拟信号，然后由每个发送器103从基带信号转换为RF信号。每个耦合到其RF接收器106的接收天线105接收每个训练信号并将其转换为基带信号。基带信号由D/A转换器(没有示出)转换为数字信号，然后信号处理子系统107特征化该训练信号。每个信号的特征可以包括很多因素，例如，其包括，相对于接收器内部的参考信号的相位和振幅、绝对参考信号、相对参考信号、特征噪声或其他因素。每个信号的特征通常定义为当信号通过信道传送时表现信号几个方面的相位和振幅变化的向量。例如，在正交幅度调制("QAM")的调制信号中，所述特征可能是信号的几个多径映像的相位与振幅偏移的向量。另外一个例子是，在正交频分复用("OFDM")调制的信号中，它可能是OFDM频谱中几个或所有单个分量信号(sub-signal)的相位与振幅偏移的向量。

[0093] 信号处理子系统107将由每个接收天线105和相应接收器106接收到的信道特征存储起来。所有的三个发射天线104完成它们的训练信号传输之后，信号处理子系统107将已经存储了三个对于三个接收天线105中每一个的信道特征。这形成了3×3的矩阵108，其表示为信道特征矩阵"H"。每个单独的矩阵元素$H_{i,j}$是接收天线105_i接收到的传输天线104_j的训练信号传输的信道特征。

[0094] 在这点上，信号处理子系统107将矩阵H108求逆以产生H^{-1}，并且等待从发射天线104来的实际数据的传输。注意，多种在可用文献中描述的现有MIMO技术可用于确保H矩阵

108可逆。

[0095] 在实施中，要传输的数据的内容(payload)送到数据输入子系统100。然后在送到编码和调制子系统102之前，其被分配器(splitter)101分割为三部分。例如，如果内容是"abcdef"的ASCII比特，它就可以被分配器分割为三个子内容"ad"、"be"和"cf"。然后，每个子内容单独发送给编码和调制子系统102。

[0096] 通过使用适合每个信号的统计独立性和纠错能力的编码系统，单独地对每个子内容进行编码。这些包括，而不仅仅限于，Reed-Solomon编码、维特比编码(Viterbi coding)和增强编码(Turbo Codes)。最后，使用对信道合适的调制方法对这三个编码后的子内容中的每一个进行调制。示例性的调制方法是差分相移键控调制("DPSK")、64-QAM调制和OFDM。这里应该注意的是，MIMO提供的分集增益允许较高级数的调制星座图，所述调制星座图在使用相同信道的SISO(单输入单输出)系统中也是可行的。然后，每个编码并且调制后的信号通过它自己的天线104传输出去，所述传输跟随在D/A转换单元(没有示出)的D/A转换和每个发送器103的RF生成之后。

[0097] 假设有足够的空间分集存在于发送和接收天线之间，每个接收天线105将从天线104接收三个传输信号的不同组合。每个RF接收器106将每个信号接收到并将它们转换为基带信号，然后A/D转换器(没有示出)再对信号进行数字化。如果y_n是由第n个接收天线105接收到的信号，x_n是由第n个发射天线104发送的信号，N是噪声，那么这就能以下列等式描述。

[0098] $y_1=x_1h_{11}+x_2h_{12}+x_3h_{13}+N$

[0099] $y_2=x_1h_{21}+x_2h_{22}+x_3h_{23}+N$

[0100] $y_3=x_1h_{31}+x_2h_{32}+x_3h_{33}+N$

[0101] 假设这是一个具有三个未知量的三个等式的系统，那么这就是信号处理子系统107推导出x_1、x_2和x_3的线性代数的问题了(假设N在足够低的水平，允许对信号进行解码)：

[0102] $x_1=y_1H^{-1}_{11}+y_2H^{-1}_{12}+y_3H^{-1}_{13}$

[0103] $x_2=y_1H^{-1}_{21}+y_2H^{-1}_{22}+y_3H^{-1}_{23}$

[0104] $x_3=y_1H^{-1}_{31}+y_2H^{-1}_{32}+y_3H^{-1}_{33}$

[0105] 一旦推导出三个传送的信号x_n，它们就被信号处理子系统107解调、解码和纠错，以恢复由原来由分配器101分开的三个比特流。这些比特流在合并器单元108中合并，并从数据输出109中输出为单数据流。假设系统强健性可以克服噪声损伤，那么数据输出109产生的比特流将和引入到数据输入100中的比特流一样。

[0106] 尽管所描述的现有技术系统通常有效直到四个天线，或许直到10个之多的天线，由于在该公开的背景部分中描述的原因，具有大量天线(例如125、100或1000)时其变得很不实际。

[0107] 通常，这样的现有技术系统是双向的，返回路径以完全相同的方式实现，但是反过来，在通信信道的每一侧都具有发送和接收子系统。

[0108] 图2显示了本发明的一个实施例。在其中，基站(BS)200配置有广域网(WAN)接口(例如通过T1或其它高速连接)201并且提供有一定数量的(N个)天线202。我们暂且使用术语"基站"来指代与固定位置的一组客户进行无线通信的任何无线站点。基站的示例可为无线局域网(WLAN)中的接入点，或WAN天线或天线阵列。有一些客户装置203-207，每个具有单天线，基站200通过无线方式对它们进行服务。尽管对于这个例子的目的，非常容易想到位

于办公室环境的基站，在这种环境中，其为装备有无线网络的个人计算机的用户装置203-207提供服务，但这种结构将运用于大量的应用情况，室内和室外，在这里基站服务于无线客户。例如，所述基站可以位于蜂窝电话塔上，或者位于电视广播塔上。在一个实施例中，基站200被安置于地面，用于HF频率的(例如24MHz的频率)上行传送，以将信号从电离层反射回来，如2004年4月20日提出的，序列号为No.10/817,731，题目为SYSTEM AND METHOD FOR ENHANCING NEAR VERTICAL INCIDENTCE SKYWAVE("NVIS")COMMUNICATION USING SPACE-TIME CODING的同时未决的申请描述的一样，其被支配给本申请的代理人，在这里作为参考。

[0109] 与基站200相联系的某些细节和所阐明的客户装置仅仅是为了例证的目的，而不是根据本发明的根本原理必需的。例如，该基站可以经由WAN接口201连接于多个不同类型的广域网，其包括专用广域网，例如那些用于数字视频发送的广域网。类似地，客户装置可以是任何种类的无线数据处理和/或通信装置，其包括，而不仅仅局限于，蜂窝电话、个人数字助理("PDA")、接收器和无线相机。

[0110] 在一个实施例中，基站的n个天线202在空间上是分开的，从而每一个发送和接收非空间相关的信号，就好像所述基站是现有技术MIMO的收发器一样。如在背景技术中所描述的，天线以λ/6(即1/6波长)间隔放置的实验已经做出，其成功地实现了从MIMO的吞吐量提升，但一般来说，这些基站天线越进一步分开放置，系统的性能就越好，λ/2是令人满意的最小距离。当然，本发明的根本原理不限于天线间任何特定的分离。

[0111] 注意，单基站200可以很好地将其天线放置于很远的距离，例如，在HF频谱中，天线可以有10米或更远(例如，在上面提到的NVIS实现中)。如果使用100个这样的天线，该基站的天线阵列就可以占有几个平方公里的面积。

[0112] 除了空间分集技术之外，为了提高系统的有效吞吐量，本发明的一个实施例将信号极化。通过极化来提高信道容量是一种公知的技术，其已经被卫星电视供应商使用了很多年。使用极化技术，可以使多个(例如三个)基站或用户天线互相间非常接近，并且仍然是非空间相关的。尽管传统的RF系统通常仅仅受益于极化的二维(例如x和y)分集，但这里描述的结构可进一步受益于极化的三维(x,y和z)分集。

[0113] 除了空间和极化分集之外，本发明的一种实施方式采用近乎正交的辐射方向图(pattern)，以经由方向图分集来改善链路性能。方向图分集可改善MIMO系统的容量和误码率性能，且其相比于其他天线分集技术的优点可参见以下文章：

[0114] [17]L.Dong,H.Ling,and R.W.Heath Jr.,"Multiple-input multiple-output

[0115] wireless communication systems using antenna pattern diversity," Proc.IEEE Glob.Telecom.Conf.,vol.1,pp.997-1001,Nov.2002.

[0116] [18]R.Vaughan,"Switched parasitic elements for antenna diversity," IEEE

[0117] Trans.Antennas Propagat.,vol.47,pp.399-405,Feb.1999.

[0118] [19]P.Mattheijssen,M.H.A.J.Herben,G.Dolmans,and L.Leyten,"Antenna-pattern diversity versus space diversity for use at handhelds,"IEEE Trans.on Veh.Technol.,vol.53,pp.1035-1042,July2004.

[0119] [20]C.B.Dietrich Jr,K.Dietze,J.R.Nealy,and W.L.Stutzman,"Spatial,

polarization,and pattern diversity for wireless handheld terminals,"Proc.IEEE Antennas and Prop.Symp.,vol.49,pp.1271-1281,Sep.2001.

[0120] [21]A.Forenza and R.w.Heath,Jr.,"Benefit of Pattern Diversity Via2-element Array of Circular Patch Antennas in Indoor Clustered MIMO Channels", IEEE Trans.on Communications,vol.54,no.5,pp.943-954,May2006.

[0121] 通过使用方向图分集,可使得多个基站或用户天线相互之间非常接近,且尽管如此也不会在空间上相关联。

[0122] 图3提供了图2中所示的基站200和客户装置203-207的一个实施例的额外细节。为了简化的目的,该基站300仅仅显示为三个天线305和三个客户装置306-308。然而,需要注意的是,这里描述的本发明的实施例可以用几乎无限数量的天线305(即,仅仅由可用的空间和噪声来限制)和客户装置306-308来实现。

[0123] 图3与图1所示的现有技术MIMO结构类似,其中,两者在通信信道的每一端有三个天线。显著的区别是,在现有技术的MIMO系统中,图1右侧的三个天线105互相之间是固定距离(例如,集成在单一装置中),从每个天线105接收到的信号一起在信号处理子系统107中得到处理。相比之下,在图3中,图右侧的三个天线309每一个都耦合到不同的客户装置306-308上,每个所述客户装置都可以分布于基站305的范围内的任何地方。这样,每个客户装置接收到的信号可以在其解码、调制、信号处理子系统311中独立于其它两个接收到的信号而得到处理。因此,与多重输入(即天线105)多重输出(即天线104)的"MIMO"系统相比较,图3显示了多重输入(即天线305)分布式输出(即天线305)系统,以下指"MIDO"系统。

[0124] 注意,本申请使用与之前的申请不同的术语用法,以更好地符合学术界及行业惯例。在之前所引用的2004年4月20日提交的题为"SYSTEM AND METHOD FOR ENHANCING NEAR VERTICAL INCIDENCE SKYWAVE("NVIS") COMMUNICATION USING SPACE-TIME CODING"的共同待审的申请NO.10/817,731以及2004年7月30日提交的申请NO.10/902,978(本申请是该申请的继续申请)中,"输入"和"输出"(在SIMO、MISO、DIMO以及MIDO的环境中)的意思与该术语在本申请中的表意是相反的。在之前的申请中,"输入"指输入至接收天线(例如,图3中的天线309)的无线信号,而"输出"指发射天线(例如,天线305)输出的无线信号。在学术界和无线行业中,通常使用"输入"和"输出"的反义,其中"输入"指输入至信道的无线信号(即,从天线305发送的无线信号),而"输出"指从信道输出的无线信号(即,天线309所接收的无线信号)。本申请采用此术语用法,该用法与本段之前所引用的申请中的用法相反。因此,以下给示了几个申请之间的术语用法等价形式:

[0125]

10/817,731 和 10/902,978		本申请
SIMO	=	MISO
MISO	=	SIMO
DIMO	=	MIDO
MIDO	=	DIMO

[0126] 图3所示的MIDO结构对于给定数量的发射天线实现了类似于MIMO在SISO系统上实现的容量提升。然而,MIMO和图3所示的特定MIDO实施例的一个区别是,为实现多个基站天线提供的容量提升,每个MIDO客户装置306-308仅仅要求单个接收天线,而对于MIMO,每个客户装置至少要求与希望实现的容量倍数一样多的接收天线。假设通常有一实行的限制,

其限制能够在客户装置放置多少天线(如在背景技术中解释的),典型上这就将MIMO系统限制在4个至10个天线之间(4倍至10倍的容量)。由于基站300通常从固定和装有动力的位置服务于很多客户装置,将其扩展为远超过10个天线,和用适当的距离分离天线以实现空间分集是很实际的。如所述,每个天线装备有收发器304和一部分编码、调制和信号处理部件303的处理能力。值得注意的是,在此实施例中,无论基站300扩展多少,每个客户装置306~308将仅仅要求一个天线309,因此对于单用户客户装置306~308的成本将很低,并且基站300的成本可以在大基数的用户中分担。

[0127] 在图4至图6中,显示了如何完成从基站300到客户装置306~308的MIDO传输的例子。

[0128] 在本发明的一个实施例中,MIDO传输开始之前,信道被特征化。对于MIMO系统,每个天线405对训练信号一个接一个进行传输。图4仅仅显示了第一个训练信号的传输,但对于三个天线405来说,共有三个分开的传输。每个训练信号由编码、调制和信号处理子系统403生成,通过D/A转换器转换成模拟信号,并作为RF信号通过每个RF收发器404发送出去。可用的各种不同的编码、调制和信号处理技术包括,而不局限于那些上面描述的技术(例如,Reed Solomon、维特比编码(Viterbi Coding)、QAM、DPSK、QPSK调制等等)。

[0129] 每个客户装置406~408通过其天线409接收训练信号并通过收发器410将该训练信号转换成基带信号。A/D转换器(没有示出)在该信号被编码、调制和信号处理子系统411处理的地方将其转换成数字信号。然后信号特征逻辑单元320识别所得信号的特征(例如,识别上述的相位和振幅失真)并将该特征存放到存储器中。这个特征处理过程类似于现有技术的MIMO系统的处理过程,一个显著的区别是,每个客户装置仅仅计算其一个天线,而不是n个天线的特征向量。例如,已知模式的所述训练信号将客户装置406的编码、调制和信号处理子系统420初始化(在产生时通过在发送的信息中接收它,或通过其他初始化处理)。当天线405以已知模式发送该训练信号的时候,编码、调制和信号处理子系统420使用相关法来找到最强的训练信号接收模式,其将相位和振幅偏移保存起来,然后其将该模式从接收到的信号中减掉。接下来,其找到与所述训练信号相关的第二强接收模式,将相位和振幅偏移保存起来,然后其将第二强模式从所述接收到的信号中减掉。该处理一直进行,直到保存了某固定数量的相位和振幅偏移(例如,8个)或可探测的训练信号模式下降到给定的背景噪声之下。该相位/振幅偏移的向量成为向量413的元素H_{11}。与此同时,客户装置407和408的编码、调制和信号处理子系统执行同样的处理,产生它们的向量元素H_{21}和H_{31}。

[0130] 信道特征存放的存储器可以是非易失性存储器,例如闪存,或硬盘,和/或易失性存储器,例如随机存取存储器(例如,SDRAM、RDAM)。此外,不同的用户装置可以同时使用不同类型的存储器来存储特征信息(例如PDA可是使用闪存,而笔记本电脑可是使用硬盘)。在各种客户装置或基站上,本发明根本的原理不限于任何特定类型的存储机构。

[0131] 如上所述,根据所使用的方案,由于每个客户装置406~408仅有一个天线,每个仅仅存储H矩阵的1×3行413~415。图4显示了第一训练信号传输后的阶段,这里,1×3行413~415的第一列存储了三个基站天线405的第一个天线的信道特征信息。其余两列存储了从其余两个基站天线的接下来的两个训练信号传输的信道特征。注意,为了例证的目的,所述三个训练信号模式在分开的时间传输。如果选择了三个训练信号模式从而互不相关,那么它们可以同时传输,因此减少训练时间。

[0132] 如图5所示，所有三个导频传输完成之后，每个客户装置506-508将已经存储起来的矩阵H的1×3行513-515发送回基站500。为了简化的目的，在图5中仅显示有一个客户装置506传送其特征信息。结合适当的纠错编码(例如Reed Solomon、维特比编码(Viterbi Coding)和/或涡强编码(Turbo Codes))，可以使用合适的调制方法(例如DPSK、64QAM、OFDM)来确保基站500准确地接收行513-515中的数据。

[0133] 图5中，尽管所有三个天线505显示出接收信号，但是对于接收每1×3行513-515的传输，基站500的单天线和单收发器已经足够了。然而，在一定条件下，使用很多或所有天线505和收发器504来接收每个传输(即，在编码、调制和信号处理子系统503中使用现有技术的单输入多重输出("SIMO")处理技术)可以实现比单天线505和收发器504更好的信噪比(SNR)。

[0134] 当基站500的编码、调制和信号处理子系统503从每个客户装置507-508接收所述1×3行513-515的时候，其将所述1×3行513-515存入3×3的H矩阵516中。对于客户装置，基站可以使用很多不同的存储技术来存储矩阵516，其包括，但不局限于，非易失性海量存储器(例如硬盘)和/或易失性存储器(例如SDRAM)。图5显示了基站已经接收和存储了来自客户装置509的1×3行513的阶段。当1×3行514和515从其余客户装置传输的时候，它们可以被传输并保存在H矩阵516中，直到整个H矩阵516被存储起来。

[0135] 参考图6，现在将描述从基站600到客户装置606-608的MIMO传输的实施例。因为每个客户装置606-608是独立的装置，所以每个装置接收不同的数据传输。这样，基站600的实施例包括位于WAN接口601和编码、调制与信号处理子系统603之间对它们进行通信联络的路由器602，其从WAN接口601接收多个数据流(格式为比特流)，分别对应于每个客户装置606-608将所述数据流按分开的数据流u_1-u_3发送。为此目的，该路由器602可以使用各种已知的路由技术。

[0136] 如图6所示，将所述三个比特流，u_1-u_3，路由进所述编码、调制和信号处理子系统603中，将它们编码为统计独立的纠错流(例如，使用Reed Solomon、维特比或增强编码)，并用对信道合适的调制方法(例如DPSK、64QAM或OFDM)将它们调制。此外，图6显示的实施例包括信号预编码逻辑单元630，基于信号特征矩阵616，该信号预编码逻辑单元630用于对从每个天线605发送来的信号进行唯一编码。特别地，在该实施例中，预编码逻辑单元630将图6中的三个比特流u_1-u_3与H矩阵616的逆矩阵相乘以生成三个新的比特流u'_1-u'_3，而不是将每个编码和调制过的比特流路由到分开的天线(如图1中所做)。然后，D/A转换器(没有示出)将所述的三个预编码比特流转换为模拟信号，收发器604和天线605将其作为RF信号发送出去。

[0137] 在解释客户装置606-608如何接收所述预特流之前，将描述预编码模块630执行的操作。类似于上面图1中MIMO的例子，三个原始比特流中每一个比特流的编码和调制过的信号将表示为u_n。在图6所示的实施例中，每个u_i包含路由器602所路由的三个比特流来的数据，每个这样的比特流将成为三个用户装置606-608其中的一个。

[0138] 然而，不象图1中的MIMO例子，那里，每个x_i有各天线104发送，在图6所示的本发明的实施例中，在每个客户装置天线609接收各u_i(加上信道中任何的噪声N)。为实现这样的结果，三个天线605中的每一个的输出(我们将其表示为v_i)是u_i和特征化每个客户装置的H矩阵的函数。在实施例中，编码、调制和信号处理子系统中的预编码逻辑单元630通过执行

下列等式来计算各v_i：

[0139] $v_1=u_1H^{-1}_{11}+u_2H^{-1}_{12}+u_3H^{-1}_{13}$

[0140] $v_2=u_1H^{-1}_{21}+u_2H^{-1}_{22}+u_3H^{-1}_{23}$

[0141] $v_3=u_1H^{-1}_{31}+u_2H^{-1}_{32}+u_3H^{-1}_{33}$

[0142] 因此，不像MIMO一那里，信道将信号变换之后在接收器计算各x_i，而这里描述的本发明的实施例在信道将信号变换之前在发送器求解每个v_i。每个天线609接收已经从其它用于其它天线609的u_i比特流中分离出来的u_i。每个收发器610将各接收到的信号转换成基带信号，这里A/D转换器(没有示出)对其进行数字化，各编码、调制和信号处理子系统611对其x_i比特流进行解调和解码，并将其比特流送到客户装置使用的数据接口612(例如，客户装置上的应用程序)。

[0143] 这里描述的本发明的实施例可以使用多种不同的编码和调制方法来实现。例如，在OFDM实现中，其中频谱被分为多个分频带，这里描述的技术可用于特征化每个单独的分频带。然而，如上所述，本发明的根本原理不限于任何特定的调制方法。

[0144] 如果客户装置606-608是便携式数据处理装置，例如PDA、笔记本电脑和/或无线电话的话，那么由于客户装置可能会从一个位置移动到另外一个，则信道特征会频繁发生改变。这样，在本发明的一个实施例中，基站的信道特征矩阵616不断地得到更新。在一个实施例中，基站600周期地(每250毫秒)发出新的训练信号到每个客户装置，每个客户装置将其信道特征向量不断地发送回基站600以确保信道特征保持准确(例如，如果环境改变或客户装置移动从而影响到信道)。在一个实施例中，在发送到每个客户装置的实际数据信号中对训练信号进行交织。典型地，所述训练信号的吞吐量远低于所述数据信号的吞吐量，因此这对系统总的吞吐率将几乎没有影响。相应地，在该实施例中，信道特征矩阵616在基站主动与各客户装置进行通信时可以不断得到更新，从而当客户装置从一个位置移动到下一个位置，或环境发生改变从而影响到信道的时候保持准确的信道特征。

[0145] 图7中所示的本发明的一个实施例使用MIMO技术来改善上行通信信道(即，从客户装置706-708到基站700的信道)。在该实施例中，基站中的上行信道特征逻辑单元741不断对从每个客户装置来的信道进行分析和特征化。特别地，每个客户装置706-708发送训练信号到基站700，那里信道特征逻辑单元741分析以产生N×M的信道特征矩阵741，这里N是客户装置的数量，M是基站所使用的天线的数量。图7所示的实施例在基站使用三个天线705和三个客户装置706-708，这导致了存放于基站700的3×3信道特征矩阵741。客户装置可以将图7所示的MIMO上行传输用于将数据发送回基站700和将信道特征向量传送回基站700，如图5所示。但是和图5所示的实施例不同的是，在图5中，每个客户装置的信道特征向量以分开的时间进行传输，而图7所示的方法允许从多个客户装置同时将信道特征向量传输回基站700，从而大大降低信道特征向量对回程信道吞吐率的影响。

[0146] 如上所述，每个信号的特征可以包括很多因素，例如，其包括相对于接收器内部的参考信号、绝对参考信号、相对参考信号、特征噪声或其他因素的相位和振幅。例如，在正交幅度调制所调制的信号中，所述特征可以是信号的几个多径映像的相位和振幅偏移向量。另一个例子是，在正交频分复用所调制的信号中，所述特征可以是OFDM频谱中几个或所有单个分量信号的相位和振幅偏移向量。所述训练信号可以由各客户装置的编码和调制子系统711生成，D/A转换器(未示出)将该训练信号转换成模拟信号，然后各客户装置的发送器

709将其从基带信号转换成RF信号。在一个实施例中，为了确保训练信号的同步，客户装置仅仅在基站请求的时候传送训练信号（例如，在循环（round robin）的情况下）。此外，可以在从各客户装置发送来的实际数据信号中对训练信号进行交织，或者训练信号可以和所述实际数据信号一起传输。因此，即使客户装置706-708是移动的，上行信道特征逻辑单元741也可以连续地传输和分析该训练信号，从而确保信道特征矩阵741保持更新。

[0147] 本发明的前述实施例所支持的总的信道容量可以被定义为min(N,M)，这里，M是客户装置的数量，而N是基站天线的数量。也就是说，容量由基站侧或客户侧的天线数量所限定。如此，本发明的一个实施例使用同步技术来确保在给定时间内不超过min(N,M)个天线在发送/接收。

[0148] 在典型的情况下，基站700的天线705的数量将少于客户装置706-708的数量。图8显示了一个示例性的情况，其允许5个客户装置804-808与具有三个天线802的基站进行通信。在这个实施例中，确定总的客户装置804-808的数量并且检测到必要的信道特征信息（例如，上面的描述）之后，基站800选择第一群与其进行通信的三个客户810（因为min(N,M)=3，所以在此例中是三个客户）。在与第一群客户810通信了指定时间之后，基站就选择另一群与其通信的三个客户811。为了均匀分配通信信道，基站800选择没有包含在第一群中的两个客户装置807、808。此外，由于额外的天线是可用的，基站800就选择包含在第一群中的额外的客户装置806。在一个实施例中，基站800以这种方式在客户群众循环，从而能够有效地分配给每个客户在时间上相同数量的吞吐量。例如，为了均匀分配吞吐量，基站可以接着选择除客户装置806之外的三个客户装置的任何组合（即，由于客户装置806用于在开始的两个循环中与基站进行通信）。

[0149] 在一个实施例中，除了标准的数据通信之外，基站可以使用前述技术来传送训练信号到各客户装置和从各客户装置接收训练信号和信号特征数据。

[0150] 在一个实施例中，某些客户装置或客户装置群可以分配到不同水平的吞吐量，例如，可以把客户装置区分优先次序，从而可以确保相对较高优先级的客户装置比较低优先级的客户装户有更多的通信周期（即，更多的吞吐量）。基于一定数量的变量，可以对客户的"优先级"进行选择，所述变量包括，例如，用户的对无线带宽的预订费（例如，用于愿意为额外吞吐量付出更多），和/或通信到/从客户装置的数据类型（例如，实时通信，譬如电话语音和视频，获得高于非实时通信的优先级，例如电子邮件）。

[0151] 在基于各客户装置要求的当前负载，基站动态分配吞吐量的实施例中。例如，如果客户装置804直播视频流，而其它装置805-808在执行例如电子邮件的非实时功能，那么基站800可以给该客户804分配相对较多的吞吐量。然而，应该注意的是，本发明的根本原理不限于任何特定的吞吐量分配技术。

[0152] 如图9所示，两个客户装置907、908可以非常接近，使得所述客户的信道特征在实际上是一样的。结果，基站将接收和存储两个客户装置907、908的实际上相等的信道特征向量，因此这将不能产生对于各客户唯一的、空间分布的信号。相应地，在一个实施例中，基站将确保相互距离非常接近的任何两个或更多客户装置被分配给不同的群。例如，在图9中，基站900首先和客户装置904、905以及908的第一群910通信，然后和客户装置905、906、907的第二群911通信，这确保了客户装置907和908在不同的群中。

[0153] 可选择地，在一个实施例中，基站900同时和客户装置907以及908进行通信，但使

用已知的信道复用技术来对通信信道进行复用。例如，基站可以使用时分复用（"TDM"），频分复用（"FDM"）或码分多址（"CDMA"）技术来分开客户装置907和908之间单个的、空间相关的信号。

[0154] 尽管上述各客户装置装备有单天线，但可以通过使用具有多个天线的客户装置来实现本发明的根本原理以提高吞吐量。例如，当用在上述的无线系统上时，具有2个天线的客户将实现2倍的吞吐量提升，具有3个天线的客户将实现3倍的吞吐量提升，等等（即，假设天线之间的空间和角度分离是足够的）。当通过具有多个天线的客户装置循环的时候，基站可以应用同样的一般规则。例如，其可以将每个天线看作分开的客户，并将吞吐量分配给那个"客户"，就如同它是任何其它客户一样（例如，确保每个客户提供有足够或相当的通信周期）。

[0155] 如上所述，本发明的一个实施例使用上述的MIDO和/或MIMO信号传输技术在近乎垂直入射天波（"NVIS"）中提高信噪比和吞吐量。参考图10，在本发明的一个实施例中，装备有N个天线1002的矩阵的第一NVIS基站1001用于和M个客户装置1004进行通信。所述NVIS天线1002和多种用户装置的天线1004以和垂直方向约成15度以内的角度将信号上行传送以获得想要的NVIS并且将地面波干扰效应降到最低。在一个实施例中，天线1002和客户装置1004使用上述的多种MIDO和MIMO技术在NVIS频谱中的指定频率（例如在载波频率或低于23MHz的频率，但通常低于10MHz的频率上）支持多个独立的数据流1006，从而显著提高了在指定频率的吞吐量（即，以和统计独立的数据流的数量成正比）。

[0156] 服务于给定基站的所述NVIS天线相互之间可以有很远的物理距离。假设低于10MHz的长波长和信号传播的长距离（300英里的往返距离），几百码，甚至是几英里的天线物理间隔能够在分集上提供益处。在这样的条件下，单独的天线信号可以被收回到中心位置，用传统的有线或无线通信系统对其进行处理。可选择地，每个天线可以具有本地设备来处理其信号，然后使用传统的有线或无线通信系统来将该数据传输回中心位置。在本发明的一个实施例中，NVIS基站1001具有到因特网1010（或其它广域网）的宽带链路1015，从而提供给客户装置1003远程、高速、无线网络访问。

[0157] 在一种实施方式中，基站和/或用户可利用极化/方向图分集（pattern diversity）技术，以在提供分集与提升吞吐量的同时，减小阵列大小和/或用户距离。例如，在具有HF传输的DIMO系统中，由于极化/方向图分集，用户可位于同一位置且他们的信号不会相关联。特别地，通过使用方向图分集，一用户可经由地波面与基站进行通信，而其他用户可经由NVIS面与基站进行通信。

[0158] 本发明的附加实施方式

[0159] 1、利用I/Q不平衡来进行DIDO-OFDM预编码

[0160] 本发明的一种实施方式采用用于对具有正交频分复用（OFDM）的分布式输入分布式输出（DIDO）系统中的同相正交（I/Q）不平衡进行补偿的系统和方法。简言之，根据本实施方式，用户设备对信道进行估计，并将该信息回馈至基站；基站计算出预编码矩阵，以消除I/Q不平衡所导致的载波之间和用户之间的干扰；以及并行数据流经由DIDO预编码而被发送至多个用户设备；该用户设备经由零强制（ZF）、最小均方误差（MMSE）或最大似然（ML）接收机来对数据进行解调，以抑制剩余干扰。

[0161] 如下所详述的，本发明该实施方式的一些显著特征包括，但不限于：

[0162] 预编码以用于消除OFDM系统中来自镜像调(mirror tone)的载波间干扰(ICI)(因I/Q不匹配所导致)；

[0163] 预编码以用于消除DIDO-OFDM系统中的用户间干扰和ICI(因I/Q不匹配所导致)；

[0164] 用于经由采用块对角化(BD)的DIDO-OFDM系统中的ZF接收机来消除ICI(因I/Q不匹配所导致)的技术；

[0165] 用于经由DIDO-OFDM系统中的预编码(在发射机处)和ZF或MMSE滤波器(在接收机处)来消除用户间干扰和ICI(因I/Q不匹配所导致)的技术；

[0166] 用于经由DIDO-OFDM系统中的预编码(在发射机处)和类似于最大似然(ML)检测器的非线性检测器(在接收机处)来消除用户间干扰和ICI(因I/Q不匹配所导致)的技术；

[0167] 使用基于信道状况信息的预编码以用于消除OFDM系统中来自镜像调的载波间干扰(ICI)(因I/Q不匹配所导致)；

[0168] 使用基于信道状况信息的预编码以用于消除DIDO-OFDM系统中来自镜像调的载波间干扰(ICI)(因I/Q不匹配所导致)；

[0169] 在基站处使用I/Q不匹配已知DIDO预编码器(I/Q mismatch aware DIDO precoder)，以及在用户终端处使用I/Q已知DIDO接收机；

[0170] 在基站处使用I/Q不匹配已知DIDO预编码器(I/Q mismatch aware DIDO precoder)，在用户终端处使用I/Q已知DIDO接收机，以及使用I/Q已知信道估计器；

[0171] 在基站处使用I/Q不匹配已知DIDO预编码器，在用户终端处使用I/Q已知DIDO接收机，以及使用I/Q已知信道估计器和I/Q已知DIDO反馈生成器(该生成器将信道状况信息从用户终端发送至站点)；

[0172] 在基站处使用I/Q不匹配已知DIDO预编码器，以及使用I/Q已知DIDO配置器(该配置器使用I/Q信道信息来执行各种功能，包括用户选择、自适应编码和调制、空时频映射或预编码器选择)；

[0173] 使用I/Q已知DIDO接收机(该接收机经由采用块对角化(BD)预编码器的DIDO-OFDM系统中的ZF接收机来消除ICI(因I/Q不匹配所导致))；

[0174] 使用I/Q已知DIDO接收机(该接收机经由DIDO-OFDM系统中的预编码(在发射机处)和类似于最大似然(ML)检测器的非线性检测器(在接收机处)来消除用户间干扰和ICI(因I/Q不匹配所导致))，以及

[0175] 使用I/Q已知DIDO接收机(该接收机经由DIDO-OFDM系统中的ZF或MMSE滤波器来消除ICI(因I/Q不匹配所导致))。

[0176] a、背景

[0177] 典型无线通信系统的发送和接收信号包含同相正交(I/Q)分量。在实际的系统中，该同相正交分量可能会由于混频和基带操作中的缺陷而失真。这些失真(distortion)表现为I/Q相位、增益和延迟不匹配。相位不平衡是由调制器/解调器中的正弦(sine)和余弦(cosine)并未正确正交而导致的。增益不平衡是由同相正交分量之间的不同增幅而导致的。由于模拟电路中的I和Q轨道(rail)之间的延迟不同，还可能存在附加失真，该失真称之为延迟不平衡。

[0178] 在正交频分复用(OFDM)系统中，I/Q不平衡会导致来自发射调的载波间不平衡(ICI)。该影响已在一些资料中得到了研究，且在以下资料中，已提出了用于对单输入单输

出SISO-OFDM系统中的I/Q不匹配进行补偿的方法:M.D.Benedetto和P.Mandarini,"Analysis of the effect of the I/Q baseband filter mismatch in an OFDM modem,"Wireless personal communications,pp.175-186,2000;S.Schuchert和R.Hasholzner,"A novel I/Q imbalance compensation scheme for the reception of OFDM signals,"IEEE Transaction on Consumer Electronics,Aug.2001;M.Valkama,M.Renfors和V.Koivunen,"Advanced methods for I/Q imbalance compensation in communication receivers,"IEEE Trans.Sig.Proc,Oct.2001;R.Rao和B.Daneshrad,"Analysis of I/Q mismatch and a cancellation scheme for OFDM systems,"IST Mobile Communication Summit,June2004;A.Tarighat,R.Bagheri和A.H.Sayed,"Compensation schemes and performance analysis of IQ imbalances in OFDM receivers,"Signal Processing,IEEE TransactionS on[还可参见Acoustics,Speech,and Signal Processing,IEEE TransactionS on],vol.53,pp.3257-3268,Aug.2005。

[0179] 以下资料中示出了该工作向多输入多输出MIMO-OFDM系统的扩展:R.Rao和B.Daneshrad,"I/Q mismatch cancellation for MIMO OFDM systems,"in Personal,Indoor and Mobile Radio CommunicationS,2004;PIMRC2004.15th IEEE International Symposium on,vol.4,2004,pp.2710-2714.对于空间复用(SM),请参见R.M.Rao,W.Zhu,S.Lang,C.Oberli,D.Browne,J.Bhatia,J.F.Frigon,J.Wang,P;Gupta,H.Lee,D.N.Liu,S.G. Wong,M.Fitz,B.Daneshrad,和O.Takeshita,"Multiantenna testbeds for research and education in wireless communications,"IEEE Communications Magazine,vol.42,no.12,pp.72-81,Dec.2004;S.Lang,M.R.Rao和B.Daneshrad,"Design and development of a5.25GHz software defined wireless OF DM communication platform,"IEEE Communications Magazine,vol.42,no.6,pp.6-12,June2004;对于正交空时分组码(OSTBC),请参见A.Tarighat和A.H.Sayed,"MIMO OFDM receivers for systems with IQ imbalances,"IEEE Trans.Sig.Proc,vol.53,pp.3583-3596,Sep.2005。

[0180] 不幸的是,目前并不存在介绍如何对分布式输入分布式输出(DIDO)通信系统中的I/Q增益和相位不平衡误差进行校正的资料。以下所述的本发明实施方式提供了一种解决这些问题的方案。

[0181] DIDO系统包括一具有分布式天线的基站,该基站在利用相同于传统SIO系统的无线资源(即,相同的时隙持续时间和频带),发送并行数据流(经预编码的)至多个用户,以增强下行链路吞吐量。S.G. Perlman和T.Cotter 于2004年7月30提交的题为"System and Method for Distributed Input-Distributed Output Wireless Communications"的申请No.10/902,978("在先申请")给出了DIDO系统的详细说明,该申请被转让给了本申请的受让人,且该申请作为参考而被结合于此。

[0182] 存在多种方式实现DIDO预编码器。一种方案是以下资料中所描述的块对角化(BD):Q.H.Spencer,A.L.Swindlehurst和M.Haardt,"Zero forcing methods for downlink spatial multiplexing in multiuser MIMO channels,"IEEE Trans.Sig.Proc,vol.52,pp.461-471,Feb.2004;K.K.Wong,R.D.Murch,和K.B.letaief,"A joint channel diagonalization for multiuser MIMO antenna systems,"IEEE Trans.Wireless Comm.,vol.2,pp.773-786,Jul2003;L.U.Choi和R.D.Murch,"A transmit

preprocessing technique for multiuser MIMO systems using a decomposition approach,"IEEE Trans.Wireless Comm.,vol.3,pp.20-24,Jan2004;Z.Shen,J.G.Andrews,R.W.Heath和B.L.Evans,"Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization,"被接受发表在IEEE Trans.Sig.Proc.Sep.2005;Z.Shen,R.Chen,J.G.Andrews,R.W.Heath和B.L.Evans,"Sum capacity of multiuser MIMO broadcast channels with block diagonalization,"被提交至IEEE Trans.Wireless Comm.,Oct.2005;R.Chen,R.W. Heath,和J.G. Andrews,"Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers,"被接受至IEEE Trans.on Signal Processing,2005。这些材料中所给出的用于I/Q补偿的方法设想了BD预编码器,且该预编码器可被扩展至DIDO预编码的任何类型。

[0183] 在DIDO-OFDM系统中,I/Q不匹配会导致两种影响:ICI和用户间干扰。与SISO-OFDM系统中相类似,前者是由于来自镜像调的干扰造成的。后者是由于以下事实引起的,即I/Q不匹配会破坏DIDO预编码器的正交,从而在用户之间产生干扰。可通过在此所述的方法,在发射机和接收机处消除此两类干扰。在此描述了三种用于DIDO-OFDM系统中的I/Q补偿的方法,且针对具有和不具有I/Q不匹配,比较了他们的性能。基于利用DIDO-OFDM原型所执行的仿真和实际测量,展示了结果。

[0184] 本实施方式是在先申请的扩展。特别地,这些实施方式与在先申请的以下特征有关:

[0185] 在先申请中所描述的系统,其中I/Q轨道会受到增益和相位不平衡的影响;

[0186] 在发射机处,使用针对信道估计所采用的训练信号来计算具有I/Q补偿的DIDO预编码器;以及

[0187] 信号特征数据考虑到了由于I/Q不平衡所导致的失真,且在发射机处,根据本材料所提出的方法,使用该信号特征数据来计算DIDO预编码器。

[0188] b、本发明的实施方式

[0189] 首先,将描述本发明的数学模型和架构。

[0190] 在提示本方案之前,解释核心数学概念是非常有用的,我们通过假设I/Q增益和相位不平衡(本描述中并未包含相位延迟,但该相位延迟将在DIDO-OFDM形式的算法中被自动处理)来对其进行解释。为解释基本思想,假设我们想将两个复数s=sI+jsQ和h=hI+jhQ相乘,且使得x=h*s。我们使用下标来代表同相正交分量。调用以下等式:

[0191] $x_I=s_Ih_I-s_Qh_Q$

[0192] 及

[0193] $x_Q=s_Ih_Q+s_Qh_I$

[0194] 其矩阵形式可重写为:

[0195] $$\begin{bmatrix} x_I \\ x_Q \end{bmatrix}=\begin{bmatrix} h_I & -h_Q \\ h_Q & h_I \end{bmatrix}\begin{bmatrix} s_I \\ s_Q \end{bmatrix}$$

[0196] 通过信道矩阵OD来标识归一化变换。现假设s为所发送的符号,且h为信道。可通过创建以下非归一化变换来对I/Q增益和相位不平衡的存在进行建模:

[0197] $$\begin{bmatrix} x_I \\ x_Q \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} s_I \\ s_Q \end{bmatrix} \quad (A)$$

[0198] 该技巧的作用是确认可写为：

[0199] $$\begin{bmatrix} h_{11} & -h_{12} \\ h_{21} & h_{22} \end{bmatrix} = \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & h_{12}+h_{21} \\ h_{12}+h_{21} & h_{22}-h_{11} \end{bmatrix}$$

[0200] $$= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix}$$

[0201] 现对(A)进行重写：

[0202] $$\begin{bmatrix} x_I \\ x_Q \end{bmatrix} = \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix}\begin{bmatrix} s_I \\ s_Q \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix}\begin{bmatrix} s_I \\ s_Q \end{bmatrix}$$

[0203] $$= \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix}\begin{bmatrix} s_I \\ s_Q \end{bmatrix} + \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}-h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix}\begin{bmatrix} s_I \\ -s_Q \end{bmatrix}$$

[0204] 我们进行以下定义：

[0205] $$H_e = \frac{1}{2}\begin{bmatrix} h_{11}+h_{22} & h_{12}-h_{21} \\ -(h_{12}-h_{21}) & h_{11}+h_{22} \end{bmatrix}$$

[0206] 及

[0207] $$H_c = \frac{1}{2}\begin{bmatrix} h_{11}-h_{22} & -(h_{12}+h_{21}) \\ h_{12}+h_{21} & h_{11}-h_{22} \end{bmatrix}$$

[0208] 这两个矩阵具有归一化结构，因此可被表示为复数形式：

[0209] $h_e=h_{11}+h_{22}+j(h_{21}-h_{12})$

[0210] 及

[0211] $h_c=h_{11}-h_{22}+j(h_{21}+h_{12})$

[0212] 通过使用所有这些知识，我们可将有效等式推导回具有两个信道(等价信道h_e和共轭信道h_c)的标量形式。因此，(5)中的有效变换变为：

[0213] $x=h_e s+h_c s^*$

[0214] 我们将第一信道称为等价信道，第二信道称为共轭信道。如果不存在I/Q增益和相位不平衡，则该等价信道即为我们所要观察的信道。

[0215] 通过使用相类似的论据，具有I/Q增益和相位不平衡的离散时间MIMO×M系统的输入-输出关系可示为(通过使用标量等价形式来建立他们的矩阵对应形式)：

[0216] $$\mathbf{x}[t] = \sum_{\ell=0}^{L} \mathbf{h}_e[\ell]\mathbf{s}[t-\ell] + \mathbf{h}_c[\ell]\mathbf{s}^*[t-\ell]$$

[0217] 其中，t为离散时间指数，$\mathbf{h}_e, \mathbf{h}_c \in \mathbb{C}^{M\times N}$，$s=[s_1,\ldots,s_N]$，$x=[x_1,\ldots,x_M]$且L为信道

抽头(channel tap)数。

[0218] 在DIDO-OFDM系统中，表示了频域中所接收的信号。如果满足以下等式，则从信号和系统重新调用：

[0219] $\mathrm{FFT}_K\{s[i]\}=S[k]$ 则 $\mathrm{FFT}_K\{s^*[i]\}=S^*[(-k)]=S^*[K-k]$ for $k=0,1,\ldots,K-1$

[0220] 利用OFDM，对于副载波k，MIMO-OFDM系统的等价输入-输出关系为：

[0221] $$\bar{\mathbf{x}}[k]=\mathbf{H}_e[k]\bar{\mathbf{s}}[k]+\mathbf{H}_c[k]\bar{\mathbf{s}}^*[K-k] \quad (1)$$

[0222] 其中，$k=0,1,\ldots,K-1$为OFDM副载波索引，H_e和H_c分别代表等价和共轭信道矩阵，定义如下：

[0223] $$\mathbf{H}_e[k]=\sum_{l=0}^{L}\mathbf{h}_e[l]e^{-j\frac{2\pi lk}{K}}$$

[0224] 及

[0225] $$\mathbf{H}_c[k]=\sum_{l=0}^{L}\mathbf{h}_c[l]e^{-j\frac{2\pi lk}{K}}$$

[0226] (1)中的第二基值为来自镜像调的干扰。可通过构建以下堆式(stacked)矩阵系统(请仔细注意共轭值)来对其进行处理：

[0227] $$\begin{bmatrix}\bar{\mathbf{x}}[k]\\ \bar{\mathbf{x}}^*[K-k]\end{bmatrix}=\begin{bmatrix}\mathbf{H}_e[k] & \mathbf{H}_c[k]\\ \mathbf{H}_c^*[K-k] & \mathbf{H}_e^*[K-k]\end{bmatrix}\begin{bmatrix}\bar{\mathbf{s}}[k]\\ \bar{\mathbf{s}}^*[K-k]\end{bmatrix}$$

[0228] 其中$\bar{\mathbf{s}}=[s_1,s_2]$和$\bar{\mathbf{x}}=[x_1,x_2]$分别为发送和接收符号在频域中的向量。

[0229] 通过使用该方法，可建立有效矩阵，以用于DIDO操作。例如，利用DIDO2×2输入-输出关系(假设每个用户具有单个接收天线)，第一用户设备可考虑以下等式(在不存在噪声时)：

[0230] $$\begin{bmatrix}\bar{x}_1[k]\\ \bar{x}_1^*[K-k]\end{bmatrix}=\begin{bmatrix}\mathbf{H}_e^{(1)}[k] & \mathbf{H}_c^{(1)}[k]\\ \mathbf{H}_c^{(1)*}[K-k] & \mathbf{H}_e^{(1)*}[K-k]\end{bmatrix}\mathbf{W}\begin{bmatrix}\bar{s}_1[k]\\ \bar{s}_1^*[K-k]\\ \bar{s}_2[k]\\ \bar{s}_2^*[K-k]\end{bmatrix} \quad (2)$$

[0231] 而第二用户注意以下等式：

[0232] $$\begin{bmatrix}\bar{x}_2[k]\\ \bar{x}_2^*[K-k]\end{bmatrix}=\begin{bmatrix}\mathbf{H}_e^{(2)}[k] & \mathbf{H}_c^{(2)}[k]\\ \mathbf{H}_c^{(2)*}[K-k] & \mathbf{H}_e^{(2)*}[K-k]\end{bmatrix}\mathbf{W}\begin{bmatrix}\bar{s}_1[k]\\ \bar{s}_1^*[K-k]\\ \bar{s}_2[k]\\ \bar{s}_2^*[K-k]\end{bmatrix} \quad (3)$$

[0233] 其中，$\mathbf{H}_e^{(m)},\mathbf{H}_c^{(m)}\in C^{1\times 2}$分别代表矩阵$H_e$和$H_c$的第m行，且$\mathbf{W}\in C^{4\times 4}$为DIDO预编码矩阵。根据(2)和(3)，可注意到用户m所接收的符号$\bar{x}_m[k]$受1/0不平衡所导致的两个干扰源(即，来自镜像调的载波间干扰(即，$\bar{s}_m[K-k]$)以及用户间干扰(即，$\bar{s}_p[k]$以及$\bar{s}_p^*[K-k]$

,p≠m))的影响。(3)中的DIDO预编码矩阵W被设计成用于消除这两个干扰项。

[0234] 可用于此处的DIDO预编码器存在多个不同的实施方式,这取决于接收机处所应用的联合检测。在一种实施方式中,可采用根据合成信道$\left[\mathbf{H}_1^{(eq)},\mathbf{H}_2^{(eq)}\right]$(而非$\mathbf{H}_j^{(eq)}$)所计算的块对角化(BD)(请参见例如,Q.H.Spencer,A.L.Swindlehurst,和M.Haardt,"Zeroforcing methods for downlink spatial multiplexing in multiuser MIMO channels,"IEEE Trans.Sig.Proc,vol.52,pp.461-471,Feb.2004.K.K.Wong,R.D.Murch,和K.B.Letaief,"A joint channel diagonalization for multiuser MIMO antenna systems,"IEEE Trans.Wireless Comm.,vol.2,pp.773-786,jul2003;L. U.Choi和R.D.Murch,"A transmit preprocessing technique for multiuser MIMO systems using a decomposition approach,"IEEE Trans.Wireless Comm.,vol.3,pp.20-24,jan2004;Z.Shen,J.G. Andrews,R.W. Heath,和B.L Evans,"Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization,"被接受发表在IEEE Trans.Sig.Proc,Sep.2005;Z.Shen,R.Chen,J.G. Andrews,R.W. Heath,和B.L Evans,"Sum capacity of multiuser MIMO broadcast channels with block diagonalization,"被提交至IEEE Trans.Wireless Comm.,(Oct.2005),因此,目前DIDO系统选择预编码器,以使得:

[0235] $$\tilde{H}_{eq}\triangleq\begin{bmatrix}\mathbf{H}_1^{(1)}[k] & \mathbf{H}_1^{(2)}[k]\\ \mathbf{H}_1^{(1)}[K-k] & \mathbf{H}_1^{(2)}[K-k]\\ \mathbf{H}_2^{(1)}[k] & \mathbf{H}_2^{(2)}[k]\\ \mathbf{H}_2^{(1)}[K-k] & \mathbf{H}_2^{(2)}[K-k]\end{bmatrix}\mathbf{W}=\begin{bmatrix}\alpha_{11} & 0 & 0 & 0\\ 0 & \alpha_{22} & 0 & 0\\ 0 & 0 & \alpha_{33} & 0\\ 0 & 0 & 0 & \alpha_{44}\end{bmatrix}\Delta\begin{bmatrix}H_{eq}^{(1,1)} & H_{eq}^{(1,2)}\\ H_{eq}^{(2,1)} & H_{eq}^{(2,2)}\end{bmatrix} \quad (4)$$

[0236] 其中,$\alpha_{i,i}$为常数,且$H_{eq}^{(j,i)}\in C^{k\times 2}$。该方法是非常有益的,因为通过使用该预编码器,由于在发射机处完全消除了I/Q增益和相位不平衡的影响可使DIDO预编码器的其他方面保持原样。

[0237] 还可将DIDO预编码器设计为预先消除用户间干扰,而不预先消除因IQ不平衡所导致的ICI。利用该方法,接收机(而非发射机)可通过采用以下所述的接收滤波器之一来对IQ不平衡进行补偿。因此,(4)中的预编码设计标准可被修改为:

[0238] $$\tilde{H}_{eq}\triangleq\begin{bmatrix}\mathbf{H}_1^{(1)}[k] & \mathbf{H}_1^{(2)}[k]\\ \mathbf{H}_1^{(1)}[K-k] & \mathbf{H}_1^{(2)}[K-k]\\ \mathbf{H}_2^{(1)}[k] & \mathbf{H}_2^{(2)}[k]\\ \mathbf{H}_2^{(1)}[K-k] & \mathbf{H}_2^{(2)}[K-k]\end{bmatrix}\mathbf{W}=\begin{bmatrix}\alpha_{11} & \alpha_{12} & 0 & 0\\ \alpha_{21} & \alpha_{22} & 0 & 0\\ 0 & 0 & \alpha_{33} & \alpha_{34}\\ 0 & 0 & \alpha_{43} & \alpha_{44}\end{bmatrix}\Delta\begin{bmatrix}H_{eq}^{(1,1)} & H_{eq}^{(1,2)}\\ H_{eq}^{(2,1)} & H_{eq}^{(2,2)}\end{bmatrix} \quad (5)$$

[0239] $$\bar{\mathbf{x}}_1[k]=\begin{bmatrix}H_{eq}^{(1,1)} & H_{eq}^{(1,2)}\end{bmatrix}\begin{bmatrix}\bar{\mathbf{s}}_1[k]\\ \bar{\mathbf{s}}_2[k]\end{bmatrix} \quad (6)$$

[0240] 及

[0241] $$\bar{\mathbf{x}}_2[k]=\left[H_{uc}^{(2,1)}\ H_{uc}^{(2,2)}\right]\begin{bmatrix}\bar{\mathbf{s}}_1[k]\\ \bar{\mathbf{s}}_2[k]\end{bmatrix} \quad (7)$$

[0242] 其中对于第m个发送信号而言，$\bar{\mathbf{s}}_m[k]=[\bar{s}_m[k]\ \bar{s}_m[K-k]]^T$，且$\bar{\mathbf{x}}_m[k]=[\bar{x}_m[k]\ \bar{x}_m[K-k]]^T$为用户m所接收的符号向量。

[0243] 在接收侧，为了对发送符号向量$\bar{\mathbf{s}}_m[k]$进行估计，用户m采用ZF滤波器，且所估计的符号向量被给定为：

[0244] $$\hat{\mathbf{s}}_m^{(ZF)}[k]=[(H_{uc}^{(m,m)H}H_{uc}^{(m,m)})^{-1}H_{uc}^{(m,m)H}]\bar{\mathbf{x}}_m[k] \quad (8)$$

[0245] 虽然ZF滤波器最易于理解，但接收机还可应用任意数量的本领域技术人员所公知的其他滤波器。一种大众选择为MMSE滤波器，其中：

[0246] $$\hat{\mathbf{s}}_m^{(MMSE)}[k]=(H_{uc}^{(m,m)H}+\mathrm{p}^{-1}I)^{-1}H_{uc}^{(m,m)H}\bar{\mathbf{x}}_m[k] \quad (9)$$

[0247] 且p为信噪比。可选地，用户可执行最大似然符号检测(或者球解码器、迭代变化)。例如，第一用户可使用ML接收机，并求解以下优化：

[0248] $$\hat{\mathbf{s}}_m^{(ML)}[k]=\arg\min_{\mathbf{s}_1,\mathbf{s}_2\in S}\left\|\bar{\mathbf{y}}_1[k]-\left[H_{uc}^{(1,1)}\ H_{uc}^{(1,2)}\right]\begin{bmatrix}\mathbf{s}_1[k]\\ \mathbf{s}_2[k]\end{bmatrix}\right\| \quad (10)$$

[0249] 其中，S为所有可能的向量s的集合，且取决于星座图大小。该ML接收机给出可较好的性能，但在接收机处要求更高的复杂度。类似的一组等式可应用于第二用户。

[0250] 注意，(6)和(7)中的$H_{uc}^{(1,2)}$和$H_{uc}^{(2,1)}$被假设为具有零项，该假设仅在发射预编码器能够完全消除针对(4)中的标准的用户间干扰的情况下有效。类似的，$H_{uc}^{(1,2)}$和$H_{uc}^{(2,1)}$仅在发射预编码器能够完全消除载波间干扰(即，来自镜像调)的情况下为对角矩阵。

[0251] 图13显示了具有I/Q补偿的DIDO-OFDM系统的架构的一种实施方式，所述DIDO-OFDM系统包括位于基站(BS)内的IQ-DIDO预编码器1302、发送信道1304、位于用户设备内的信道估计逻辑1306以及ZF、MMSE或ML接收机1308。所述信道估计逻辑1306经由训练信号而对信道$H_{uc}^{(m,m)}$和$H_{uc}^{(m,m)}$进行估计，并将这些估计反馈至AP内的预编码器。BS计算DIDO预编码器权重(矩阵W)，以预先消除因I/Q增益和相位不平衡所导致的干扰以及用户干扰，并将数据通过无线信道1304发送至用户。用户设备m采用ZF、MMSE或ML接收机1308。通过利用单元1301所提供的信道估计来消除剩余干扰，并对数据进行解调。

[0252] 可采用以下三个实施方式来实现这I/Q补偿算法。

[0253] 方法1-TX补偿：在该实施方式中，发射机根据(4)中的标准来计算预编码矩阵。在接收机处，用户设备采用"简化的"ZF接收机，其中$H_{uc}^{(1,1)}$和$H_{uc}^{(2,2)}$被假设为对角矩阵。因此，公式(8)简化为：

[0254] $$\hat{\mathbf{s}}_m[k]=\begin{bmatrix}1/\alpha_{m,1} & 0\\ 0 & 1/\alpha_{m,2}\end{bmatrix}\bar{\mathbf{x}}_m[k] \quad (10)$$

[0255] 方法2-RX补偿：在该实施方式中，发射机基于R.Chen,R.W.Heath,and J.G.Andrews,"Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers,"accepted to IEEE Trans.on Signal Processing,2005中描述的传统BD方法，计算预编码矩阵，且不针对(4)中的标准来消除载波间和用户间干扰。利用该方法，(2)和(3)中的预编码矩阵简化为：

[0256] $$\boldsymbol{W}=\begin{bmatrix} w_{1,1}[k] & 0 & w_{1,2}[k] & 0 \\ 0 & w_{2,1}^{*}[K-k] & 0 & w_{2,2}^{*}[K-k] \\ w_{2,1}[k] & 0 & w_{2,2}[k] & 0 \\ 0 & w_{1,1}^{*}[K-k] & 0 & w_{1,2}^{*}[K-k] \end{bmatrix} \quad (12)$$

[0257] 在接收机处，用户设备如(8)中那样采用ZF被动器。注意，该方法并不如上述方法1那样，在发射机处预先消除干扰。因此，其在接收机处消除载波间干扰，但并不能消除用户间干扰。此外，相比于方法1要求反馈$\boldsymbol{H}_{2,2}^{[m]}$和$\boldsymbol{H}_{2,1}^{[m]}$，在方法2中，用户仅需要反馈针对发射机的向量$\boldsymbol{H}_{2}^{[m]}$，以计算DIDO预编码器。因此，方法2特别适于具有低速率反馈信道的DIDO系统。另一方面，方法2需要用户设备处具有稍微较高的计算复杂度，以在(8)(而非(11))中计算ZP接收机。

[0258] 方法3-TX-RX补偿：在一种实施方式中，将上述两个方法合并。发射机如(4)那样计算预编码矩阵，而接收机根据(8)来对发送符号进行估计。

[0259] I/Q不平衡(无论是相位不平衡、增益不平衡，抑或是延迟不平衡)会对无线通信系统中的信号质量造成有害的降级。针对此原因，以往的电路均被设计成具有较低的不平衡。然而，如上所述，可通过使用发射预编码形式的数字信号处理和/或特定接收机，修正该问题。本发明的一种实施方式包括具有多个新功能单元的系统，每个单元对于实现OFDM通信系统或DIDO-OFDM通信系统中的I/Q校正均是很重要的。

[0260] 本发明的一种实施方式使用基于信道状况信息的预编码，以消除OFDM系统中来自镜像调的载波间干扰(ICI)(因I/Q不匹配导致)。如图11所示，根据本实施方式的DIDO发射机包括用户选择器单元1102、多个编码调制单元1104、对应的多个映射单元1106、DIDO IQ已知预编码单元1108、多个RF发射机单元1114、用户反馈单元1112以及DIDO配置器单元1110。

[0261] 所述用户选择器单元1102基于反馈单元1112所获取的反馈信息，选择与多个用户U_1-U_K相关联的数据，并将该信息提供给多个编码调制单元1104中的每个编码调制单元1104。每个编码调制单元1104对每个用户的信息比特进行编码和解调，并将它们发送至映射单元1106。该映射单元1106将输入比特映射至复数符号，并将结果发送至DIDO IQ已知预编码单元1108。该DIDO IQ已知预编码单元1108利用反馈单元1112从用户获取的信道状况信息，计算DIDO IQ已知预编码权重，并对从映射单元1106获取的输入符号进行预编码。每一个预编码数据流均由DIDO IQ已知预编码单元1108发送至OFDM单元1115，该OFDM单元1115计算IFFT，并加入循环前缀。该信息被发送至D/A单元1116，该D/A单元1116进行数模转换，并将其发送至RF单元1114。该RF单元1114将基带信号升频至中频/射频，并将其发送至发射天线。

[0262] 所述预编码器对常规调和镜像调一起进行操作，以补偿I/Q不平衡。可使用任意数量的预编码器设计标准，包括ZF、MMSE或加权MMSE设计。在优选实施方式中，预编码器可完全移除因I/Q不匹配所导致的ICI，从而使得接收机不需要执行任何附加补偿。

[0263] 在一种实施方式中，所述预编码器使用块对角化标准，以在不完全消除每一用户的I/Q影响(这需要附加接收机处理)的情况下，完全消除用户间干扰。在另一实施方式中，所述预编码器使用零强制标准来完全消除因I/Q不平衡所导致的用户间干扰以及ICI干扰。该实施方式可在接收机处使用传统的DIDO-OFDM处理器。

[0264] 本发明的一种实施方式使用基于信道状况信息的预编码，以消除DIDO-OFDM系统中来自镜像调的载波间干扰(ICI)(因I/Q不匹配所导致)，且每一用户采用IQ已知DIDO接收机。如图12所示，在本发明的一种实施方式中，系统(包括接收机1202)包括多个RF单元1208、相应地多个A/D单元1210、IQ已知信道估计器单元1204以及DIDO反馈生成器单元1206。

[0265] 所述RF单元1208接收从DIDO发射机单元1114发送的信号，将该信号降频至基带，并将该降频后的信号提供给A/D单元1210。之后，该A/D单元1210对该信号进行模数转换，并将其发送至OFDM单元1213，该OFDM单元1213移除循环前缀，并进行FFT，以将该信号报告至频域。在训练周期期间，OFDM单元1213将输出发送至IQ已知信道估计单元1204，该IQ已知信道估计单元1204在频域中计算信道估计。可选地，可在时域中计算所述信道估计。在数据周期(data period)期间，OFDM单元1213将输出发送至IQ已知接收机单元1202。该IQ已知接收机单元计算IQ接收机，并对所述信号进行解调/解码，以获取数据1214。所述IQ已知信道估计单元1204发送所述信道估计至DIDO反馈生成器单元1206，该反馈生成器单元1204可对所述信道估计进行量化，并经由反馈控制信道1112而将其发回发射机。

[0266] 图12所示的接收机1202可在任意数量的本领域技术人员所公知的标准(包括ZF、MMSE、最大似然或MAP接收机)下工作。在一优选实施方式中，接收机使用MMSE滤波器来消除因镜像调上的IQ不平衡所导致的ICI。在另一优选实施方式中，接收机使用类似于最大似然搜索的非线性检测器来联合检测镜像调上的符号。该方法具有良好的性能，但具有更高的复杂度。

[0267] 在一种实施方式中，使用IQ已知信道估计器1204来确定接收机系数，以移除ICI。因此，我们要求了DIDO-OFDM系统(使用基于信道状况信息的预编码来消除来自镜像调的载波间干扰(ICI)(因I/Q不匹配所导致))、IQ已知DIDO接收机以及IQ已知信道估计器的权益。所述信道估计器可使用传统的训练信号，或可使用在同相正交信号上发送的专门构建的训练信号。可实施任意数量的估计算法，包括最小二乘法、MMSE或最大似然。所述IQ已知信道估计器为IQ已知接收机提供输入。

[0268] 信道状况信息可通过信道互易性或通过反馈信道而被提供给站点。本发明的一实施方式包括DIDO-OFDM系统，该系统具有I/Q已知预编码器，以及用于将来自用户终端的信道状况信息传输至站点的I/Q已知反馈信道。该反馈信道可为物理或逻辑控制信道。其可在随机访问信道中被专用或共享。可通过使用用户终端(我们也要求了该用户终端的权益)处的DIDO反馈生成器来生成反馈信息。所述DIDO反馈生成器将所述I/Q已知信道估计器的输出作为输入，其可量化信道系数，或可使用任意数量本领域所公知的有限反馈算法。

[0269] 用户的分配、调制及编码率、至空时频编码时隙的映射可根据所述DIDO反馈生成

器的结果而变化。因此，一实施方式包括IQ已知DIDO配置器，该配置器使用来自一个或多个用户的IQ已知信道估计来配置DIDO IQ已知预编码器，选择调制率、编码率、允许发送的用户的子集、以及他们的空空时频编码时隙的映射。

[0270] 为了评价所提出的补偿方法的性能，将比较三个DIDO2×2系统：

[0271] 1、具有I/Q不匹配：通过所有的调(除了DC调和边缘调)进行发送，且不对I/Q不匹配进行补偿；

[0272] 2、具有I/Q补偿：通过所有的调进行发送，且通过使用上述"方法1"来对I/Q不匹配进行补偿；

[0273] 3、理想的：仅通过奇数个调进行发送，以避免用户间干扰以及因I/Q不匹配所导致的载波间(即，来自镜像调的)干扰。

[0274] 在此之后，展示了真实传播情形中利用DIDO-OFDM原型进行测量所获取的结果。图14给示了从上述三个系统所获取的64-QAM星座图。这些星座图是在同一用户位置以及固定平均信噪比(~45dB)的情况下获取的。第一星座图1401是非常嘈杂的(由于I/Q不平衡所导致来自镜像调的干扰)。第二星座图1402示出了一些改进(由于I/Q补偿)。注意，第二星座图1402并没有星座图1403所示的理想情况那样纯净(由于存在可能产生载波间干扰(ICI)的相位噪声)。

[0275] 图15示出了在具有和不具有I/Q不匹配的情况下，64-QAM和3/4编码率的DIDO2×2系统的平均SER(符号差错率)1501和每用户实际吞吐量(goodput)1502。OFDM带宽为250KHZ，具有64个调且循环前缀长度L_{cp}=4。由于在理想情况下，我们仅通过调的子集来发送数据，因此根据平均每调的发射功率(而非总的发射功率)来评价SER和实际吞吐量性能，以保证不同情况之间的公平比较。此外，在以下结果中，我们使用发射功率的归一化值(以分贝标示)，因为我们此处的目标是比较不同方案的相对(而非绝对)性能。图15示出了在存在I/Q不平衡的情况下，SER饱和且未达到目标SER(~10^{-3})，这与A.Tarighat and A.H.Sayed,"MIMO OFDM receivers for systems with IQ imbalances,"IEEE Trans.Sig.Proc,vol.53,pp.3583-3596,Sep.2005中报告的结果相一致。该饱和效应是由于以下事实导致的，即信号功率和干扰功率(来自镜像调的)随着TX功率的增大而增大。然而，通过所提出的I/Q补偿方法，可消除干扰，并获得较好地SER性能。注意，由于64-QAM调制需要较大的发射功率，因此，可因为DAC中的振幅饱和效应而导致SER在高SNR处会具有细微的增大。

[0276] 此外，可观察到，在存在I/Q补偿的情况下，SER性能非常接近理想情况。此两种情况之间，TX功率的2dB间隙是由于相位噪声(该相位噪声可能会在相邻OFDM调之间产生附加干扰)造成的。最后，实际吞吐量曲线1502示出了当应用I/Q方法时，其相比于理想情况可发送两倍的数据，因为我们使用了所有的数据调而非仅奇数调(针对理想情况)。

[0277] 图16图示了在具有I/Q补偿或不具有I/Q补偿的情况下，不同QAM星座图的SER性能。我们可观察到，在此实施方式中，所提出的方法对于64-QAM星座图而言是特别有利的。对于4-QAM和16-QAM而言，I/Q补偿方法会产生比具有I/Q不匹配的情况更差的性能，这可能是因为所提出的方法要求更大的功率来进行数据发送以及来自镜像调的干扰消除。此外，由于星座点之间的较大的最小距离，4-QAM和16-QAM并不如64-QAM那样受到I/Q不匹配的影响。参见A.Tarighat,R.Bagheri,和A.H.Sayed,"Compensation schemes and performance

analysis of IQ imbalances in OFDM receivers," Signal Processing, IEEE Transactions on[还可参见Acoustics, Speech, and Signal Processing, IEEE Transactions on], vol.53, pp.3257-3268, Aug.2005.还可观察图16并通过将I/Q不匹配与针对4-QAM和16-QAM的理想情况进行比较而得出该结论。因此，对于4-QAM和16-QAM的情况而言，具有干扰消除(来自镜像调的)的DIDO预编码器所需要的附加功率并不能为I/Q补偿的小小利益作保。注意，可通过采用上述I/Q补偿方法2和3来解决该问题。

[0278] 最后，在不同的传播情况下，测量了上述三个方法的相对SER性能。还描述了在存在I/Q不匹配的情况的SER性能，以供参考。图17给示了针对载波频率为450.5MHZ且带宽为250KHz的64-QAM DIDO2×2系统，在两个不同的用户位置所测量的SER。在位置1，用户与处于不同房间且处于NLOS(无视距)状态的BS相距~6λ。在位置2，用户与具有LOS(视距)的BS相距~λ。

[0279] 图17示出了所有三种补偿方法均比不进行补偿的情况表现突出。然而，应该注意的是，在任何信道情形下，方法3均胜过其他两种补偿方法。方法1和2的相对性能取决于传播情况。通过实际测量活动，可得出方法1大体上胜过方法2，因为其预先消除了(在发射机处)I/Q不平衡所导致的用户间干扰。当该用户间干扰很小时，如图17的曲线图1702所示，方法2可胜过方法1，因为其不会遭受因I/Q补偿预编码器所导致的功率损耗。

[0280] 到目前为止，已通过仅考虑有限组传播情形(如图17所示)而对不同方法进行了比较。在此之后，在理想i.i.d.(独立且具同分布的)信道中测量这些方法的相对性能。利用发射和接收侧的I/Q相位和增益不平衡来仿真DIDO-OFDM系统。图18示出了在仅发射机侧具有增益平衡的情况下(即，在第一发射链路的1轨上具有增益0.8，在其他轨上具有增益1)，所提出的方法的性能。可看出，方法3胜过了所有其他方法。此外，与图17的曲线图1702中位置2处获得结果相比，在i.i.d.信道中，方法1可比方法2更好地执行。

[0281] 因此，给出了三种新型方法来补偿上述DIDO-OFDM系统中的I/Q不平衡，方法3胜过所提出的其他补偿方法。在具有低速率反馈信道的系统中，可使用方法2来减小DIDO预编码所需的反馈量，但会导致较差的SER性能。

[0282] 11，自适应DIDO发送方案

[0283] 将描述用于增强分布式输入分布式输出(DIDO)系统的性能的系统和方法的另一实施方式。该方法通过跟踪变化的信道状态，动态地将无线资源分配给不同的用户设备，以在满足某些目标误码率的同时增大吞吐量。所述用户设备对信道质量进行估计，并将其反馈至基站(BS)；该基站对获取自用户设备的信道质量进行处理，以选择用于下一次发送的最佳用户设备集合、DIDO方案、调制/编码方案(MCS)以及阵列配置；所述基站经由预编码而将并行数据发送至多个用户设备，且信号在接收机处被解调。

[0284] 还描述一为DIDO无线链路有效分配资源的系统。该系统包括具有DIDO配置器的DIDO基站，该基站对接收自用户的反馈进行处理，以选择用于下一次发送的最佳用户集合，DIDO方案、调制/编码方案(MCS)以及阵列配置；DIDO系统中的接收机，该接收机对信道和其他相关参数进行测量，以生成DIDO反馈信号；以及DIDO反馈控制信道，用于将来自用户的反馈信息传输给基站。

[0285] 如以下所详述的，本发明该实施方式的一些显著特征可包括，但不限于：

[0286] 用于基于信道质量信息，自适应地选择用户数量、DIDO发送方案(即，天线选择或

复用)、调制/编码方案(MCS)以及阵列配置,以最小化SER,或最大化每用户的频谱效率或下行链路频谱效率的技术;

[0287] 用于定义多组DIDO发送模式以作为DIDO方案和MCS的组合的技术;

[0288] 用于根据信道状态将不同DIDO模式指派给不同的时隙、OFDM调和DIDO子流的技术;

[0289] 用于基于不同用户的信道质量将不同DIDO模式动态地指派给不同用户的技术;

[0290] 用于基于在时域、频域和空域中所计算的链路质量度量对自适应DIDO切换进行激活的标准;

[0291] 用于基于查找表对自适应DIDO切换进行激活的标准。

[0292] 如图19所示的在基站处具有DIDO配置器的DIDO系统,该系统可基于信道质量信息,自适应地选择用户数量、DIDO发送方案(即,天线选择或复用)、调制/编码方案(MCS)以及阵列配置,以最小化SER,或最大化每用户的频谱效率或下行链路频谱效率;

[0293] 如图20所示的在基站处具有DIDO配置器且在每个用户设备处具有DIDO反馈生成器的DIDO系统,该系统使用所估计的信道状况和/或接收机处的其他参数(类似于所估计的SNR),以生成输入至DIDO配置器的反馈消息。

[0294] DIDO系统,该系统具有DIDO配置器(在基站处)、DIDO反馈生成器以及DIDO反馈控制信道(该DIDO反馈信道用于将DIDO特定配置信息从用户传输至基站)。

[0295] a、背景

[0296] 在多输入多输出(MIMO)系统中,可构想分集方案(例如,正交空时分组码(OSTBC)(参见V.Tarokh,H.Jafarkhani,and A.R.Calderbank,"Spacetime block codes from orthogonal designs,"IEEE Trans.Info.Th.,vol.45,pp.1456-467,Jul.1999)或天线选择(参见R.W.Heath Jr.,S.Sandhu,and A.J.Paulraj,"Antenna selection for spatial multiplexing systems with linear receivers,"IEEE Trans.Comm.,vol.5,pp.142-144,Apr.2001),以防止信道衰减,提高链路可靠性(该可靠性可转换为更佳的覆盖率)。另一方面,空间复用(SM)可以以多个并行数据发送作为手段来增强系统吞吐量。参见G.J.Foschini,G.D.Golden,R.A.Valenzuela,和P.W.Wolniansky,"Simplified processing for high spectral effciency wireless communication employing multielement arrays,"IEEE Jour.Select.Areas in Comm.,vol.17,no.11,pp.1841-1852,Nov.1999。根据来源于L.Zheng和D.N.C.Tse,"Diversity and multiplexing:a fundamental tradeoff in multiple antenna channels,"IEEE Trans.Info.Th.,vol.49,no.5,pp.1073-1096,May2003的理论分集/复用折中,这些益处可在MIMO系统中同时实现。一实际实施形式为通过跟踪变化的信道状态,在分集和复用发送方案之间进行自适应切换。

[0297] 现已提出了大量自适应MIMO发送技术。R.W.Heath和A.J.Paulraj,"Switching between diversity and multiplexing in MIMO systems,"IEEE Trans.Comm.,vol.53,no.6,pp.962-968,Jun.2005中的分集/复用切换方法被设计成基于瞬时信道质量信息,改进针对固定速率发送的BER(比特误码率)。可选地,可如S.Catreux,V.Erceg,D.Gesbert,和R.W.Heath,Jr.,"Adaptive modulation and MIMO coding for broadband wireless data networks,"IEEE Comm.Mag.,vol.2,pp.108-115,June2002("Catreux")中那样,采用

统计信道信息来对自适应进行激活，从而减小反馈开销以及控制消息的数量。Catreux中的自适应发送算法被设计成基于信道时/频选择指示符，针对正交频分复用(OFDM)系统中的预定目标误码率，增强频谱效率。还针对窄带系统，提出了类似的低反馈自适应方法，该方法利用信道空间选择性来在分集方案与空间复用之间进行切换。参见例如A.Forenza，M.R.McKay，A.Pandharipande，R.W.Heath，Jr.，和I.B.Collings，"Adaptive MIMO transmission for exploiting the capacity of spatially correlated channels，" accepted to the IEEE Trans.on Veh.Tech.，Mar.2007；M.R.McKay，I.B.Collings，A.Forenza，and R.W.Heath，Jr.，"Multiplexing/beamforming switching for coded MIMO in spatially correlated Rayleigh channels，"被接受至IEEE Trans.on Veh.Tech.，Dec.2007；A.Forenza，M.R.McKay，R.W.Heath，Jr.，和I.B.Collings，"Switching between OSTBC and spatial multiplexing with linear receivers in spatially correlated MIMO channels，"Proc.IEEE Veh.Technol.Conf.，vol.3，pp.1387-1391，May2006；M.R.McKay，I.B.Collings，A.Forenza，和R.W.Heath Jr.，"A throughput-based adaptive MIMO BICM approach for spatially correlated channels，"出现在Proc.IEEE ICC，June2006。

[0298] 在该资料中，我们将各种先前公开中所展现的工作范围扩展至DIDO-OFDM系统。参见例如R.W.Heath和A.J.Paulraj，"Switching between diversity and multiplexing in MIMO systems，"IEEE Trans.Comm.，vol.53，no.6，pp.962-968，Jun.2005；S.Catreux，V.Erceg，D.Gesbert，和R.W.Heath Jr.，"Adaptive modulation and MIMO coding for broadband wireless data networks，"IEEE Comm.Mag.，vol.2，pp.108-115，June2002；A.Forenza，M.R.McKay，A.Pandharipande，R.W.Heath Jr.，和I.B.Collings，"Adaptive MIMO transmission for exploiting the capacity of spatially correlated channels，"IEEE Trans.on Veh.Tech.，vol.56，n.2，pp.619-630，Mar.2007；M.R.McKay，I.B.Collings，A.Forenza，和R.W.Heath Jr.，"Multiplexing/beamforming switching for coded MIMO in spatially correlated Rayleigh channels."被接受至IEEE Trans，on Veh.Tech.，Dec.2007；A.Forenza，M.R.McKay，R.W.Heath Jr.，和I.B.Collings，"Switching between OSTBC and spatial multiplexing with linear receivers in spatially correlated MIMO channels，"Proc.IEEE Veh.Technol.Conf.，vol.3，pp.1387-1391，May2006；M.R.McKay，I.B.Collings，A.Forenza，和R.W.Heath Jr.，"A throughput-based adaptive MIMO BICM approach for spatially correlated channels，"出现在Proc.IEEE ICC，June2006。

[0299] 在此描述了新型自适应DIDO发送策略，该策略以基于信道质量信息在不同数量的用户、不同数量的发射天线以及发送方案之间进行切换作为一种手段来改进系统性能。注意，M.Sharif和B.Hassibi，"On the capacity of MIMO broadcast channel with partial side information，"IEEE Trans.Info.Th.，vol.51，p.506522，Feb.2005以及W.Choi，A.Forenza，J.G.Andrews，和R.W.Heath Jr.，"Opportunistic space division multiple access with beam selection，"出现在IEEE Trans.on Communications中已提出了在多用户MIMO系统中自适应选择用户的方案。然而，这些公开中的机会性的(opportunistic)空分复用接入(OSDMA)方案被设计成通过利用多用户分集来最大化总的

PART 2 of CN103501193B

(Presented in 2 parts)

容量，且他们仅能够实现脏纸(dirty paper)码的部分理论容量，因为并未在发射机处完全预先消除干扰。在在此所述的DIDO发送算法中，采用块对角化来预先消除用户间干扰。然而，所提出自适应发送策略可以应用于任何DIDO系统，无需考虑预编码技术的类型。

[0300] 本专利申请描述了上述本发明以及在先申请的实施方式的扩展，包括但不限于以下附加特征：

[0301] 1、可由无线客户装置采用在先申请中用于信道估计的训练符号来对自适应DIDO方案中的链路质量度量进行评价。

[0302] 2、如在先申请中所述那样，基站接收来自客户端设备的信号特征数据。在当前实施方式中，信号特征数据被定义为用于对自适应进行激活的链路质量度量。

[0303] 3、在先申请描述了一用于选择天线和用户数量的机制，并定义了吞吐量分配。此外，可如在先申请那样，将不同级别的吞吐量动态地指派给不同客户端。本发明的当前实施方式定义了与该选择和吞吐量分配相关的新型标准。

[0304] b、本发明的实施方式

[0305] 所提出的自适应DIDO技术的目标为通过将时间、频率以及空间中的无线资源动态分配给系统中的不同用户来增强每用户的频谱效率或下行链路频谱效率。该整体自适应标准用于在满足目标误码率的同时，提高吞吐量。根据传播状态，还可使用该自适应算法经由分集方案来改进用户的链路质量(或覆盖率)。图21显示的流程图描述了自适应DIDO方案的步骤。

[0306] 在2102，基站(BS)收集来自所有用户的信道状况信息。在2104，根据所接收的CSI，基站在时域/频域/空域计算链路质量度量。在2106，使用这些链路质量度量来选择将在下一传输中被服务的用户，以及针对每一用户的发送模式。注意，发送模式包括调制/编码以及DIDO方案的不同组合。最后，在2108，BS经由DIDO预编码而将数据发送至用户。

[0307] 在2102，基站选择来自所有用户设备的信道状况信息(CSI)。在2104，基站使用该CSI来确定所有用户设备的瞬时或统计信道质量。在DIDO-OFDM系统中，可在时域、频域和空域对信道质量(或链路质量度量)进行估计。之后，在2106，基站使用链路质量度量来确定最佳用户子集以及用于当前传播状态的发送模式。DIDO发送模式集合被组合为DIDO方案(即，天线选择或复用)、调制/编码方案(MCS)以及阵列配置的组合。在2108，通过使用所选用户数量以及发送模式，将数据发送至用户设备。

[0308] 可通过查找表(LUT)(该查找表是基于DIDO系统不同传播环境中的误码率性能而被预先计算的)来进行模式选择。这些LUT将信道质量信息映射至误码率性能。为了构建LUT，可根据SNR评价DIDO系统在不同传播情形中误码率性能。从误码率曲线可看出，可计算实现某一预定目标误码率所需的最小SNR。我们将该SNR需求定义为SNR阈值。之后，在不同的传播情形以及针对不同的DIDO发送模式来评价SNR阈值，并将其存储在LUT中。例如，可使用图24和图26中SER结果来构建LUT。之后，根据该LUT，基站可选择针对活动用户的发送模式，该模式可在满足预定目标误码率的同时提高吞吐量。最后，基站经由DIDO预编码而将数据发送至所选用户。注意，可将不同DIDO模式指派给不同的时隙、OFDM调以及DIDO子流，以使得可在时域、频域和空域进行自适应。

[0309] 图19-图20显示了采用DIDO自适应的系统的一种实施方式。引入了若干新的功能单元来实施所提出的DIDO自适应算法。具体而言，在一种实施方式中，DIDO配置器1910可基

于用户设备所提供的信道质量信息1912，执行多种功能，包括选择用户数量、DIDO发送方案(即，天线选择和复用)、调制/编码方案(MCS)以及阵列配置。

[0310] 用户选择器单元1902基于由DIDO配置器1910所获取的反馈信息，选择与多个用户U1-UM相关联的数据，并将该信息提供每多个编码调制单元1904中的每个编码调制单元。每个编码调制单元1904对每个用户的信息比特进行编码和调制，并将他们发送至映射单元1906。该映射单元1906将输入比特映射至复数符号，并将其发送至预编码单元1908。编码调制单元1904和映射单元1906均利用获取自DIDO配置器单元1910的信息，选择为每一用户所采用的调制/编码方案类型。所述信息可由配置器单元1910通过利用反馈单元1912所提供的每一用户的信道质量信息来计算。DIDO预编码单元1908利用由DIDO配置器单元1910所获取的信息来计算DIDO预编码权重，并对获取自映射单元1906的输入符号进行预编码。由DIDO预编码单元1906将每一预编码后的数据流发送至OFDM单元1915，该OFDM单元1915计算IFFT并加入循环前缀。将该信息发送至D/A单元1916，该D/A单元1916进行数模转换，并将最终的模拟信号发送至RF单元1914。该RF单元1914将基带信号升频至中频/射频，并将其发送至发射天线。

[0311] 每一客户端设备的RF单元2008接收从DIDO发射机单元1914发送的信号，将该信号降频至基带，并将降频之后的信号提供给A/D单元2010。之后，该A/D单元2010将该信号从模拟转换为数字，并将其发送至OFDM单元2013。该OFDM单元2013移除循环前缀，并执行FFT，以将信号报告至频域。在训练周期，OFDM单元2013将输出发送至信道估计单元2004，该信道估计单元2004在频域中计算信道估计。可选地，可在时域计算信道估计。在数据周期期间，OFDM单元2013将输出发送至接收机单元2002，该接收机单元2002对信号进行解调/解码，以获取数据2014。所述信道估计单元2004将信道估计发送至DIDO反馈生成器单元2006，该DIDO反馈生成器单元2006可对信道估计进行量化，并经由反馈控制信道1912而将其发回发射机。

[0312] 所述DIDO配置器1910可使用在基站处得到的信息，或在优选实施方式中，额外使用工作于每一用户设备处的DIDO反馈生成器2006(参见图20)的输出。该DIDO反馈生成器2006使用所估计的信道状况2004和/或接收机处的类似于所估计的SNR的其他参数来生成将被输入至DIDO配置器1910的反馈消息。所述DIDO反馈生成器2006可在接收机处对信息进行压缩、量化和/或使用本领域所公知的一些有限反馈策略。

[0313] 所述DIDO配置器1910可使用从DIDO反馈控制信道1912恢复的信息。DIDO反馈控制信道为逻辑或物理控制信道，该信道可用于将DIDO反馈生成器2006的输出从用户发送至基站。控制信道1912可以以任意数量的本领域所公知的方式实施，且可为逻辑或物理控制信道。作为物理信道，其可包括指派给用户的专用时隙/频隙。其还可为由所有用户共享的随机访问信道。所述控制信道可被预先指派，或可由现有的控制信道中预定方式的侵占比特(stealing bits)来创建。

[0314] 在以下论述中，将在真实传播环境中描述通过利用DIDO-OFDM原型进行测量所获取的结果。这些结果表明了自适应DIDO系统中潜在增益的可实现性。首先展现不同级别DIDO系统的性能，表明可增大天线/用户数量，以实现更大的下行链路吞吐量。之后，描述与用户设备的位置有关的DIDO性能，表明需要跟踪变化的信道状态。最后，对采用分集技术的DIDO系统的性能进行描述。

[0315] i、不同级别DIDO系统的性能

[0316] 利用越来越多的发射天线(N=M,其中M为用户数量)来评价不同DIDO系统的性能。将以下系统的性能进行比对:SISO、DIDO2×2、DIDO4×4、DIDO6×6以及DIDO8×8。DIDON×M指在BS处具有N个发射天线以及M个用户的DIDO。

[0317] 图22显示了发射/接收天线布局。以方形阵列配置来布置发射天线2201,且用户位于发射阵列周围。在图22,T指代"发射"天线,U指"用户设备"2202。

[0318] 8元发射阵列中的不同天线子集处于活动状态,这取决于针对不同测量所选取的N值。对于每一DIDO级别(N),选择可对8元阵列的固定大小约束所针对的最大真实地区进行覆盖的天线子集。该标准被测量可增强给定N值的空间分集。

[0319] 图23示出了针对适合可用真实地区(即,虚线)的不同DIDO级别的阵列配置。方形虚框具有24"×24"的尺寸,对应于450MHz载波频率处的~λ×λ。

[0320] 基于与图23相关的评述以及参考图22,现将定义并比较以下系统中每一系统的性能:

[0321] 具有T1和U1的SISO(2301)

[0322] 具有T1,2和U1,2的DIDO2×2(2302)

[0323] 具有T1,2,3,4和U1,2,3,4的DIDO4×4(2303)

[0324] 具有T1,2,3,4,5,6和U1,2,3,4,5,6的DIDO6×6(2304)

[0325] 具有T1,2,3,4,5,6,7,8和U1,2,3,4,5,6,7,8的DIDO8×8(2305)

[0326] 图24示出了在4-QAM和1/2FEC(前向纠错)率情况下,上述DIDO系统中SER、BER、SE(频谱效率)和实际吞吐量性能与发射(TX)功率的函数关系。观察得出,SER和BER性能会因N值增大而下降。该影响是由以下两个现象造成的:对于固定的TX功率,DIDO阵列的输入功率会在越来越多的用户(或数据流)之间被分割;空间分集会随着实际DIDO信道中的用户数量的增大而减小。

[0327] 如图24所示,为了比较不同级别DIDO系统的相对性能,将目标BER固定为10^{-4}(该值可根据系统而变化),该值大致对应于SER=10^{-2}。我们将对应于该目标的TX功率值称之为TX功率阈值(TPT)。对于任何N,如果TX功率低于TPT,我们假设不可能在DIDO级别N下进行发送,且我们需要切换至更低级别的DIDO。此外,在图24,可观察得出,当TX功率超过针对任意N值的TPT时,SE和实际吞吐量性能会达到饱和。根据这些结果,可将自适应发送策略设计成在不同级别DIDO之间进行切换,以增强针对固定预定目标误码率的SE或实际吞吐量。

[0328] ii、可变用户位置情况下的性能

[0329] 该试验的目标在于,经由在空间关联信道中进行仿真,评价不同用户位置的DIDO性能。DIDO2×2系统被视为具有4QAM以及1/2FEC率。如图25所示,用户1位于发射阵列的侧射(broadside)方向,而用户2的位置从侧射方向变为端射(endfire)方向。发射天线间隔~λ/2,且与用户相隔~2.5λ。

[0330] 图26示出了针对用户设备2的不同位置,SER和每用户的SE结果。从发射阵列的边射方向测量,用户设备的到达角度(AOA)为0°至90°。观察得出,随着用户设备的角距增大,DIDO性能将会提升,因为DIDO信道内存在更大的分集。此外,在目标SER=10^{-2}处,AOA2=0°与AOA2=90°这两种情况之间存在10dB的间隙,该结果与图35中针对角度扩展10°所获得仿真结果一致。此外,注意,对于AOA1=AOA2=0°的情况而言,两个用户之间可能存在耦合效

应(因他们的天线相邻近所导致),这可能会使得他们的性能与图35中的仿真结果不同。

[0331] 111、针对D1D08×8的优选情形

[0332] 图24显示了D1D08×8产生比更低级D1D0更大的SE,但具有更高TX功率需求。该分析的目标在于示出存在该情况,即D1D08×8不仅在峰值频谱效率(SE)方面,而且还在TX功率需求(或TPT)方面,胜过D1D02×2,以实现所述峰值SE。

[0333] 注意,在i.i.d.(瑞想)信道中,TX功率在D1D08×8与D1D02×2的SE之间存在~6dB的间隙。该间隙是因该事实导致的,即D1D08×8将TX功率在8个数据流之间进行了分割,而D1D02×2仅在两个流之间进行分割。该结果经由图32中的仿真而被示出。

[0334] 然而,在空间关联信道中,TPT为传播环境特性(例如,阵列朝向、用户位置、角度扩展)的函数。例如,图35示出了针对两个不同用户设备位置之间的低角度扩展的~15dB间隙。本申请图26中展示了相类似的结果。

[0335] 类似于MIMO系统,当用户位于TX阵列的端射方向时,D1D0系统的性能会下降(因缺少分集所导致),该影响可通过利用当前D1D0原型进行测量而观察得出。因此,一种示出D1D08×8胜过D1D02×2的方式为将用户置于相对于D1D02×2阵列的端射方向。在此情形,D1D08×8胜过了D1D02×2,因为8-天线阵列提供了更高的分集。

[0336] 在该分析中,考虑了以下系统:

[0337] 系统1:4-QAM的D1D08×8(每时隙发送8个并行数据流);

[0338] 系统2:64-QAM的D1D02×2(每4个时隙,对发送用户X和Y进行一次发送)。对于此系统,我们考虑TX和RX天线位置的四种组合:a)T1,T2U1,2(端射方向);b)T3,T4U3,4(端射方向);c)T5,T6U5,6(与端射方向相隔~30°);d)T7,T8U7,8(NLOS(无视距));

[0339] 系统3:64-QAM的D1D08×8;以及

[0340] 系统4:64-QAM的MISO8×1(每8个时隙,对发送用户X进行一次发送)。

[0341] 对于所有这些情况,使用3/4的FEC率。

[0342] 图27给示了用户的位置。

[0343] 在图28中,SER结果示出了由于不同的阵列方向和用户位置的在系统2a和2c之间的a~15dB的间隙(与在图35中的仿真结果相似)。在第二行中的第一子图示出了SE曲线饱和的TX功率的值(即,对应于BER1e-4)。我们观察到系统1比系统2对于较低的TX功率需求(小于~5dB)产生了更大的每个用户的SE。而且,由于D1D08×8在D1D02×2上的复用增益,D1D08×8相比于D1D02×2的好处对于DL(下行链路)SE和DL实际吞吐量来说更加明显。由于波束成形的阵列增益(即,具有MISO8×1的MRC),系统4比系统1具有更低的TX功率需求(小于5dB)。但是系统4相比于系统1仅产生了每个用户的SE的1/3。系统2比系统1的性能差(即,对于较大的TX功率需求产生了较低的SE)。最后,系统3比系统1对于较大的TX功率需求(~15dB)产生了大得多的SE(由于较大的阶数(larger order)调制)。

[0344] 根据这些结果,可以推断出以下结论:

[0345] 一种信道情形被确认为D1D08×8胜过D1D02×2(即对于较低的TX功率需求产生了较大的SE);

[0346] 在该信道情形中,D1D08×8比D1D02×2和MISO8×1产生了更大的每个用户的SE和DL SE;以及

[0347] 可以通过以较大的TX功率需求(大于~15dB)为代价使用高阶调制(即64-QAM,而

不是4-QAM)来进一步增大DIDO8×8的性能。

[0348] iv、具有天线选择的DIDO

[0349] 下面,我们评估在由IEEE学报接收的在Signal Processing上在2005年由R.Chen、R.W.Heath和J.G.Andrews发表的"Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers"中描述的天线选择算法的好处。我们用两个用户、4-QAM和1/2的FEC率来显现用于一个特定DIDO系统的结果。以下系统在图27中被比较:

[0350] 具有T1,2和U1,2的DIDO2×2;以及

[0351] 具有T1,2,3和U1,2的使用天线选择的DIDO3×2。

[0352] 发射天线位置和用户装置位置与图27中相同。

[0353] 图29示出了具有天线选择的DIDO3×2与DIDO2×2系统(不具有选择)相比可以提供~5dB的增益。注意信道几乎是静态的(即没有多普勒效应),所以选择算法适用于路径损耗和信道空间相关,而不是快速衰减。我们应当在具有高的多普勒效应的情形中看到不同的增益。而且,在该特定实验中,观察到天线选择算法选择天线2和3用于发送。

[0354] iv、用于LUT的SNR阈值

[0355] 在选择[0171],我们声明了模式选择通过LUT实现。LUT可以通过评估SNR阈值来被预计算以实现在不同传播环境中用于DIDO发送模式的特定预定义的目标误码率性能。下面,我们提供了具有和不具有天线选择和可变化的用户数量的DIDO系统的性能,所述性能可以用作构造LUT的指导。虽然图24、图26、图28、图29通过用DIDO原型实际测量得到,但下面的图通过仿真得到。以下BER结果假设没有FEC。

[0356] 图30示出了在独立同分布信道中不同的DIDO预编码方案的平均BER性能。标有"没有选择"的曲线是指使用BD的情况。在同一个图中,天线选择(ASel)的性能对于不同数量的额外天线(对于不同数量的用户)而被示出。可以看出,随着额外天线的数量增长,ASel提供更好的分集增益(以高SNR区的BER曲线的斜率为特征),产生了更好的覆盖。例如,如果我们将目标BER固定到10^{-2}(对于未编码的系统的实际值),则由ASel提供的SNR增益随着天线的数量增长。

[0357] 图31示出了对于不同的目标BER的作为在独立同分布信道中的额外发射天线的数量的函数的ASel的SNR增益。可以看出,仅通过添加1或2个天线,ASel与BD相比产生了巨大的SNR增益。在以下部分中,我们将仅对于1或2个额外天线的情况通过将目标BER固定到10^{-2}(对于未编码的系统)来评估ASel的性能。

[0358] 图32示出了对于在独立同分布信道中具有1和2个额外天线的BD和ASel的作为用户数量(M)的函数的SNR阈值。我们观察到由于对于较大数量的用户的较大的接收SNR需求,SNR阈值随着M增大。注意,我们假设对于任意数量的用户为固定的总发射功率(而不同数量的发射天线)。此外,图32示出了由于天线选择的增益对于在独立同分布信道中的任意数量的用户来说是恒定的。

[0359] 下面,我们示出了在空间相关信道中的DIDO系统的性能。我们通过在X.Zhuang、E.W. Vook、K.L. Baum、T.A.Thomas和M.Cudak于2004年9月在IEEE802.16Broadband Wireless Access Working Group上发表的"Channel models for link and system level simulations"中描述的COST-259空间信道模型仿真每个用户的信道。我们生成用于

每个用户的单一群。作为一种情况研究，我们假设了NLOS信道，在发射机有均匀线性阵列(ULA)，元件间隔为0.5λ。对于2个用户系统的情况，我们对于第一和第二用户分别用到达的AOA1和AOA2的平均角来仿真群。AOA相对于ULA的侧面方向而被测量。当在系统中有多于两个的用户，我们生成具有在范围[-Φ∞，Φ∞]中的均匀间隔的平均AOA的用户的群，其中我们定义

[0360] $$\Phi_M \triangleq \frac{\Delta\phi(M-1)}{2} \qquad (13)$$

[0361] K是用户的数量，Δφ是用户的平均AOA之间的角距。注意角度范围[-Φ∞，Φ∞]中心为0°，对应于ULA的侧射方向。下面，我们用BD和ASel发送方案和不同的用户数量来研究作为信道角度分布(AS)和用户间的角距的函数的DIDO系统的BER性能。

[0362] 图33示出了用于位于同一角度方向(即相对于ULA的侧射方向，AOA1=AOA2=0°)的具有不同AS值的两个用户的相对于每个用户的平均SNR的BER。可以看出，随着AS增大，BER性能改善且接近独立同分布情况。实际上，较高的AS在统计上产生了在两个用户的特征模式之间的较少覆盖和BD预编码器的更好的性能。

[0363] 图34示出了与图33相似的结果，但在用户之间具有较高的角距。我们考虑AOA1=0°，AOA2=90°(即90°角距)。在低的AS的情况下实现了最好的性能。实际上，对于高的角距的情况，当角距低时，在用户的特征模式之间有较少的交迭。有趣的是，我们观察到对于刚刚提到的相同的理由，在低AS中的BER性能要好于独立同分布信道。

[0364] 接下来，对于不同的相关情形中的10^{-2}的目标BER，我们计算SNR阈值。图35绘出了对于用户的平均AOA的不同值的作为AS的函数的SNR阈值。对于低的用户角距，具有合理的SNR需求(即18dB)的可靠的发送仅对于以高AS为特征的信道是可能的。另一方面，当用户在空间上分离时，需要较小的SNR来满足相同的目标BER。

[0365] 图36示出了用于5个用户的情况的SNR阈值。根据(13)中的定义生成具有不同的角距Δφ的值的用户平均AOA。我们观察到对于Δφ=0°和AS<15°，由于用户之间的小的角距，BD性能很差，没有满足目标BER。对于增大的AS，满足固定的目标BER的SNR需求减小。另一方面，对于Δφ=30°，在低的AS获得最小的SNR需求，与图35中的结果一致。随着AS增大，SNR阈值饱和至独立同分布信道中的一个。注意，具有5个用户的Δφ=30°对应于[-60°，60°]的AOA范围，这对于具有120°扇形单元的蜂窝系统中的基站是典型的。

[0366] 接下来，我们研究了在空间相关信道中的ASel发送方案的性能。图37比较了对于两个用户情况的具有1个和2个额外天线的BD和ASel的SNR阈值。我们考虑了用户间的角距的两种不同情况：{AOA1=0°，AOA2=0°}以及{AOA1=0°，AOA2=90°}。用于BD方案(即没有天线选择)的曲线与在图35中相同。我们观察到ASel分别对于高的AS产生了具有1个和2个额外天线的8dB和10dB的SNR增益。随着AS减小，在BD上的由于ASel的增益由于在MIMO广播信道中的自由度的数量减少而变得更小。有趣的是，对于AS=0°(即接近于LOS信道)以及情况{AOA1=0°，AOA2=90°}，ASel没有提供由于在空间域中的差异的任何增益。图38示出了与图37相似的结果，但是对于5个用户的情况。

[0367] 我们对于BD和ASel发送方案计算了作为在系统中的用户数量(M)的函数的SNR阈值(假设10^{-2}的一般目标BER)。SNR阈值对应于平均SNR，以使得总的发射功率对于任意M是恒定的。我们假设在方位角范围[-Φ∞，Φ∞]=[-60°，60°]内每个用户群的平均AOA之间的最大

间隔。然后，用户之间的角距是 $\Delta\phi=120^\circ/(M-1)$。

[0368] 图39示出了对于具有不同AS值的BD方案的SNR阈值。我们观察到由于用户之间的大的角距，对于具有相对较小数量的用户(即$K\leq 20$)的AS=0.1°(即低的角度扩散)，获得最低的SNR需求。然而，对于M>50，由于$\Delta\phi$非常小以及BD不能实行，SNR需求远远大于40dB。此外，对于AS>10°，SNR阈值对于任意M几乎保持恒定，在空间相关信道中的DIDO系统接近独立同分布信道的性能。

[0369] 为了减小SNR阈值的值并改善DIDO系统的性能，我们应用ASel发送方案。图40示出了对于具有1个和2个额外天线的BD和ASel的在具有AS=0.1°的空间相关信道中的SNR阈值。为了参考，我们还报告了用于在图32中示出的独立同分布情况的曲线。可以看到，对于较少的用户(即$M\leq 10$)，由于在DIDO广播信道中缺少分集，天线选择没有帮助降低SNR需求。随着用户数量增加，ASel从多用户分集中受益，产生了SNR增益(即对于M=20为4dB)。此外，对于$M\leq 20$，在高的空间相关信道中的具有1个或2个额外天线的ASel的性能是相同的。

[0370] 然后我们计算对于两种另外的信道情形的SNR阈值：图41中的AS=5°和图42中的AS=10°。图41与图40相比，示出了由于较大的角度扩散，ASel产生了也用于相对较少数量的用户(即$M\leq 10$)的SNR增益。如在图42中报告的，对于AS=10°，SNR阈值进一步减少，由于ASel的增益变得更高。

[0371] 最后，我们总结了目前对于相关信道提出的结果。图43和图44示出了分别具有1个和2个额外天线的作为对于BD和ASel方案的用户数量(M)和角度扩散(AS)的函数的SNR阈值。注意，AS=30°的情况实际上对应于独立同分布信道，我们在图中使用AS的这个值用于图形表示。我们观察到，虽然BD被信道空间相关所影响，ASel产生了对于任意AS的几乎相同的性能。此外，对于AS=0.1°，由于多用户分集，ASel对于低的M与BD性能相似，而对于大的M(即$M\geq 20$)超过BD。

[0372] 图49比较了在SNR阈值方面不同的DIDO方案的性能。所考虑的DIDO方案是：BD、ASel、具有特征模式选择(BD-ESel)的BD以及最大比合并(MRC)。注意MRC没有预先消除在发射机处的干扰(不像其它方法)，但在用户被空间分离的情况下提供了较大的增益。在图49中，我们绘出了当两个用户分别位于与发射阵列的侧射方向成-30°和30°时，对于DIDO N×2系统的对于目标$BER=10^{-3}$的SNR阈值。我们观察到，对于低的AS，MRC方案与其它方案相比提供了3dB的增益，因为用户的空间信道被很好地分离，用户间的干扰的影响很小。注意，在DIDO N×2上的MRC的增益是由于阵列增益。对于大于20°的AS，QR-ASel方案超过其它方案并与不具有选择的BD2×2相比产生了大约10dB的增益。QR-ASel和BD-ESel对于AS的任意值提供了大约相同的性能。

[0373] 上面描述的是用于DIDO系统的新的自适应发送技术。该方法在DIDO发送模式之间动态转换到不同的用户来增强用于固定的目标误码率的吞吐量。不同级别的DIDO系统的性能在不同的传播情况下被测量，观察到在吞吐量的巨大增益可以通过动态选择作为传播情况的函数的DIDO模式和用户数量来实现。

[0374] III.频率和相位差的预补偿

[0375] a.背景

[0376] 如之前所述，无线通信系统使用载波来传送信息。这些载波通常是正弦波，其振幅和/或相位响应于被发送的信息而被调制。正弦波的标称频率已知为载波频率。为了创建该

波形，发射机合成一个或两个正弦波，并使用升频转换来创建重叠在具有指定载波频率的正弦波上的调制后的信号。这可以通过直接转换来实现，其中，信号在载波上或通过多个升频转换阶段被直接调制。为了处理该波形，接收机必须解调所接收到的RF信号，并有效地移除调制载波。这需要接收机合成一个或多个正弦信号来反转在发射机处的调制过程，已知为降频转换。遗憾的是，在发射机和接收机生成的正弦波信号从不同的基准振荡器获得。没有基准振荡器创建了完美(perfect)的频率基准；实际上，通常与实际频率有一些偏差。

[0377] 在无线通信系统中，在发射机和接收机处的基准振荡器的输出的差异在接收机处创建了已知为载波频率偏移或简单的频率偏移的现象。本质上，在降频转换之后，在所接收的信号中有一些剩余调制(对应于发送和接收载波中的差异)。这创建了在所接收的信号中的失真，导致了较高的比特误码率和较低的吞吐量。

[0378] 存在用于处理载波频率偏移的不同技术。大多数方法估计在接收机处的载波频率偏移，然后应用载波频率偏移校正算法。载波频率偏移估计算法使用以下方法是盲目的(blind)：偏移QAM(T.Fusco和M.Tanda，"Blind Frequency-offset Estimation for OFDM/OQAM Systems，"Signal Processing，IEEE Transactions on[也参见Acoustics，Speech，and Signal Processing，IEEE Transactions on]，vol.55，pp.1828-1838，2007)；周期特性(E.Serpedin，A.Chevreuil，G. B.Giannakis和P. Loubaton，"Blind channel and carrier frequency offset estimation using periodic modulation precoders，" Signal Processing，IEEE Transactions on[也参见Acoustics，Speech，and Signal Processing，IEEE Transactions on]，vol.48，no.8，pp.2389-2405，Aug.2000)；或者正交频分复用(OFDM)结构方法中的循环前缀(J.J.van de Beek，M.Sandell和P.O.Borjesson，"ML estimation of time and frequency offset in OFDM systems，" Signal Processing，IEEE Transactions on[也参见Acoustics，Speech，and Signal Processing，IEEE Transactions on]，vol.45，no.7，pp.1800-1805，July1997；U.Tureli，H.Liu和M.D.Zoltowski，"OFDM blind carrier offset estimation：ESPRIT，"IEEE Trans.Commun.，vol.48，no.9，pp.1459-1461，Sept.2000；M.Luise，M.Marselli和R.Reggiannini，"Low-complexity blind carrier frequency recovery for OFDM signals over frequency-selective radio channels，"IEEE Trans.Commun.，vol.50，no.7，pp.1182-1188，July2002)。

[0379] 可替换地，专用训练信号可以被利用，包括重复的数据符号(P. H.Moose，"A technique for orthogonal frequency division multiplexing frequency offset correction，"IEEE Trans.Commun.，vol.42，no.10，pp.2908-2914，Oct.1994)；两个不同的符号(T.M.Schmidl and D.C.Cox，"Robust frequency and timing synchronization for OFDM，"IEEE Trans.Commun.，vol.45，no.12，pp.1613-1621，Dec.1997)；或者周期性插入的已知的符号序列(M.Luise和R.Reggiannini，"Carrier frequency acquisition and tracking for OFDM systems，"IEEE Trans.Commun.，vol.44，no.11，pp.1590-1598，Nov.1996)。校正可以以模拟或数字方式发生。接收机还可以使用载波频率偏移估计来预校正所发送的信号以消除偏移。由于多载波和OFDM系统对频率偏移的灵敏性，载波频率偏移校正对于多载波和OFDM系统被广泛地研究(J.J.van de Beek，M.Sandell和P.O.Borjesson，"ML estimation of time and frequency offset in OFDM systems，"

Signal Processing,IEEE Transactions on[也参见Acoustics,Speech,and Signal Processing,IEEE Transactions on],vol.45,no.7,pp.1800-1805,July1997;U.Tureli,H.Liu和M.D.Zoltowski,"OFDM blind carrier offset estimation:ESPRIT,"IEEE Trans.Commun.,vol.48,no.9,pp.1459-1461,Sept.2000;T.M.Schmidl和D.C.Cox,"Robust frequency and timing synchronization for OFDM,"IEEE Trans.Commun.,vol.45,no.12,pp.1613-1621,Dec.1997;M.Luise,M.Marselli和R.Reggiannini,"Low-complexity blind carrier frequency recovery for OFDM signals over frequency-selective radio channels,"IEEE Trans.Commun.,vol.50,no.7,pp.1182-1188,July2002)。

[0360] 频率偏移估计和校正对于多天线通信系统或更一般的MIMO(多输入多输出)系统是重要的问题。在MIMO系统中,发射天线被锁定到一个频率基准,接收机被锁定到另一个频率基准,在发射机和接收机之间有单个偏移。提出了几种算法来使用训练信号处理这个问题(K.Lee和J.Chun,"Frequency-offset estimation for MIMO and OFDM systems using orthogonal training sequences,"IEEE Trans.Veh.Technol.,vol.56,no.1,pp.146-156,Jan.2007;M.Ghogho和A.Swami,"Training design for multipath channel and frequency offset estimation in MIMO systems,"Signal Processing,IEEE Transactions on[也参见Acoustics,Speech,and Signal Processing,IEEE Transactions on],vol.54,no.10,pp.3957-3965,Oct.2006;以及adaptive tracking C.Oberli和B.Daneshrad,"Maximum likelihood tracking algorithms for MIMOOFDM,"in Communications,2004IEEE International Conference on,vol.4,June20-24,2004,pp.2468-2472)。在MIMO系统中遇到了更重要的问题,其中,发射天线没有被锁定到同一频率基准,但接收天线被锁定到一起。这实际上发生在空分多址存取(SDMA)系统的上行链路中,SDMA系统被视为MIMO系统,其中不同用户对应于不同的发射天线。在这种情况下,频率偏移的补偿更加复杂。具体地,频率偏移创建了在不同的被发送的MIMO流中的干扰。可以使用复杂的联合估计和均衡算法来进行校正(A.Kannan,T.P. Krauss和M.D.Zoltowski,"Separation of cochannel signals under imperfect timing and carrier synchronization,"IEEE Trans.Veh.Technol.,vol.50,no.1,pp.79-96,Jan.2001),以及在频率偏移估计之后的均衡(T.Tang和R.W.Heath,"Joint Frequency offset estimation and interference cancellation for MIMO-OFDM systems[mobile radio],"2004.VTC2004-Fall.2004IEEE60thVehicular Technology Conference,vol.3,pp.1553-1557,Sept.26-29,2004;X.Dai,"Carrier Frequency offset estimation for OFDM/SDMA systems using consecutive pilots,"IEEE Proceedings-Communications,vol.152,pp.624-632,Oct.7,2005),一些工作处理了剩余相位偏移和跟踪误差的相关问题,其中剩余相位偏移在频率偏移估计之后被估计和补偿,但这个工作仅考虑了SDMAOFDMA系统的上行链路(E. Haring,S.Bieder和A.Czylwik,"Residual carrier and sampling frequency synchronization in multiuser OFDM systems,"2006.VTC2006-Spring.IEEE63rd Vehicular Technology Conference,vol.4,pp.1937-1941,2006)。当所有发射和接收天线具有不同的频率基准时,在MIMO系统中发生最严重的情况。关于这个话题仅有的可用工作只处理了在平衰减信道中的估计误差的渐近分析(O.Besson和P.Stoica,"On parameter estimation of MIMO flat-fading channels with frequency offsets,"Signal

Processing,IEEE Transactions on[也参见Acoustics,Speech,and Signal Processing,IEEE Transactions on],vol.51,no.3,pp.602-613,Mar.2003)。

[0381] 当MIMO系统的不同发射天线不具有相同的频率基准,且接收天线独立地处理信号时,已经被深入研究的情况发生。这发生在已知为分布式输入输出(DIDO)通信系统(在文献中也称为MIMO广播信道)中发生。DIDO系统包括具有分布式天线的一个接入点,所述天线发送平行数据流(经由预编码)到多个用户来增强下行链路的吞吐量,此时使用相同的无线资源(即相同的时隙持续时间和频带)作为常规的SISO系统。DIDO系统的详细描述在S.G.Perlman和T.Cotter,2004年7月提交的题为"System and method for distributed input-distributed output wireless communications"的美国专利申请20060023803中提出。有许多实施DIDO预编码器的方式,一个解决方案是块对角化(BD),在例如以下文献中描述:Q.H.Spencer,A.L.Swindlehurst 和M.Haardt,"Zero-forcing methods for downlink spatial multiplexing in multiuser MIMO channels,"IEEE Trans.Sig.Proc,vol.52,pp.461-471,Feb.2004;K.K.Wong,R.D.Murch和K.B.Letaief,"A joint-channel diagonalization for multiuser MIMO antenna systems,"IEEE Trans.Wireless Comm.,vol.2,pp.773-786,jul2003;L.U.Choi和R.D.Murch,"A transmit preprocessing technique for multiuser MIMO systems using a decomposition approach,"IEEE Trans.Wireless Comm.,vol.3,pp.20-24,Jan2004;Z.Shen,J.G.Andrews,R.W.Heath和B.L Evans,"Low complexity user selection algorithms for multiuser MIMO systems with block diagonalization,"被接受发表在IEEE Trans.Sig.Proc,Sep.2005;Z.Shen,R.Chen,J.G.Andrews,R.W.Heath和B.L Evans,"Sum capacity of multiuser MIMO broadcast channels with block diagonalization,"被提交至IEEE Trans.Wireless Comm.,Oct.2005;R.Chen,R.W.Heath和J.G.Andrews,"Transmit selection diversity for unitary precoded multiuser spatial multiplexing systems with linear receivers,"被接受至IEEE Trans.on Signal Processing,2005。

[0382] 在DIDO系统中,发送预编码被用于分离用于不同用户的数据流。当发射天线射频链没有共享同一频率基准时,载波频率偏移导致了与系统实施相关的几个问题。当这发生时,每个天线以稍微不同的载波频率有效发送,这破坏了DIDO预编码器的完整性,导致每个用户遭受额外的干扰。下面提出的是对这个问题的几种解决方案。在解决方案的一个实施方式中,DIDO发射天线通过有线的,光学的或无线的网络来共享一个频率基准。在解决方案的另一个实施方式中,一个或多个用户估计频率偏移差异(天线对之间的偏移中的相对差异)并将该信息发送回发射机。然后发射机预校正频率偏移并继续进行用于DIDO的训练和预编码器估计相位。该实施方式在反馈信道存在延迟时有问题。原因是可能有由校正过程创建的剩余相位误差,该校正过程未考虑随后的信道估计。为了解决这个问题,一个另外的实施方式使用新的频率偏移和相位估计器,通过估计延迟解决了这个问题。基于仿真和通过DIDO-OFDM原型执行的实际测量来给出结果。

[0383] 在该文件中提出的频率和相位偏移补偿方法可能对由于接收机处的噪声的估计误差比较灵敏。因此,另一实施方式提出了用于时间和频率偏移估计的方法,在低的SNR条件下也很强健。

[0384] 有不同的用于执行时间和频率偏移估计的方法。由于其对同步误差的灵敏性，这些方法中的许多方法专门为OFDM波形而提出。

[0385] 这些算法没有典型地使用OFDM波形的结构，因此对于单一载波和多载波波形一般是足够的。下面描述的算法在使用已知的基准符号(例如，训练数据)以协助同步的技术的类之中。许多方法是Moose的频率偏移估计器的扩展(见P.H.Moose，"A technique for orthogonal frequency division multiplexing frequency offset correction，"IEEE Trans.Commun.，vol.42，no.10，pp.2908-2914，Oct.1994)。Moose提出了使用两个重复的训练信号并使用在所接收到的信号之间的相位差来获得频率偏移。Moose的方法仅能够校正分数(fractional)频率偏移。Moose的方法的扩展由Schmidl和Cox提出(T.M.Schmidl and D.C.Cox，"Robust frequency and timing synchronization for OFDM，"IEEE Trans.Commun.，vol.45，no.12，pp.1613-1621，Dec.1997)。他们的主要创新在于使用一个周期性的OFDM符号和另外的差分编码的训练符号。在第二个符号中的差分编码实现的整数偏移校正。Coulson考虑了在T.M.Schmidl和D.C.Cox，"Robust frequency and timing synchronization for OFDM，"IEEE Trans.Commun.，vol.45，no.12，pp.1613-1621，Dec.1997中描述的类似的设置，并在A.J.Coulson，"Maximum likelihood synchronization for OFDM using a pilot symbol:analysis，"IEEE J.Select.Areas Commun.，vol.19，no.12，pp.2495-2503，Dec.2001和A.J.Coulson，"Maximum likelihood synchronization for OFDM using a pilot symbol:algorithms，"IEEE J.Select.Areas Commun.，vol.19，no.12，pp.2486-2494，Dec.2001中提供了算法和分析的详细论述。一个主要的差别是Coulson使用了重复的最大长度序列来提供好的相关特性。他还建议使用线性调频(chirp)信号，因为其在时域和频域中的恒定的包络属性。Coulson考虑了实际的细节但没有包括整数估计。多个重复的训练信号被Minn et.al.in H.Minn，V.K.Bhargava和K.B.Letaief，"A robust timing and frequency synchronization for OFDM systems，"IEEE Trans.Wireless Commun.，vol.2，no.4，pp.822-839，July2003所考虑，但训练的结构没有被优化。Shi和Serpedin提出了训练结构具有形成帧同步的想法的一些最优性(K.Shi和E.Serpedin，"Coarse frame and carrier synchronization of OFDM systems:a new metric and comparison，"IEEE Trans.Wireless Commun.，vol.3，no.4，pp.1271-1284，July2004)。本发明的一个实施方式使用了Shi和Serpedin的方法来执行帧同步和分数频率偏移估计。

[0386] 在文献中的许多方法集中在帧同步和分数频率偏移校正上。整数偏移校正使用在T.M.Schmidl和D.C.Cox，"Robust frequency and timing synchronization for OFDM，"IEEE Trans.Commun.，vol.45，no.12，pp.1613-1621，Dec.1997中的另外的训练符号来被解决。例如，Morrelli等在M.Morelli，A.N.D'Andrea和U.Mengali，"Frequency ambiguity resolution in OFDM systems，"IEEE Commun.Lett.，vol.4，no.4，pp.134-136，Apr.2000中得到了T.M.Schmidl和D.C.Cox，"Robust frequency and timing synchronization for OFDM，"IEEE Trans.Commun.，vol.45，no.12，pp.1613-1621，Dec.1997的改进版本。使用不同的前导码结构的可替换的方法由Morelli和Mengali提出M.Morelli和U.Mengali，"An improved frequency offset estimator for OFDM applications，"IEEE Commun.Lett.，vol.3，no.3，pp.75-77，Mar.1999)。这个方法使用了在M个重复的相同训练符号之间的相关

性来通过因子增大分数频率偏移估计器的范围。这是最好的线性无偏估计器且接受了最大的偏移(具有合适的设计),但不提供好的时序同步。

[0387] 系统描述

[0388] 本发明的一个实施方式使用基于信道状况信息的预编码来消除DIDO系统中的频率和相位偏移。见图11以及对于该实施方式的描述的上面的相关描述。

[0389] 在本发明的一个实施方式中,每个用户使用装备有频率偏移估计器/补偿器的接收机。如图45中所示出的,在本发明的一个实施方式中,包括接收机的系统包括多个RF单元4508、相应的多个A/D单元4510、装备有频率偏移估计器/补偿器4512的接收机以及DIDO反馈生成器单元4506。

[0390] RF单元4508接收从DIDO发射机单元发送的信号,将信号降频转换到基带,并将降频转换后的信号提供到A/D单元4510。然后A/D单元4510将信号从模拟转换为数字,并将它发送到频率偏移估计器/补偿器单元4512。频率偏移估计器/补偿器单元4512估计频率偏移并补偿频率偏移,如在这里所描述的。然后将补偿后的信号发送到OFDM单元4513。OFDM单元4513移除循环前缀并运行快速傅立叶变换(FFT)来将信号报告给频域。在训练期间,OFDM单元4513将输出发送到信道估计单元4504来计算在频域中的信道估计。可替换地,信道估计可以在时域中被计算。在数据周期期间,OFDM单元4513将输出发送到DIDO接收机单元4502,该DIDO接收机单元4502对信号进行解调/解码以获得数据。信道估计单元4504将信道估计发送到DIDO反馈生成器单元4506,该DIDO反馈生成器单元4506可以量化信道估计并经由反馈控制信道将它们发送回发射机,如所示出的。

[0391] 对用于DIDO2×2情形的算法的一个实施方式的描述

[0392] 下面描述的是用于DIDO系统中的频率/相位偏移补偿的算法的实施方式。DIDO系统模型开始在具有和不具有频率/相位偏移的情况下被描述。为了简便,提供了DIDO2×2系统的特定实施。然而,本发明的基本原理还可以在高阶DIDO系统中被实施。

[0393] 具有/不具有频率和相位偏移的DIDO系统模型

[0394] DIDO2×2的所接收到的信号可以对于第一用户写成:

[0395] $r_1[t]=h_{11}(w_{11}x_1[t]+w_{21}x_2[t])+h_{12}(w_{12}x_1[t]+w_{22}x_2[t])$ (1)

[0396] 以及对于第二个用户写成:

[0397] $r_2[t]=h_{21}(w_{11}x_1[t]+w_{21}x_2[t])+h_{22}(w_{12}x_1[t]+w_{22}x_2[t])$ (2)

[0398] 其中t是离散时间指数,h_{mn}和w_{mn}分别是在第m个用户和第n个发射天线之间的信道和DIDO预编码权重,x_m是对于用户m的发送信号。注意,h_{mn}和w_{mn}不是t的函数,因为我们假定信道在训练和数据发送之间的周期上是恒定的。

[0399] 在频率和相位偏移存在时,接收到的信号被表示为

[0400] $r_1[t]=e^{j(\omega_{r_1}-\omega_{t_1})(t-\tau_{11})}h_{11}(w_{11}x_1[t]+w_{21}x_2[t])+e^{j(\omega_{r_1}-\omega_{t_2})(t-\tau_{12})}h_{12}(w_{12}x_1[t]+w_{22}x_2[t])$ (3)

[0401] 以及

[0402] $r_2[t]=e^{j(\omega_{r_2}-\omega_{t_1})(t-\tau_{21})}h_{21}(w_{11}x_1[t]+w_{21}x_2[t])+e^{j(\omega_{r_2}-\omega_{t_2})(t-\tau_{22})}h_{22}(w_{12}x_1[t]+w_{22}x_2[t])$ (4)

[0403] 其中,T_s是符号周期;对于第n个发射天线,$\omega_{t_n}=2\Pi f_{t_n}$;对于第m个用户,$\omega_{r_m}=2\Pi f_{r_m}$;以及$f_{t_n}$和$f_{r_m}$分别是对于第n个发射天线和第m个用户的实际载波频率(由偏移影响)。值τ_{mn}表示在信道h_{mn}上导致相位偏移的随机延迟。图46绘出了DIDO2×2系统模型。

[0404] 对于时间，我们使用以下定义：

[0405] $\Delta\omega_{mn}=\omega_{Rm}-\omega_{Tn}$ (5)

[0406] 以用来表示在第m个用户和第n个发射天线之间的频率偏移。

[0407] 本发明的一个实施方式的描述

[0408] 根据本发明的一个实施方式的方法在图47中被示出。该方法包括以下一般步骤(包括子步骤，如示出的)：用于频率偏移估计的训练周期4701；用于信道估计的训练周期4702；经由具有补偿的DIDO预编码的数据发送4703。这些步骤在下面被详细描述。

[0409] (a)用于频率偏移估计的训练周期(4701)

[0410] 在第一训练周期期间，基站将来自每个发射天线的一个或多个训练信号发送到用户中的一个(4701a)。如在这里所描述的，"用户"是无线客户装置。对于DIDO2×2的情况，由第m个用户接收到的信号由以下给出：

[0411] $$r_m[t]=e^{j\Delta\omega_{m1}t(t-\tau_{m1})}h_{m1}p_1[t]+e^{j\Delta\omega_{m2}t(t-\tau_{m2})}h_{m2}p_2[t] \quad (6)$$

[0412] 其中，p_1和p_2分别是从第一和第二天线发送的训练序列。

[0413] 第m个用户可以使用任意类型的频率偏移估计器(即通过训练序列的卷积)并估计偏移$\Delta\omega_{m1}$和$\Delta\omega_{m2}$。然后，根据这些值，用户计算两个发射天线之间的频率偏移：

[0414] $$\Delta\omega_T=\Delta\omega_{m2}-\Delta\omega_{m1}=\omega_{T1}-\omega_{T2} \quad (7)$$

[0415] 最后，在(7)中的值被反馈回基站(4701b)。

[0416] 注意，在(6)中的p_1和p_2被设计为是正交的，从而用户可以估计$\Delta\omega_{m1}$和$\Delta\omega_{m2}$。可替换地，在一个实施方式中，相同的训练序列被用在两个连续的时隙，用户从中估计偏移。此外，为了改善(7)中的偏移的估计，上面描述的相同的计算对于DIDO系统的所有用户(不仅对于第m个用户)来说可以被完成，最后的估计可以是从所有用户获得的值的(加权后的)平均值。然而，这个解决方案需要更多的计算时间和反馈量。最后，频率偏移估计的更新只有在频率偏移随时间变化时才需要。因此，根据发射机处的时钟的稳定性，算法的步骤4701可以长期被执行(即对于每个数据发送)，使得上述反馈减少。

[0417] (b)用于信道估计的训练周期(4702)

[0418] 在第二训练周期期间，基站首先从第m个用户或从多个用户或得具有在(7)中的值的频率偏移反馈。在(7)中的值被用于预补偿在发射端的频率偏移。然后，基站将训练数据发送到所有用户来用于信道估计(4702a)。

[0419] 对于DIDO2×2系统，在第一个用户处接收到的信号由以下给出：

[0420] $$r_1[t]=e^{j\omega_{T1}t(t-\tilde{\tau}_1)}h_{11}p_1[t]+e^{j\omega_{T2}t(t-\tilde{\tau}_2)}h_{12}e^{-j\Delta\omega_T t}p_2[t] \quad (8)$$

[0421] 以及第二个用户处：

[0422] $$r_2[t]=e^{j\omega_{T1}t(t-\tilde{\tau}_1)}h_{21}p_1[t]+e^{j\omega_{T2}t(t-\tilde{\tau}_2)}h_{22}e^{-j\Delta\omega_T t}p_2[t] \quad (9)$$

[0423] 其中，$\tilde{\tau}_{mn}=\tau_{mn}+\Delta t$和$\Delta t$是在基站的第一发送和第二发送之间的随机或已知的延迟。此外，p_1和p_2分别是用户频率偏移和信道估计的从第一和第二天线发送的训练序列。

[0424] 注意，预补偿在该实施方式中仅仅被应用到第二天线。

[0425] 展开(8)，我们获得

[0426] $$r_1[t]=e^{j\Delta\omega_{m1}T_s t}e^{j\theta_{11}}[h_{11}p_1[t]+e^{j(\theta_{12}-\theta_{11})}h_{12}p_2[t]] \quad (10)$$

[0427] 类似的对于第二用户：

[0428] $$r_2[t]=e^{j\Delta\omega_{m2}T_s t}e^{j\theta_{21}}[h_{21}p_1[t]+e^{j(\theta_{22}-\theta_{21})}h_{22}p_2[t]] \quad (11)$$

[0429] 其中，$\theta_{mn}=-\Delta\omega_{mn}T_s\tilde{t}_{mn}$。

[0430] 在接收端，用户通过使用训练序列p_1和p_2来补偿剩余频率偏移。然后，用户通过训练矢量信道来进行估计(4702b)：

[0431] $$h_1=\begin{bmatrix} h_{11} \\ e^{j(\theta_{12}-\theta_{11})}h_{12}\end{bmatrix} \qquad h_2=\begin{bmatrix} h_{21} \\ e^{j(\theta_{22}-\theta_{21})}h_{22}\end{bmatrix} \quad (12)$$

[0432] 在(12)中的这些信道或信道状况信息(CSI)被反馈回基站(4702b)，基站如下面部分中描述的那样来计算DIDO预编码器。

[0433] (c)具有预补偿的DIDO预编码(4703)

[0434] 基站从用户接收(12)中的信道状况信息(CSI)并通过块对角化(BD)来计算预编码的权重(4703a)，以使得

[0435] $$\mathbf{w}_1^T\mathbf{h}_2=0, \qquad \mathbf{w}_2^T\mathbf{h}_1=0 \quad (13)$$

[0436] 其中，矢量h_1在(12)中被定义，以及$w_m=[w_{m1},w_{m2}]$。注意，在本公开中提出的本发明可以被用于除BD之外的任何其他DIDO预编码方法中。基站还通过使用(7)中的估计来预补偿频率偏移，并通过估计第二训练发送和当前发送之间的延迟(Δt_0)来预补偿相位偏移(4703a)。最后，基站经由DIDO预编码器将数据发送到用户(4703b)。

[0437] 在发送过程之后，在用户1处接收到的信号由以下给出：

[0438] $$\begin{aligned} r_1[t]&=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{11}-\Delta t_0)}h_{11}[w_{11}x_1[t]+w_{21}x_2[t]] \\ &=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{11}-\Delta t_0)}h_{12}e^{-j\Delta\omega_{m2}T_s(t-\Delta t_0)}[w_{12}x_1[t]+w_{22}x_2[t]] \\ &=\gamma_1[t]\left[h_{11}(w_{11}x_1[t]+w_{21}x_2[t])+e^{j(\Delta\omega_{m1}t-\Delta\omega_{m2}t-\theta_{12}+\theta_{11})}h_{12}(w_{12}x_1[t]+w_{22}x_2[t])\right] \\ &=\gamma_1[t]\left[(h_{11}w_{11}+e^{j(\theta_{12}-\theta_{11})}h_{12}w_{12})x_1[t]+(h_{11}w_{21}+e^{j(\theta_{12}-\theta_{11})}h_{12}w_{22})x_2[t]\right] \end{aligned} \quad (14)$$

[0439] 其中，$\gamma_1[t]=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{11}-\Delta t_0)}$。使用属性(13)，我们获得

[0440] $$r_1[t]=\gamma_1[t]\mathbf{w}_1^T\mathbf{h}_1x_1[t] \quad (15)$$

[0441] 类似地，对于用户2，我们得到：

[0442] $$\begin{aligned} r_2[t]&=e^{j\Delta\omega_{m2}T_s(t-\tilde{t}_{21}-\Delta t_0)}h_{21}[w_{11}x_1[t]+w_{21}x_2[t]] \\ &+e^{j\Delta\omega_{m2}T_s(t-\tilde{t}_{21}-\Delta t_0)}h_{22}e^{-j\Delta\omega_{m1}T_s(t-\Delta t_0)}[w_{12}x_1[t]+w_{22}x_2[t]] \end{aligned} \quad (16)$$

[0443] 展开(16)：

[0444] $$r_2[t]=\gamma_2[t]\mathbf{w}_2^T\mathbf{h}_2x_2[t] \quad (17)$$

[0445] 其中，$\gamma_2[t]=e^{j\Delta\omega_{m2}T_s(t-\tilde{t}_{m2}-\Delta t_2)}$。

[0446] 最后，用户计算剩余频率偏移和信道估计来解调数据流$x_1[t]$和$x_2[t]$(4703c)。

[0447] 一般化到DIDO N×M

[0448] 在该部分中，之前描述的技术被一般化到具有N个发射天线和M个用户的DIDO系统。

[0449] i.用户频率偏移估计的训练周期

[0450] 在第一训练周期期间，由于从N个天线发送的训练序列的由第m个用户接收到的信号由以下给出：

$$r_m[t]=\sum_{n=1}^{N}e^{j\Delta\omega_{mn}T_s(t-\tau_{mn})}h_{mn}p_n[t] \tag{18}$$

[0452] 其中，p_n是从第n个天线发送的训练序列。

[0453] 在估计偏移$\Delta\omega_{mn}$之后，$\forall n=1,\ldots,N$，第m个用户计算第一个和第n个发射天线之间的频率偏移：

[0454] $$\Delta\omega_{1,fm}=\Delta\omega_{m1}=\Delta\omega_{m1}=\omega_{Tx}-\omega_{Rx} \tag{19}$$

[0455] 最后，(19)中的值被反馈到基站。

[0456] ii.用于信道估计的训练周期

[0457] 在第二训练周期期间，基站首先从第m个用户或从多个用户获得具有在(19)中的值的频率偏移反馈。在(19)中的值被用于预补偿在发射端的频率偏移。然后，基站将训练数据发送到所有用户来用于信道估计。

[0458] 对于DIDO N×M系统，在第m个用户处接收到的信号由以下给出：

$$\begin{aligned}r_m[t]&=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}h_{m1}p_1[t]+\sum_{n=2}^{N}e^{j\Delta\omega_{mn}T_s(t-\tilde{t}_{mn})}h_{mn}e^{-j\Delta\omega_{Tx,n}T_s t}p_n[t]\\[0459]&=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}\left[h_{m1}p_1[t]+\sum_{n=2}^{N}e^{j(\theta_{mn}-\theta_{m1})}h_{mn}p_n[t]\right]\\&=e^{j\Delta\omega_{m1}T_s(t-\tilde{t}_{m1})}\sum_{n=1}^{N}e^{j(\theta_{mn}-\theta_{m1})}h_{mn}p_n[t]\end{aligned} \tag{20}$$

[0460] 其中，$\theta_{mn}=-\Delta\omega_{mn}T_s\tilde{t}_{mn}$、$\tilde{t}_{mn}=t_{mn}+\Delta t$以及$\Delta t$是基站的第一和第二发送之间的随机或已知的延迟。此外，P_n是用于频率偏移和信道估计的从第n个天线发送的训练序列。

[0461] 在接收侧，用户通过使用训练序列P_n来补偿剩余频率偏移。然后，每个用户m通过训练矢量信道来进行估计：

[0462] $$\mathbf{h}_m=\begin{bmatrix}h_{m1}\\e^{j(\theta_{m2}-\theta_{m1})}h_{m2}\\\vdots\\e^{j(\theta_{mN}-\theta_{m1})}h_{mN}\end{bmatrix} \tag{21}$$

[0463] 并反馈回基站，基站如在以下部分中描述的那样计算DIDO预编码器。

[0464] iii.具有预补偿的DIDO预编码

[0465] 基站从用户接收(12)中的信道状况信息(CSI)并通过块对角化(BD)来计算预编码的权重，以使得

[0466] $$\mathbf{w}_m^T\mathbf{h}_l = 0, \qquad \forall m \neq l, \qquad m = 1,\ldots,M \tag{22}$$

[0467] 其中，矢量$\mathbf{h}_m$在(21)中被定义，以及$\mathbf{w}_m = [w_{m1}, w_{m2}, \ldots, w_{mN}]$。基站还通过使用(19)中的估计来预补偿频率偏移，并通过估计第二训练发送和当前发送之间的延迟($\Delta\tau_m$)来预补偿相位偏移。最后，基站经由DIDO预编码器将数据发送到用户。

[0468] 在发送过程之后，在用户i处接收到的信号由以下给出：

$$\begin{aligned}
r_i[t] &= e^{j\Delta\omega_i T_s (t-\xi_m-\Delta\tau_m)} h_{ii} \sum_{m=1}^{M} w_{mi} x_m[t] + \\
&+ \sum_{m=2}^{N} e^{j\Delta\omega_i T_s (t-\xi_m-\Delta\tau_m)} h_{im} e^{-j\Delta\omega_{m-1} T_s (t-\Delta\tau_m)} \sum_{m=1}^{M} w_{mm} x_m[t] \\
&= e^{j\Delta\omega_i T_s (t-\Delta\tau_m)} e^{-j\Delta\omega_i T_s \xi_m} h_{ii} \sum_{m=1}^{M} w_{mi} x_m[t] \\
&+ \sum_{m=2}^{N} e^{j\Delta\omega_i T_s (t-\Delta\tau_m)} e^{-j\Delta\omega_m T_s \xi_m} h_{im} \sum_{m=1}^{M} w_{mm} x_m[t]
\end{aligned}$$

[0469]
$$\begin{aligned}
&= \gamma_i[t] \left[h_{ii} \sum_{m=1}^{M} w_{mi} x_m[t] + \sum_{m=2}^{N} e^{j(\phi_m - \phi_i)} h_{im} \sum_{m=1}^{M} w_{mi} x_m[t] \right] \\
&= \gamma_i[t] \left[\sum_{m=1}^{N} e^{j(\phi_m - \phi_i)} h_{im} \sum_{m=1}^{M} w_{mm} x_m[t] \right] \\
&= \gamma_i[t] \sum_{m=1}^{M} \left[\sum_{m=1}^{N} e^{j(\phi_m - \phi_i)} h_{im} w_{mm} \right] x_m[t] \\
&= \gamma_i[t] \sum_{m=1}^{M} \mathbf{w}_m^T \mathbf{h}_i x_m[t]
\end{aligned} \tag{23}$$

[0470] 其中，$\gamma_i[n] = e^{j\Delta\omega_i T_s (t-\xi_m-\Delta\tau_m)}$。使用属性(22)，我们得到：

$$r_i[t] = \gamma_i[t] \mathbf{w}_i^T \mathbf{h}_i x_i[t] \tag{24}$$

[0471] 最后，用户计算剩余频率偏移和信道估计来解调数据流$x_i[t]$。

[0472] 结果

[0473] 图48示出了具有和不具有频率偏移的DIDO2×2系统的SER结果。可以看到，所提出的方法完全消除了频率/相位偏移，产生了与不具有偏移的系统相同的SER。

[0474] 接下来，我们评估所提出的补偿方法对于频率偏移误差和/或实时偏移的波动的灵敏性。因此，我们将(14)重写为：

[0475] $$r_1[t]=e^{j2\pi\varepsilon_1 T_s(t-\hat{t}_1-\Delta t_1)}h_{11}[w_{11}x_1[t]+w_{21}x_2[t]]$$
$$+e^{j2\pi\varepsilon_2 T_s(t-\hat{t}_2-\Delta t_2)}h_{12}e^{-j(2\pi\varepsilon_1 T_s+2\pi\varepsilon_2 T_s)(t-\Delta t_2)}[w_{12}x_1[t]+w_{22}x_2[t]] \quad (25)$$

[0476] 其中，ε表示训练和数据发送之间的频率偏移的估计误差和/或变化。注意，ε的效果是破坏(13)中的正交特性，以使得(14)和(16)中的干扰项在发射机处没有完全被预先消除。因为这样，SER性能随着增大的ε值而降低。

[0477] 图48示出了对于不同的ε值的频率偏移补偿方法的SER性能。这些结果假设T_s=0.3ms(即具有3KHz带宽的信号)。我们观察到，对于ε=0.001Hz(或更少)，SER性能与没有偏移的情况相似。

[0478] f.基于时间和频率偏移估计的算法的一个实施方式的描述

[0479] 下面，我们描述执行时间和频率偏移估计的另外的实施方式(图47中的4701b)。考虑的发射信号结构在H.Minn,V.K.Bhargava和K.B.Letaief,"A robust timing and frequency synchronization for OFDM systems,"IEEE Trans.Wireless Commun.,vol.2,no.4,pp.822-839,july2003中提出，在K.Shi和E.Serpedin,"Coarse frame and carrier synchronization of OFDM systems:a new metric and comparison,"IEEE Trans.Wireless Commun.,vol.3,no.4,pp.1271-1284,July2004中被详细研究。通常具有好的相关属性的序列被用于训练。例如，对于我们的系统，Chu序列被使用，Chu序列如在D.Chu,"Polyphase codes with good periodic correlation properties(corresp.),"IEEE Trans.Inform.Theory,vol.18,no.4,pp.531-532,July1972中被描述。这些序列具有有趣的属性，即它们具有完美的循环相关。让L_{cp}表示循环前缀的长度，N_t表示分量训练序列的长度，使得$N_t=M_t$，其中M_t是训练序列的长度。在这些假设下，所发送的用于开端的符号序列可以被写成：

[0480] s[n]=t[n-N_t]对于n=-1,…,-L_{cp}

[0481] s[n]=t[n]对于n=0,…,N_t-1

[0482] s[n]=t[n-N_t]对于n=N_t,…,2N_t-1

[0483] s[n]=-t[n-2N_t]对于n=2N_t,…,3N_t-1

[0484] s[n]=-t[n-3N_t]对于n=3N_t,…,4N_t-1

[0485] 注意该训练信号的结构可以被扩展至其它长度，但重复了块结构。例如，为了使用16个训练信号，我们考虑一种结构，例如：

[0486] [CP,B,B,-B,B,B,B,-B,B,-B,-B,B,-B,B,B,-B,B,]。

[0487] 通过使用该结构，并使N_t=4M_t，所有将要描述的算法可以在没有修改的情况下被使用。有效地，我们重复训练序列。这在合适的训练信号可能不可用的情况下特别有用。

[0488] 在对符号率进行匹配的滤波和向下采样之后，考虑下面的所接收到的信号：

[0489] $$r[n]=e^{j2\pi\varepsilon n}\sum_{l=0}^{L}h[l]s[n-l-\Delta]+v[n]$$

[0490] 其中ε是未知的离散时间频率偏移，Δ是未知的帧偏移，h[l]是未知的离散时间信道系数，以及v[n]是附加噪声。为了解释以下部分中的关键思想，忽略附加噪声的存在。

[0491] 1.粗略的帧同步

[0492] 粗略的帧同步的目的是解决未知的帧偏移Δ。让我们做出以下定义：

[0493] $r_1[n] := [r[n], r[n+1], \ldots, r[n+N_c-1]]^T$,

[0494] $\tilde{r}_1[n] := [r[n+L_{cp}], r[n+1], \ldots, r[n+N_c-1]]^T$,

[0495] $r_2[n] := [r[n+N_c], r[n+1+N_c], \ldots, r[n+2N_c-1]]^T$,

[0496] $\tilde{r}_2[n] := [r[n+L_{cp}+N_c], r[n+1+L_{cp}+N_c], \ldots, r[n+L_{cp}+2N_c-1]]^T$,

[0497] $r_3[n] := [r[n+2N_c], r[n+1+2N_c], \ldots, r[n+3N_c-1]]^T$,

[0498] $\tilde{r}_3[n] := [r[n+L_{cp}+2N_c], r[n+L_{cp}+1+2N_c], \ldots, r[n+L_{cp}+3N_c-1]]^T$,

[0499] $r_4[n] := [r[n+3N_c], r[n+1+3N_c], \ldots, r[n+4N_c-1]]^T$,

[0500] $\tilde{r}_4[n] := [r[n+L_{cp}+3N_c], r[n+L_{cp}+1+3N_c], \ldots, r[n+L_{cp}+4N_c-1]]^T$.

[0501] 所提出的粗略的帧同步算法从K.Shi和E.Serpedin,"Coarse frame and carrier synchronization of OFDM systems:a new metric and comparison,"IEEE Trans.Wireless Commun.,vol.3,no.4,pp.1271-1284,July2004中的算法得到启示,根据最大似然准则获得。

[0502] 方法1-改进的粗略的帧同步:粗略的帧同步估计器解决了以下优化:

[0503] $$\hat{\Delta} = \max_{n\in\mathbb{Z}} \frac{|P_1[n]| + |P_2[n]| + |P_3[n]|}{\|r_1\|^2 + \|r_2\|^2 + \|r_3\|^2 + \|r_4\|^2 + \frac{1}{2}(\|r_1\|^2 + \|r_2\|^2 + \|r_3\|^2 + \|r_4\|^2)}$$

[0504] 其中,

[0505] $P_1[n] = r_1^*[n]r_2[n] - r_2^*[n]r_4[n] - \tilde{r}_2^*[n]\tilde{r}_3[n]$

[0506] $P_2[n] = r_2^*[n]r_1[n] - r_1^*[n]r_3[n]$

[0507] $P_3[n] = \tilde{r}_1^*[n]\tilde{r}_3[n]$.

[0508] 使得被校正的信号被定义为:

[0509] $$r_c[n] = r[n - \hat{\Delta} - \lceil L_{cp}/4 \rceil]$$

[0510] 另外的校正项被用于补偿信道中的小的初始脉冲并可以基于应用被调节。这个额外的延迟将之后被包括在信道中。

[0511] ii.分数频率偏移校正

[0512] 分数频率偏移校正在粗略的帧同步块之后。

[0513] 方法2-改进的分数频率偏移校正:分数频率偏移是下面的解:

[0514] $$\hat{\varepsilon}_f = \frac{\text{相位 } P_1[\hat{\Delta}]}{2\pi N_c}$$

[0515] 这被已知为分数频率偏移,因为算法仅可以校正偏移

[0516] $$|\hat{\varepsilon}_f| < \frac{1}{2N_c}$$

[0517] 这个问题将在下一部分中被解决。此精细频率偏移校正信号被定义为:

[0518] $$r_c[n] = e^{-j2\pi\hat{\varepsilon}_f} r_c[n]$$

[0519] 注意,方法1和2是对于在频率选择信道中工作较好的K.Shi,E.Serpedin,"Coarse frame and carrier synchronization of OFDM systems:a new metric and

comparison," IEEE Trans. Wireless Commun., vol. 3, no. 4, PP. 1271-1284, July 2004的改进。这里的一个特别创新是使用了上面所述的r和$\tilde{r}$。$\tilde{r}$的使用改进了以前的估计器，因为它忽略了由于内部符号干扰而被影响的取样。

[0520] iii. 整数频率偏移校正

[0521] 为了校正整数频率偏移，有必要写一个用于在精细频率偏移校正之后所接收到的信号的等价系统模型。将保留的定时误差吸收到信道中，没有噪声的所接收到的信号具有以下结构：

[0522] $$r_f[n] = e^{j2\pi \frac{\varepsilon}{N_c} n} \sum_{l=0}^{L_{cp}} g[l] s[n-l]$$

[0523] 其中n=0, 1, ..., $4N_f$-1。整数频率偏移是k，而未知的等价信道是g[l]。

[0524] 方法3-改进的整数频率偏移校正：整数频率偏移是以下的解：

[0525] $$\hat{k} = \arg \max_{k=0,1,\ldots,N_f-1} \mathbf{r}^* \mathbf{D}[k] \mathbf{S} (\mathbf{S}^* \mathbf{S})^{-1} \mathbf{S}^* \mathbf{D}[k]^* \mathbf{r}$$

[0526] 其中：

[0527] r=D[k]Sg

[0528] $$\mathbf{D}[k] := \mathrm{diag}\left\{1, e^{j2\pi \frac{k}{N_f}}, \ldots, e^{j2\pi \frac{(4N_f-1)k}{N_f}}\right\}$$

[0529] $$\mathbf{S} := \begin{bmatrix} s[0] & s[-1] & \cdots & \cdots & s[-L_{cp}] \\ s[1] & s[0] & s[-1] & \cdots & s[-L_{cp}+1] \\ s[4N_f-1] & s[4N_f-2] & s[4N_f-3] & \cdots & s[4N_f-1-L_{cp}] \end{bmatrix}$$

[0530] $$\mathbf{g} := \begin{bmatrix} g[0] \\ g[1] \\ \vdots \\ g[L_{cp}] \end{bmatrix}$$

[0531] 这给出了总的频率偏移的估计：

[0532] $$\hat{\varepsilon} = \frac{\hat{k}}{N_f} + \hat{\varepsilon}_f$$

[0533] 实际上，方法3具有很高的复杂性。为了降低复杂性，可以做出以下观察。首先，乘积S(S*S)$^{-1}$S可以被预计算。遗憾的是，这仍然留下了相当大的矩阵乘法。可替换的是采用具有所提出的训练序列的观察，S*S≈I。这产生了以下的降低的负载型的方法。

[0534] 方法4-低复杂性的改进的整数频率偏移校正：

[0535] 低复杂性的整数频率偏移估计器解出了

[0536] $$\hat{k} = \arg \max_{k=0,1,\ldots,N_f-1} (\mathbf{S}^* \mathbf{D}[k]^* \mathbf{r})^* (\mathbf{S}^* \mathbf{D}[k]^* \mathbf{r})$$

[0537] iv. 结果

[0538] 在该部分中，我们比较了不同的所提出的估计器的性能。

[0539] 首先，在图50中，我们比较了每种方法所需要的开销的量。注意两种新的方法将开销降低了10倍到20倍。为了比较不同的估计器的性能，MonteCarlo实验被执行。所考虑的设置是从具有3K符号每秒的符号率的线性调制构造的我们的通常的NVIS发送波形，对应于3KHz的通带带宽，以及上升的余弦脉冲成形。对于每个Monte Carlo实现，频率偏移从在$[-f_{max}, f_{max}]$上的均匀分布画生成。

[0540] 具有f_{max}=2Hz的小的频率偏移且没有整数偏移校正的仿真在图51中被示出。可以从该性能比较中看出，具有N_c/M_c=1的性能从原始估计器轻微降级，虽然实质上降低了开销，具有N_c/M_c=4的性能更好，几乎是10dB。由于在整数偏移估计中的误差，所有曲线在低的SNR点经历了曲折。在整数偏移中的小的误差可以创建大的频率误差和大的拼接平方误差。整数偏移校正可以在小的偏移中被关闭以改进性能。

[0541] 在多路径信道存在的情况下，频率偏移估计器的性能一般降低。然而，在图52中，关闭整数偏移估计器展现了非常好的性能。因此，在多路径信道中，在执行鲁棒的粗略校正之后的改进的精细校正算法是更重要的。注意，具有N_c/M_c=4的偏移性能在多路径情况下好得多。

[0542] 本发明的实施方式可以包括上面所提出的各种步骤。所述步骤可以以机器可执行指令的方式实现，所述指令使得通用或专用处理器执行特定步骤。例如，在上面所述的基站/AP和客户装置内的各种组件可以被实施为在通用或专用处理器上执行的软件。为了避免模糊本发明的有关方面，诸如计算机存储器、硬盘、输入装置等的各种公知的个人计算机组件已经从图中省去。

[0543] 可替换地，在一个实施方式中，这里所示出的各种功能模块和相关步骤可以通过包含执行步骤的硬接线逻辑的专用硬件组件（例如专用集成电路（ASIC））或通过编程计算机组件和定制硬件组件的任意组合而被执行。

[0544] 在一个实施方式中，诸如上面描述的编码、调制和信号处理逻辑903的特定模块可以在可编程数字信号处理器（DSP）（或DSP组）上被实施，所述DSP例如使用德州仪器（Texas Instruments）的TMS320x架构的DSP（例如，TMS320C6000、TMS320C5000等）。在该实施方式中的DSP可以被嵌入在个人计算机的插件卡中，例如，PCI卡。当然，在符合本发明的基本原理的情况下，各种不同的DSP架构可以被使用。

[0545] 本发明的各种部件也可以被提供为用于存储机器可执行指令的机器可读介质。机器可读介质可以包括但不限于闪存、光盘、CD-ROM、DVD ROM、RAM、EPROM、EEPROM、磁卡或光学卡、传播媒介或适于存储电子指令的其它类型的机器可读介质。例如，本发明可以被下载为计算机程序，该计算机程序可以通过包含在载波或其它传播媒介的数据信号的方式经由通信链路（例如，调制解调器或网络连接）从远程计算机（例如服务器）传送到请求计算机（例如客户端）。

[0546] 遍及前述描述，为了解释的目的，许多特定细节被提出以提供本系统和方法的全面理解。然而，对于本领域技术人员来说显而易见的是，系统和方法可以在没有这些特定细节中的一些的情况下被实现。因此，本发明的范围和实质应当根据所附权利要求被判断。

[0547] 此外，在前述描述中，许多文献被引用以提供本发明的更全面的理解。所有这些引用的参考文献通过参考被合并到本申请中。

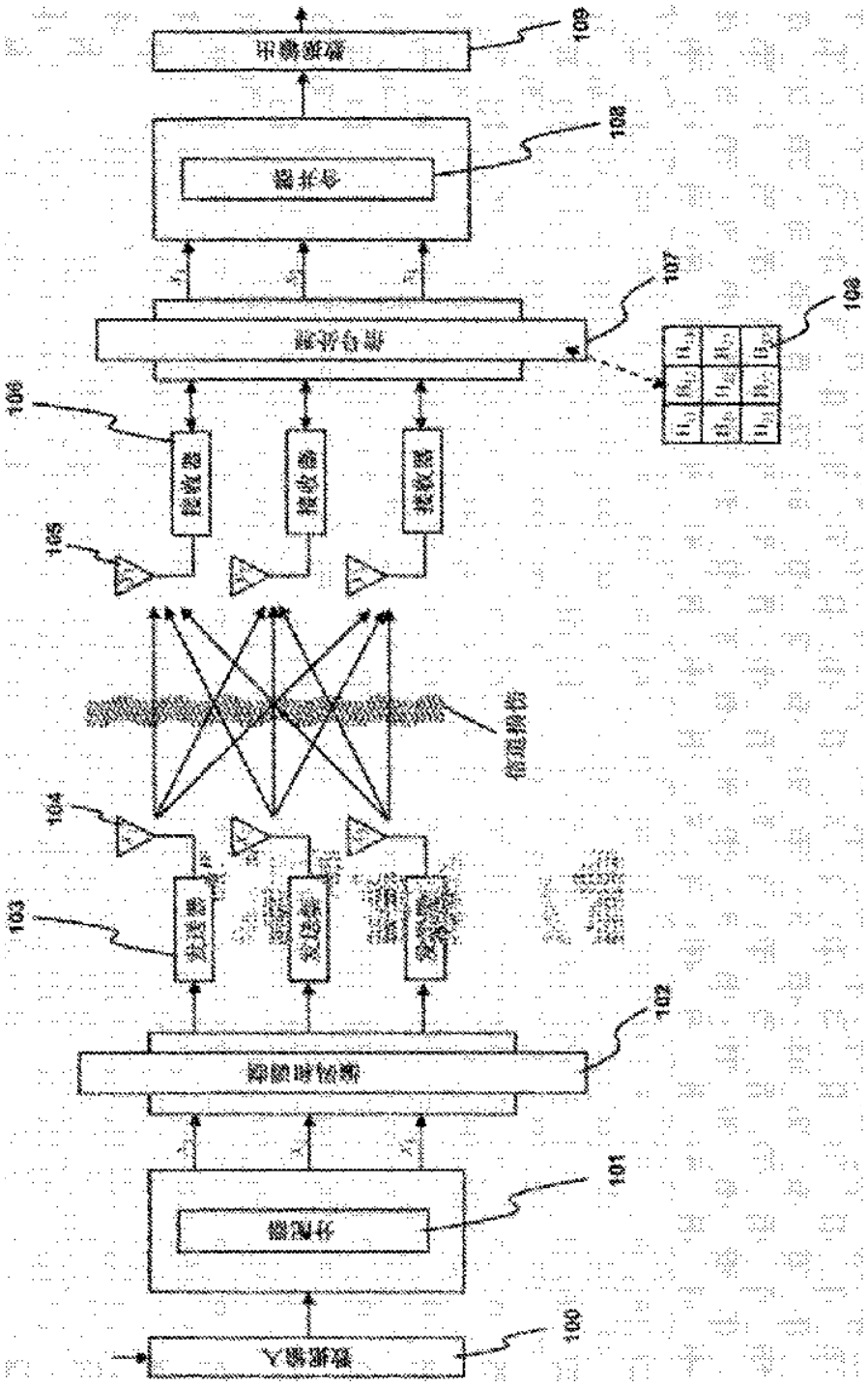


图1(现有技术)

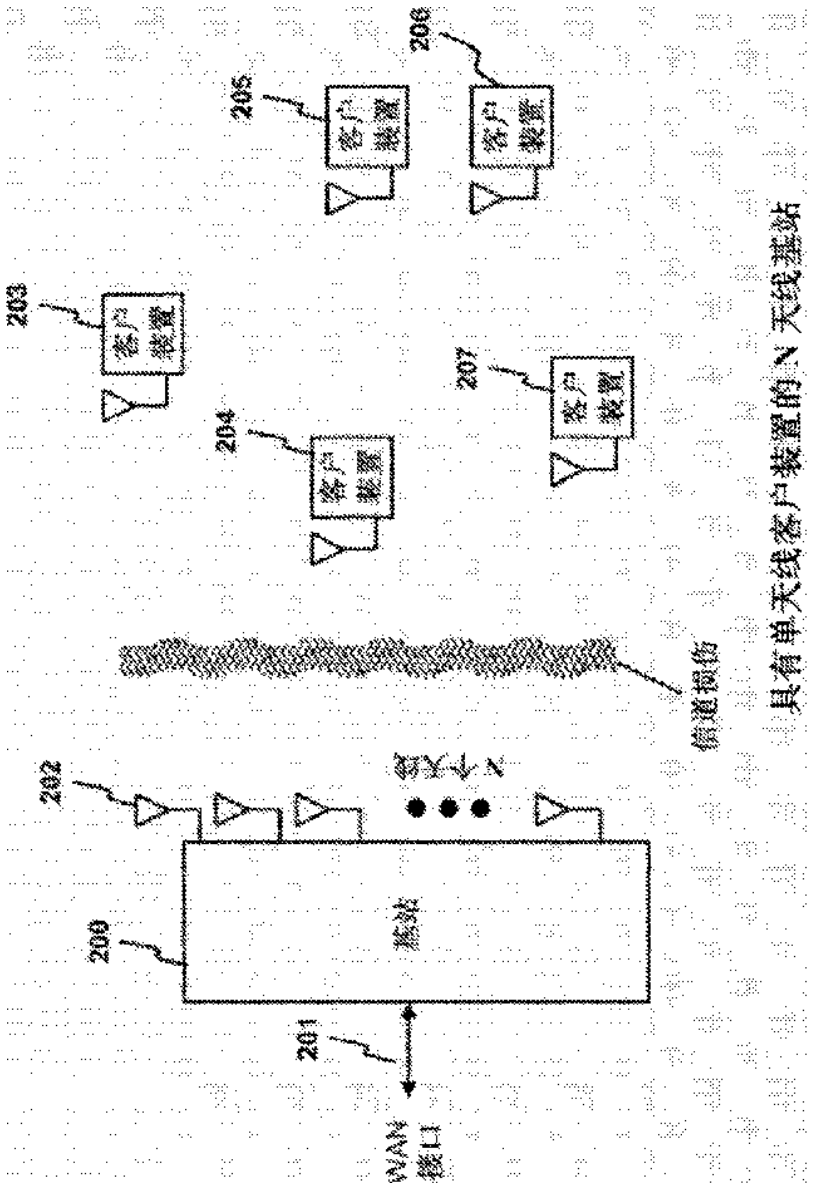

206
205
客户装置
203
207
204
具有单天线客户装置的 N 天线基站
信道损伤
202
N 天线
209
基站
201
WAN 接口

图2

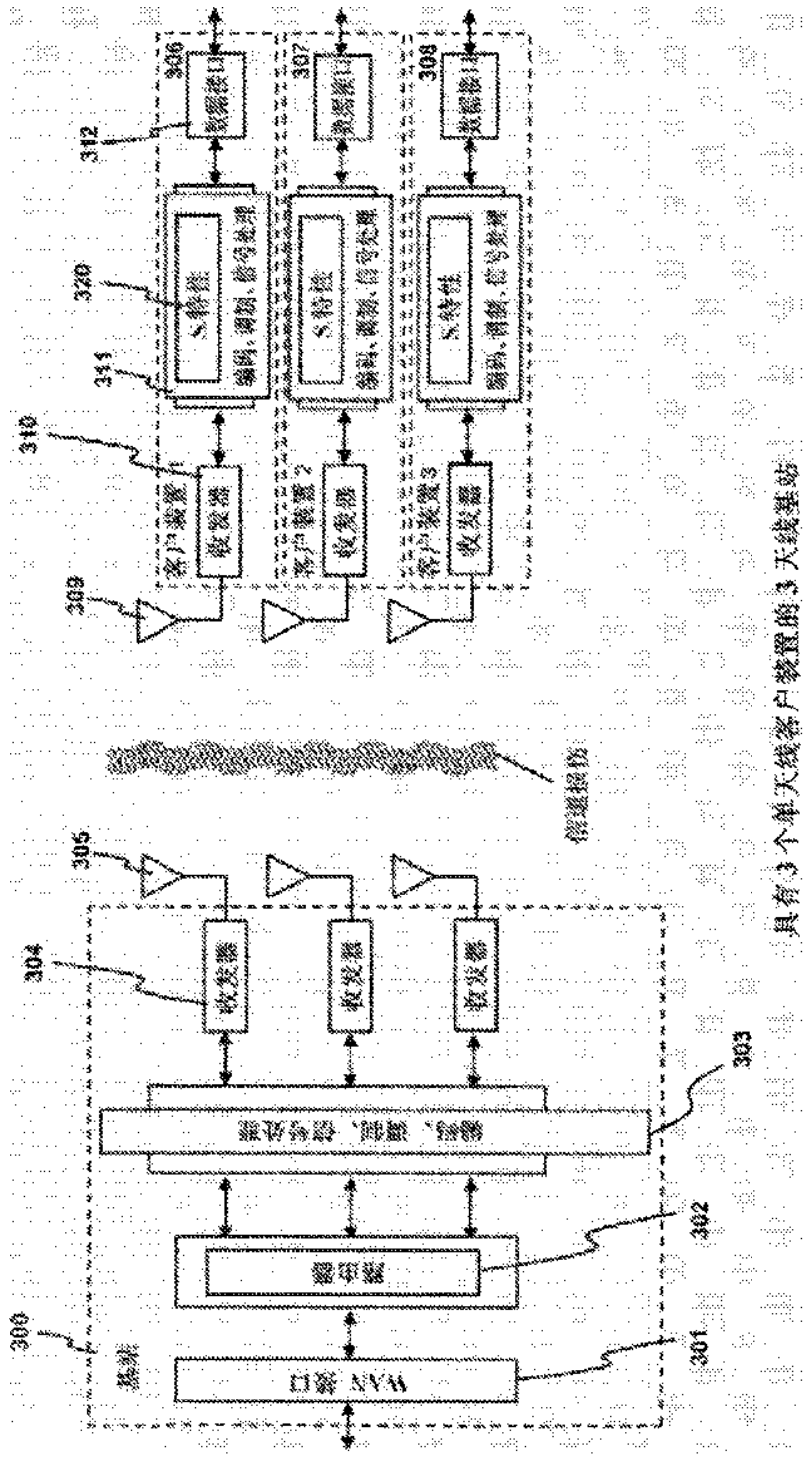

300
基站
301
MAN 接口
302
调度器
303
编码、调制、信号处理
304
收发器
305
306
307
308
309
310
311
320
312
S特性
客户装置1
客户装置2
客户装置3
微波接口
微波链路
具有3个单天线客户装置的3天线基站

图3

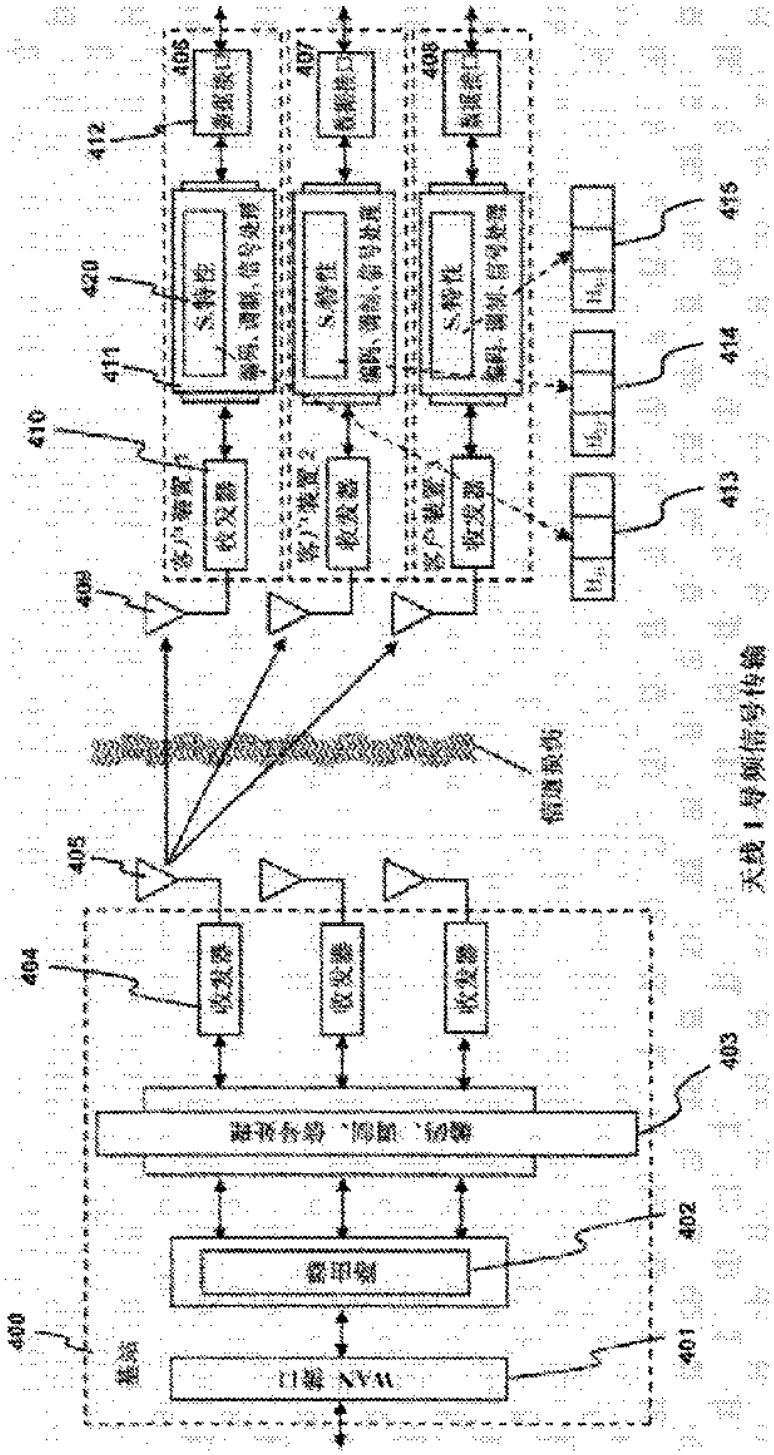

406
407
408
412
420
411
410
415
414
413
409
405
404
403
402
401
400
收发器
收发器
收发器
客户设备1
客户设备2
客户设备3
S-特性
S-特性
S-特性
编码、调制、信号处理
编码、调制、信号处理
编码、调制、信号处理
管道表面
天线1每侧信号传输
MAC
PHY
接口

图4

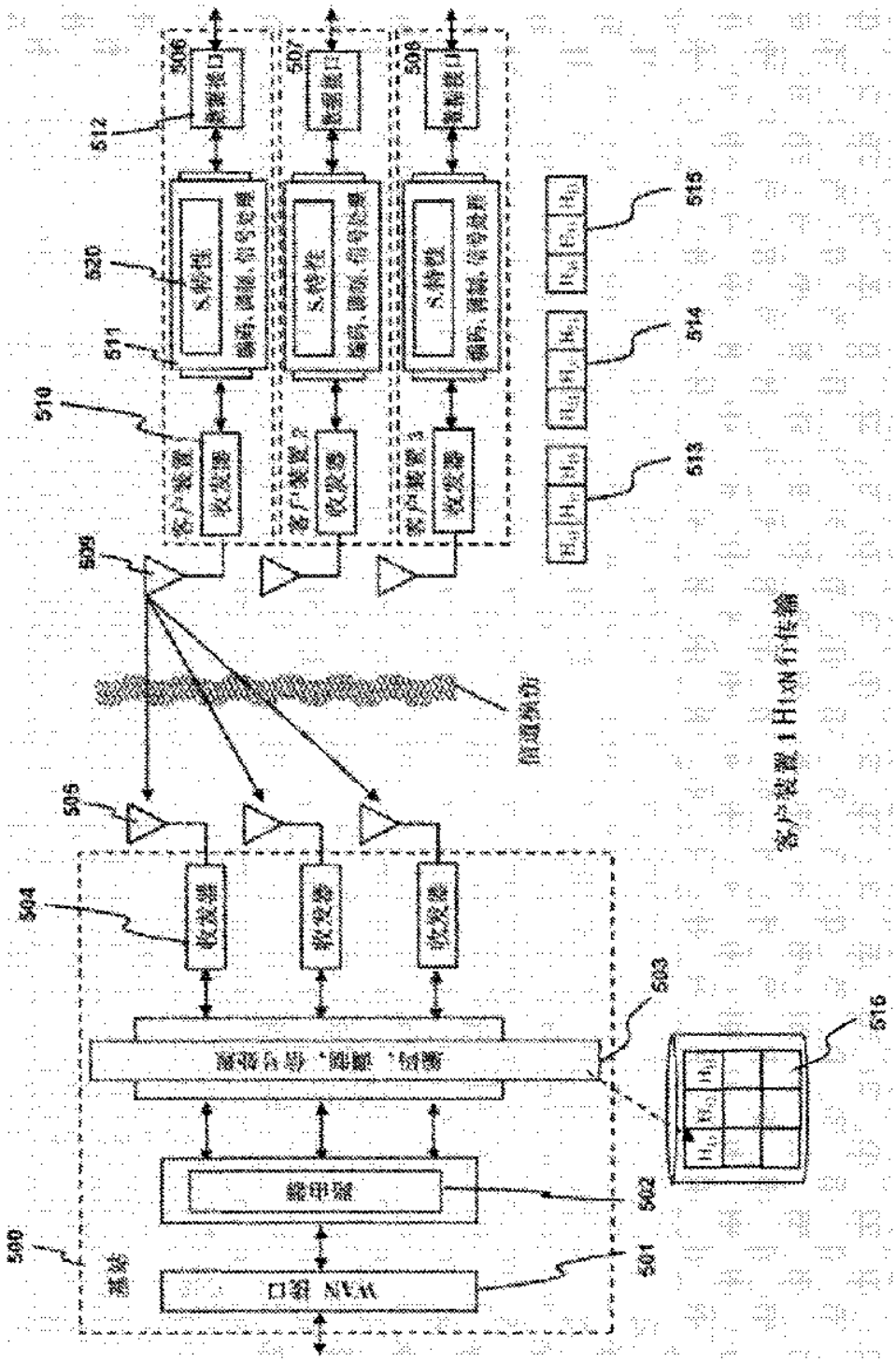

500
501
502
503
504
505
506
507
508
510
511
512
513
514
515
516
520
基站
NVM
客户设备1
客户设备2
客户设备3
S-特征
收发器
天线
信道衰落

图5

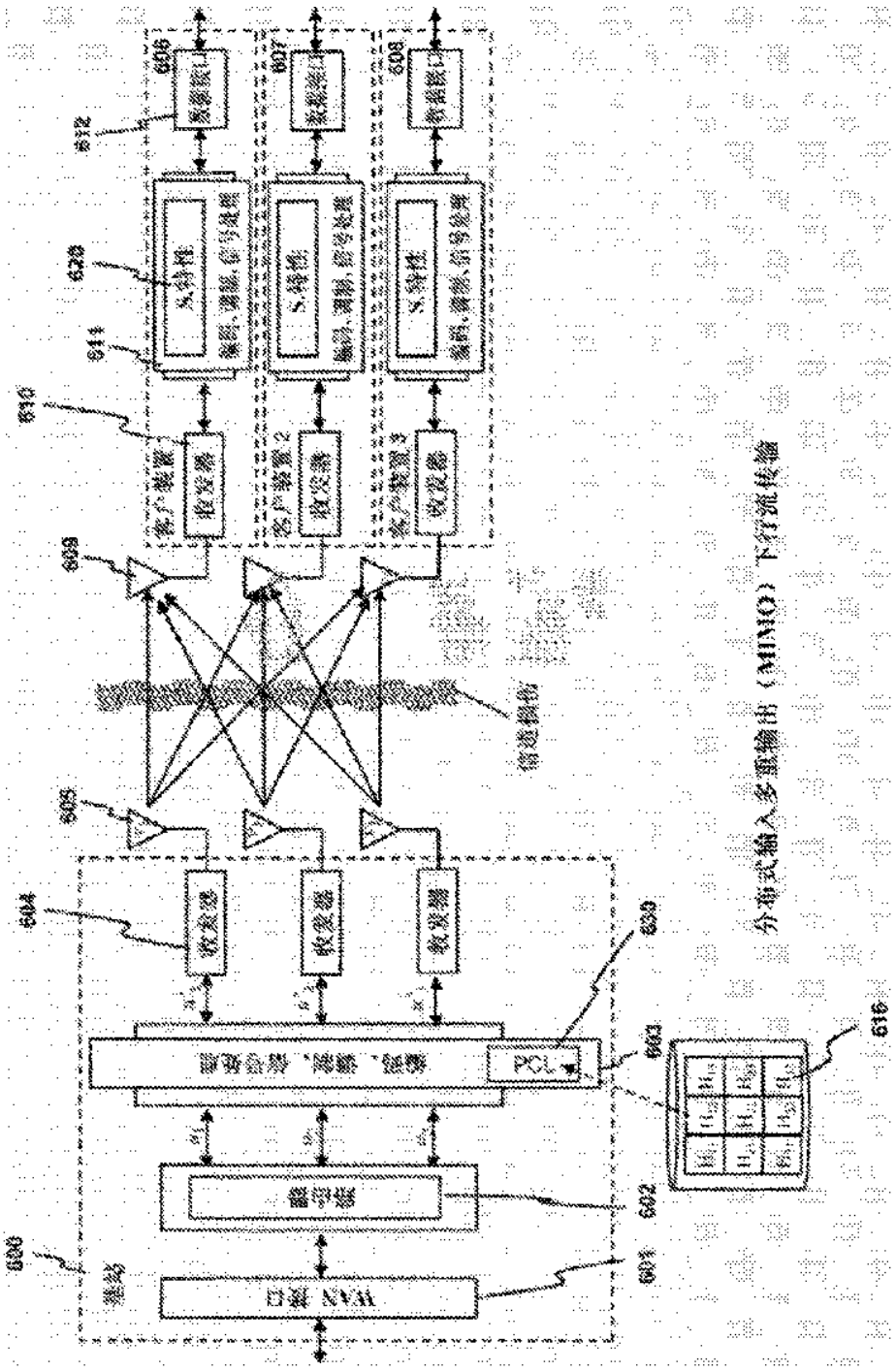


图6

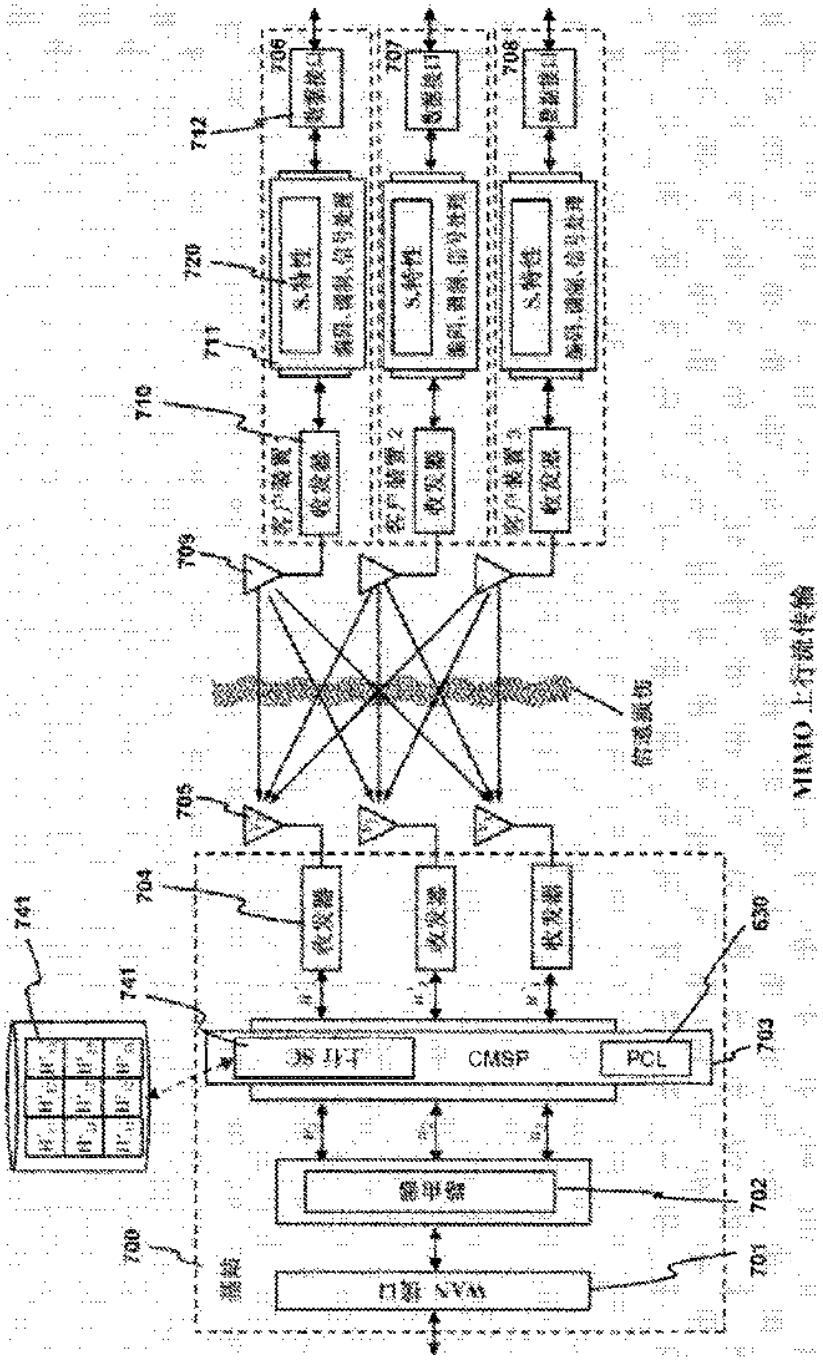

706
707
708
712
720
711
710
709
705
704
741
630
703
702
701
CMSP
PCL
MIMO上行流传输

图7

不同客户群的噪声

图8

基于延伸的客户分组

图9

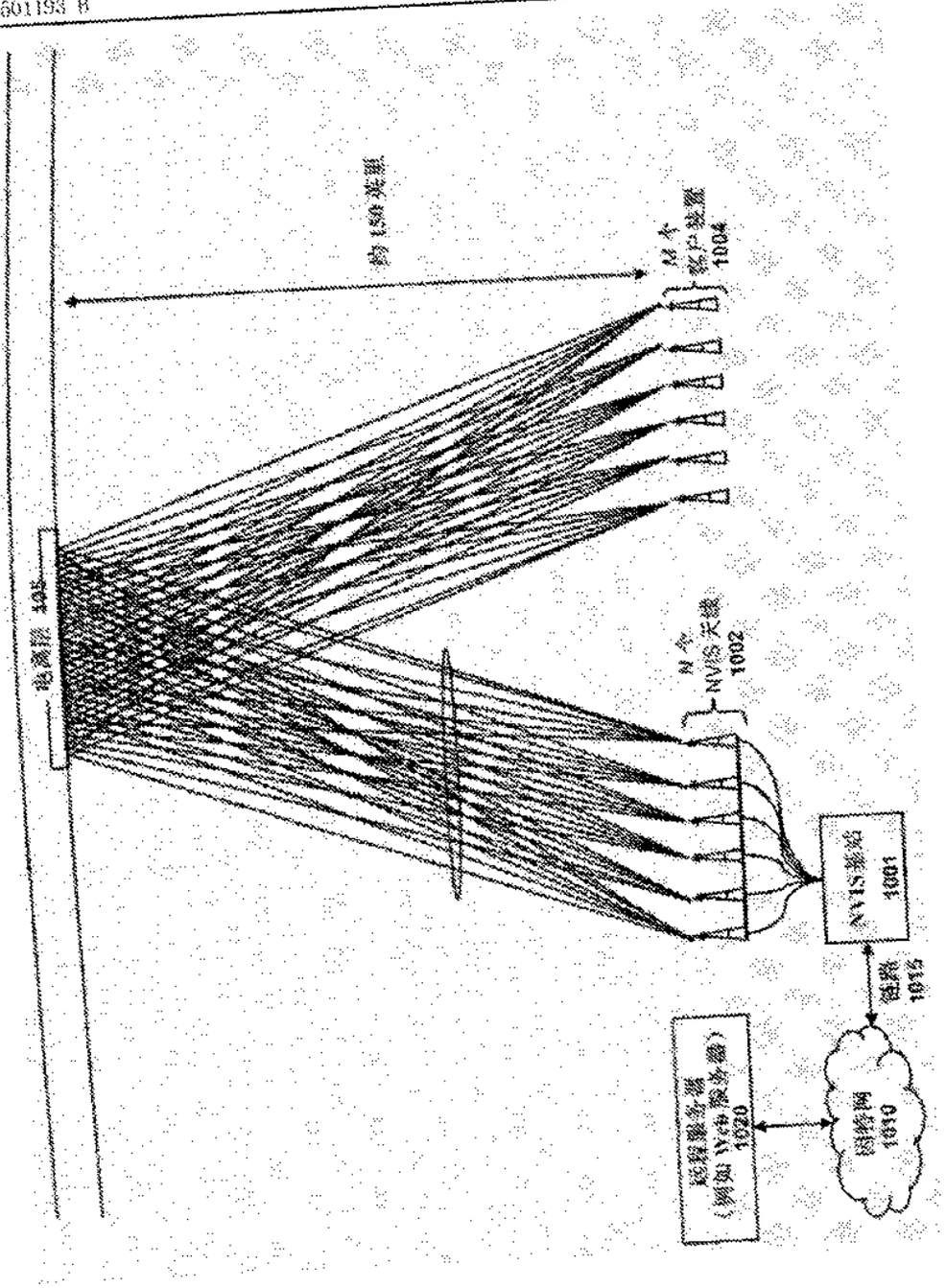

约 150 英里
M 个
客户装置
1004
地表层 105
N 个
NVIS 天线
1002
NVIS 基站
1001
链路
1015
因特网
1010
远程服务器
1020

图10

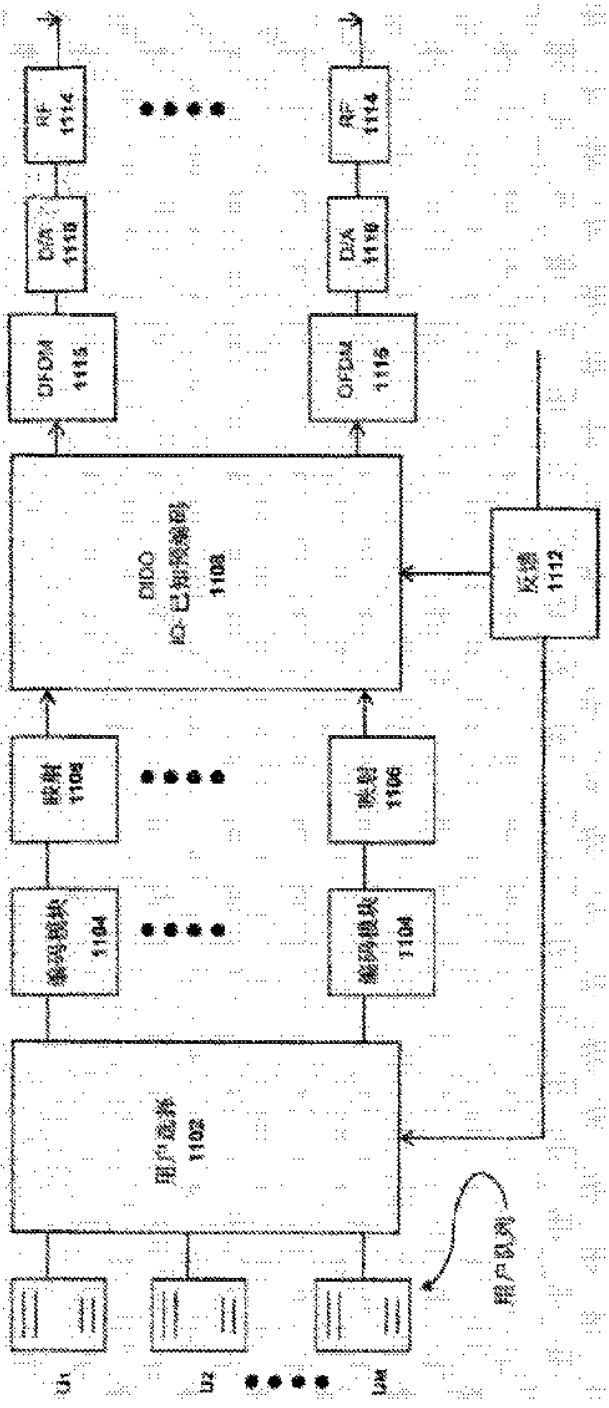

RF 1114
RF 1114
D/A 1118
D/A 1118
OFDM 1116
OFDM 1116
MIMO 1108
1112
1106
1106
1104
1104
1102
用户数据
U1
U2
UM

图11

61

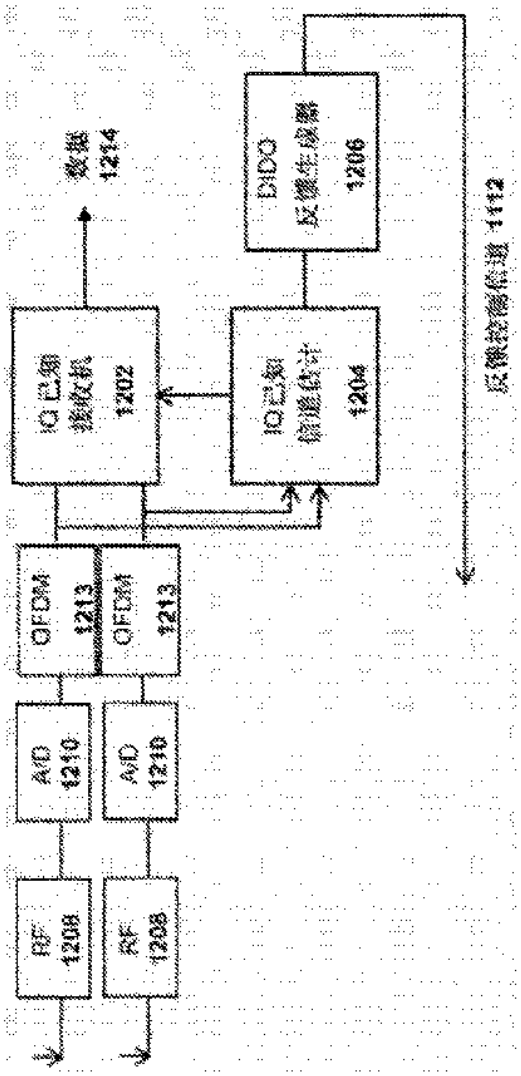

数据 1214
DIDO 反馈生成器 1206
反馈控制信道 1112
IQ 已知 接收机 1202
IQ 已知 信道估计 1204
OFDM 1213
OFDM 1213
A/D 1210
A/D 1210
RF 1208
RF 1208

图12

接入点
IQ-DIDO
预编码器
W
1302
$s_1[k]$
$s_1[k-\lambda]$
$s_M[k-\lambda]$
反馈信道
信道
1304
信道估计
1306
用户 m
ZF,MMSE
或 ML
检测器
1308
$\hat{s}_m[k]$
$\hat{s}_m[k-\lambda]$

图13

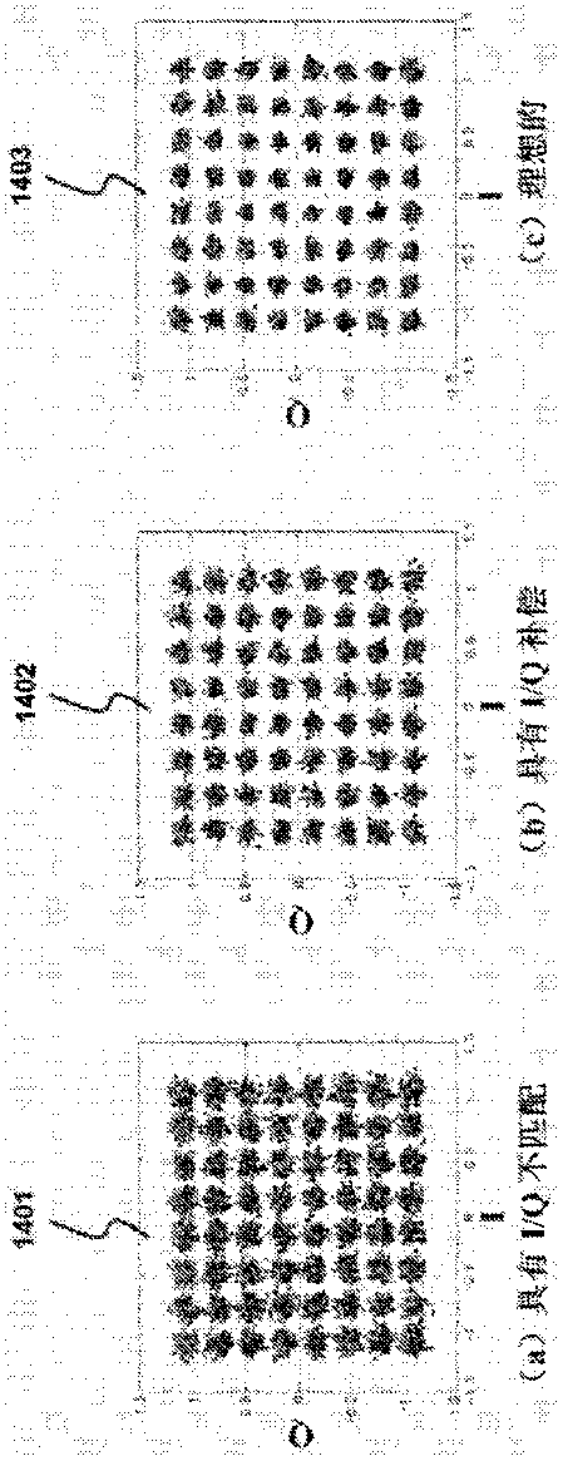


图14

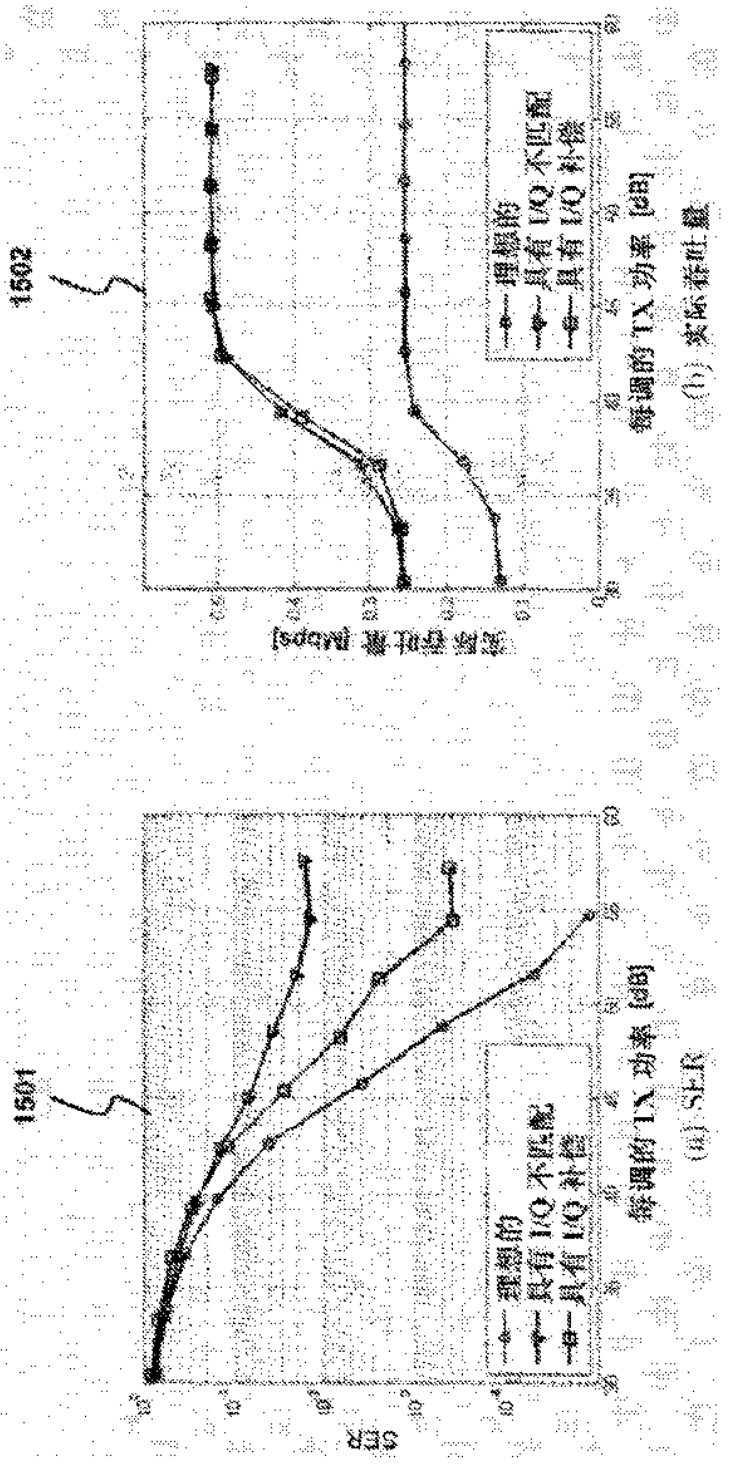

1501
SER
每调的 TX 功率 [dB]
(a) SER
理想的
具有 I/Q 不匹配
具有 I/Q 补偿
1502
实际吞吐量 [Mbps]
每调的 TX 功率 [dB]
(b) 实际吞吐量
理想的
具有 I/Q 不匹配
具有 I/Q 补偿

图15

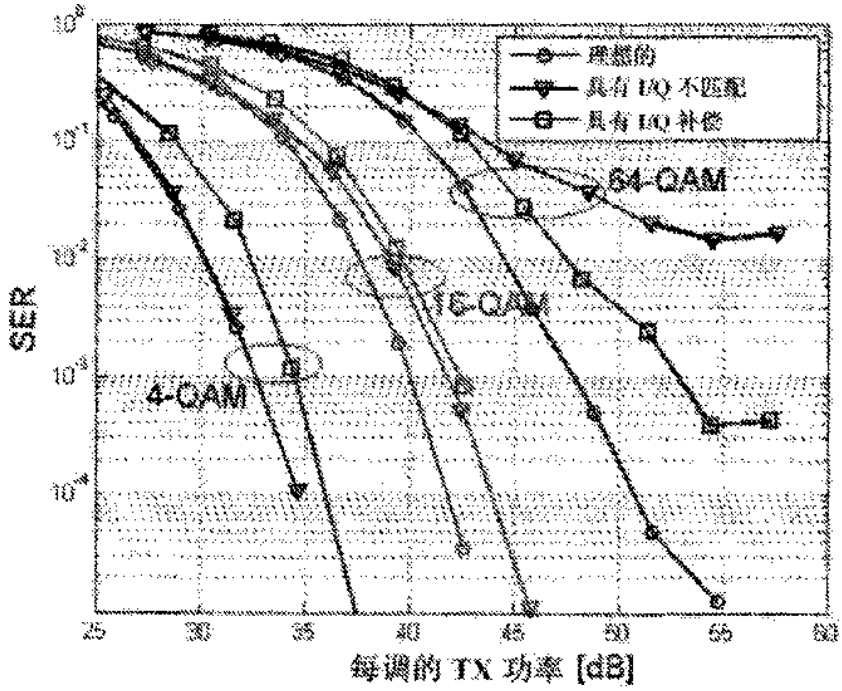


图16

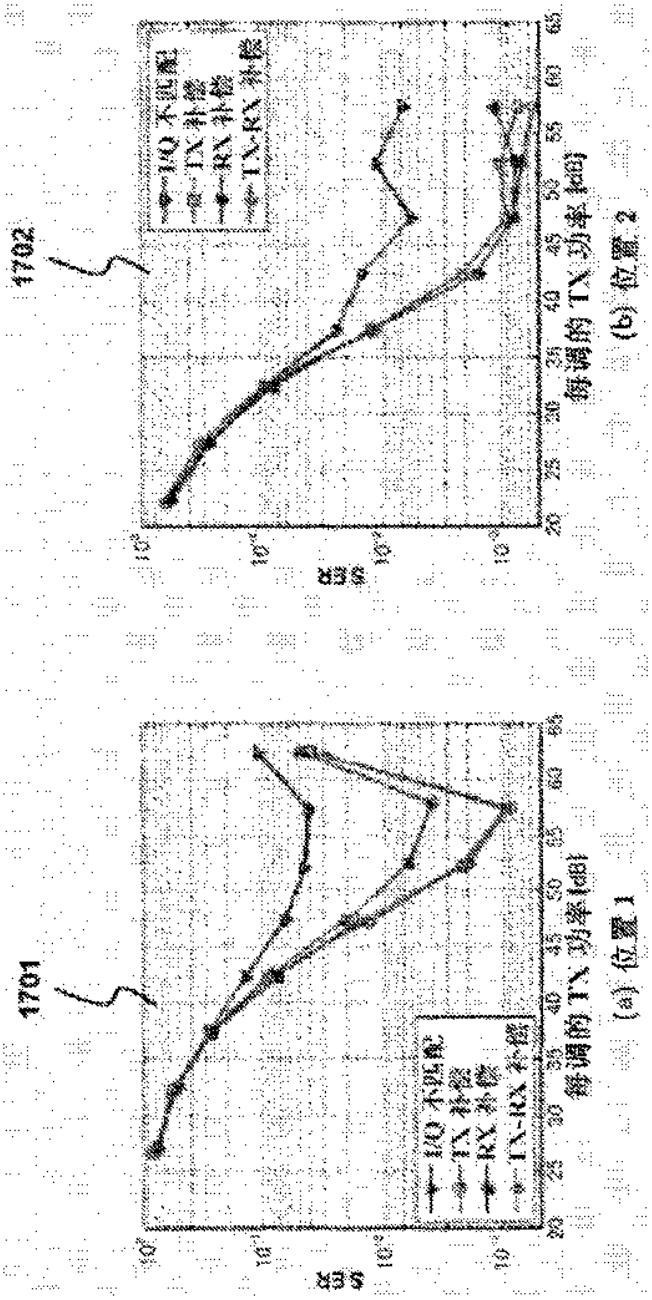


图17

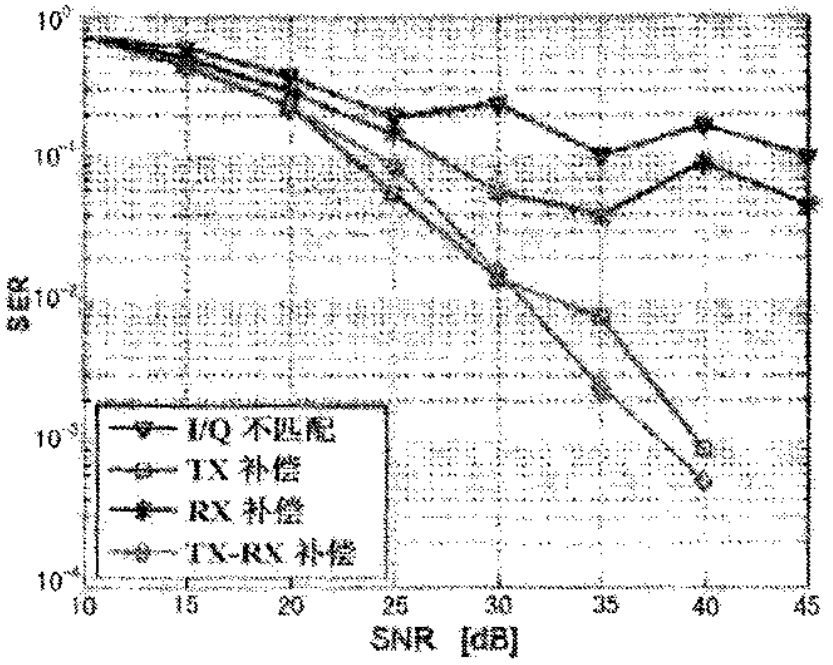


图18

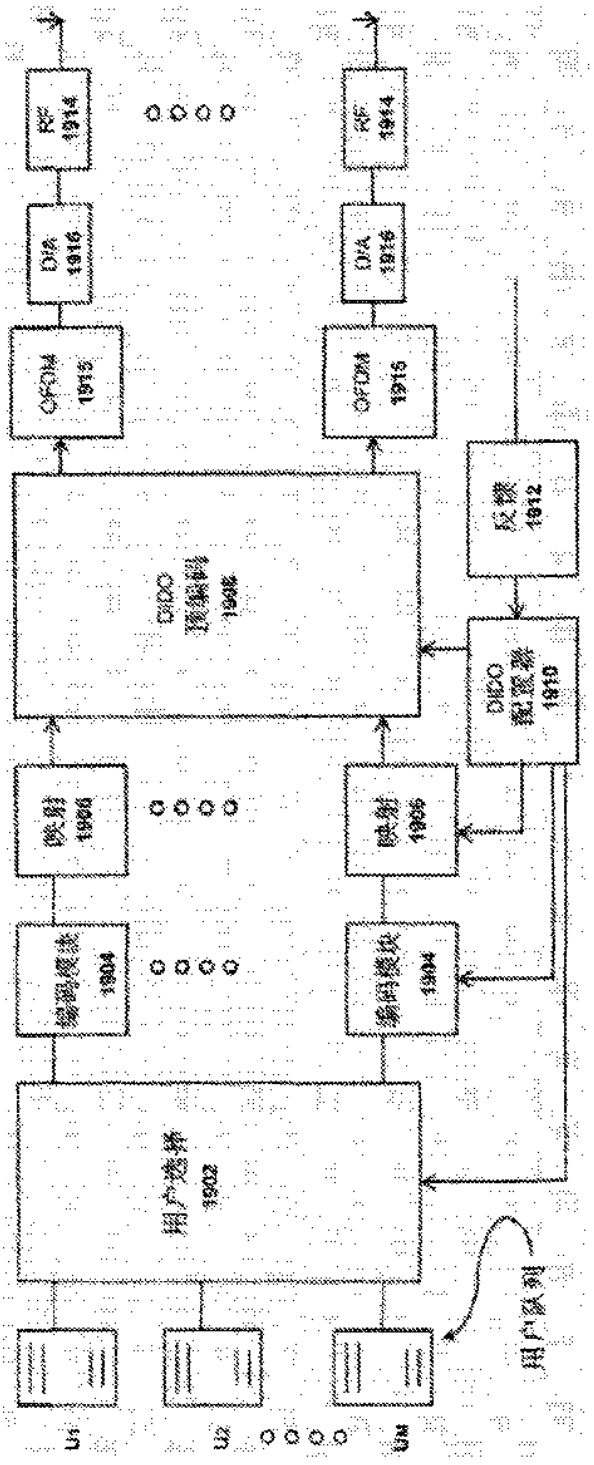

RF 1914
RF 1914
D/A 1916
D/A 1916
OFDM 1915
OFDM 1915
DIDO 预编码 1908
反馈 1912
DIDO 预编码 1910
映射 1906
映射 1906
编码模块 1904
编码模块 1904
用户选择 1902
用户队列
U1
U2
UM

图19

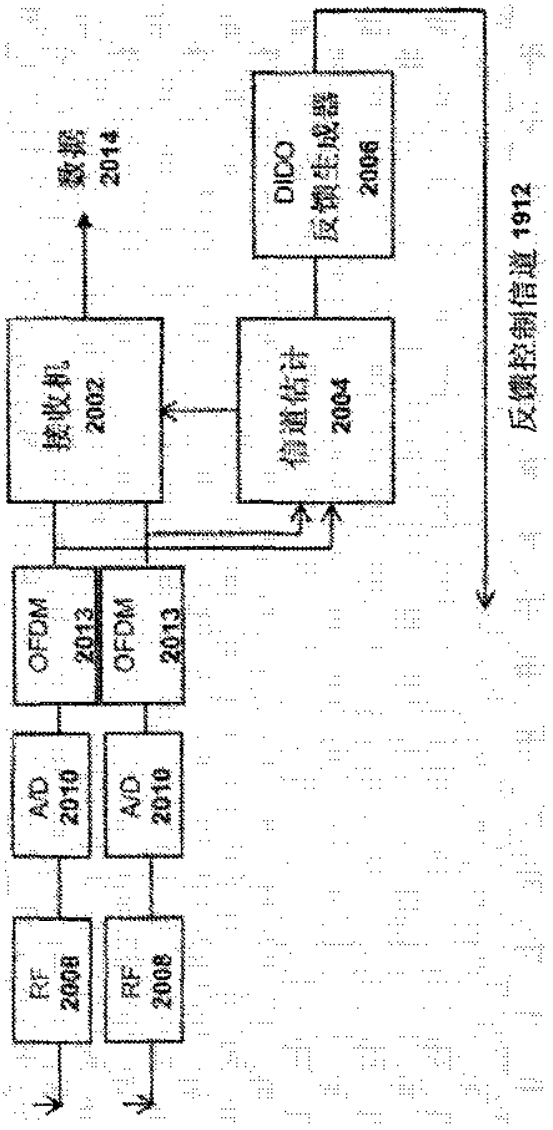

数据 2014
DICO 反馈生成器 2006
接收机 2002
信道估计 2004
反馈控制信道 1912
OFDM 2012
OFDM 2013
A/D 2010
A/D 2010
RF 2008
RF 2008

图20

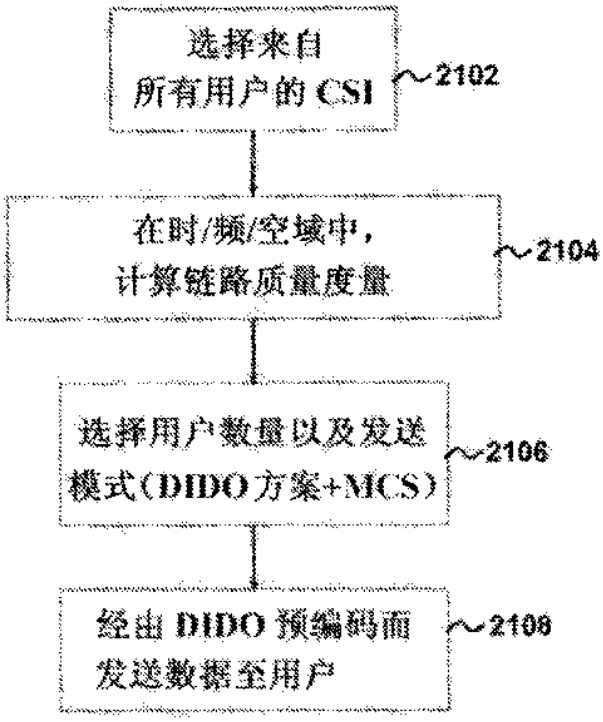

选择来自
所有用户的 CSI
2102
在时/频/空域中，
计算链路质量度量
2104
选择用户数量以及发送
模式(DIDO方案+MCS)
2106
经由 DIDO 预编码而
发送数据至用户
2108

图21

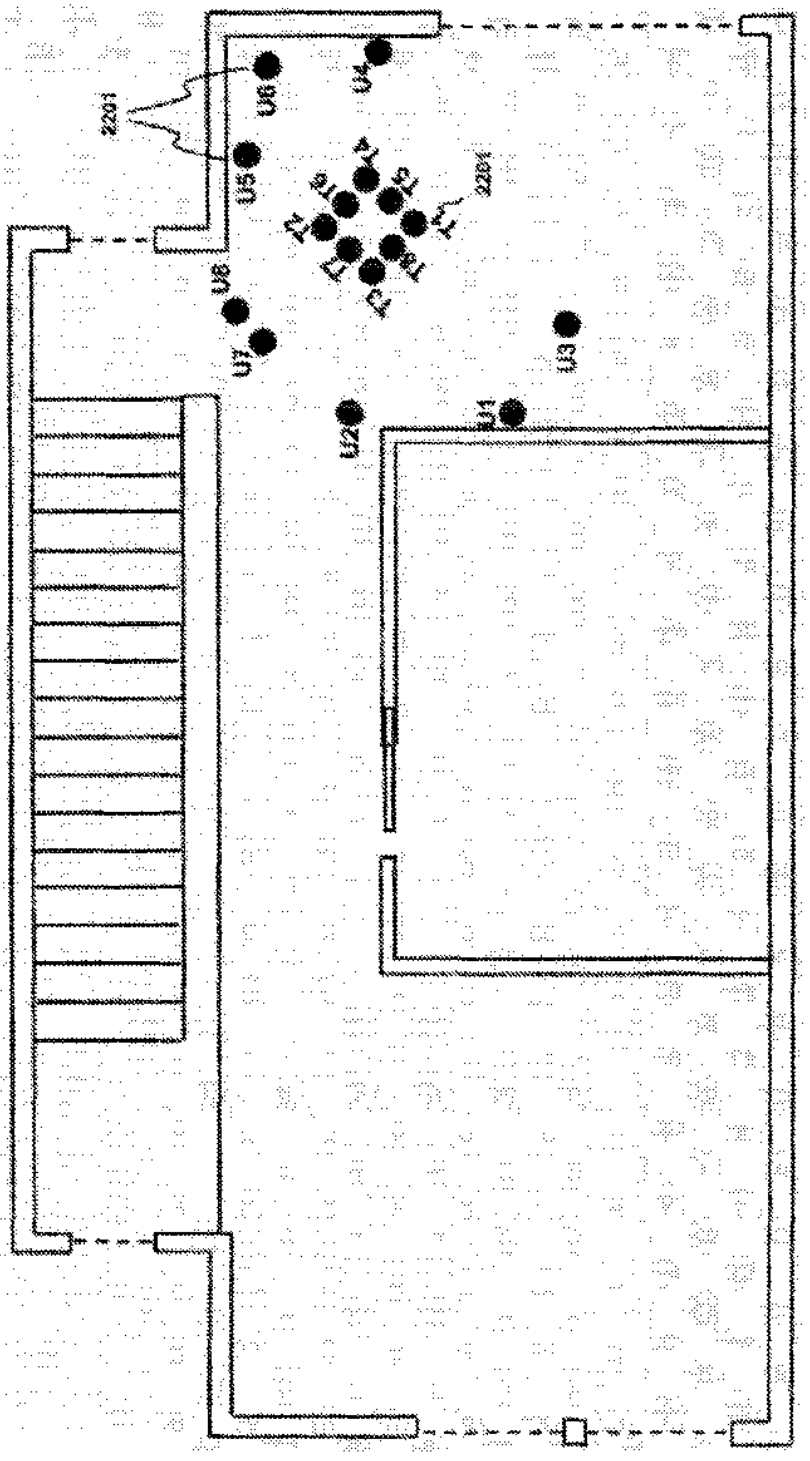


图22

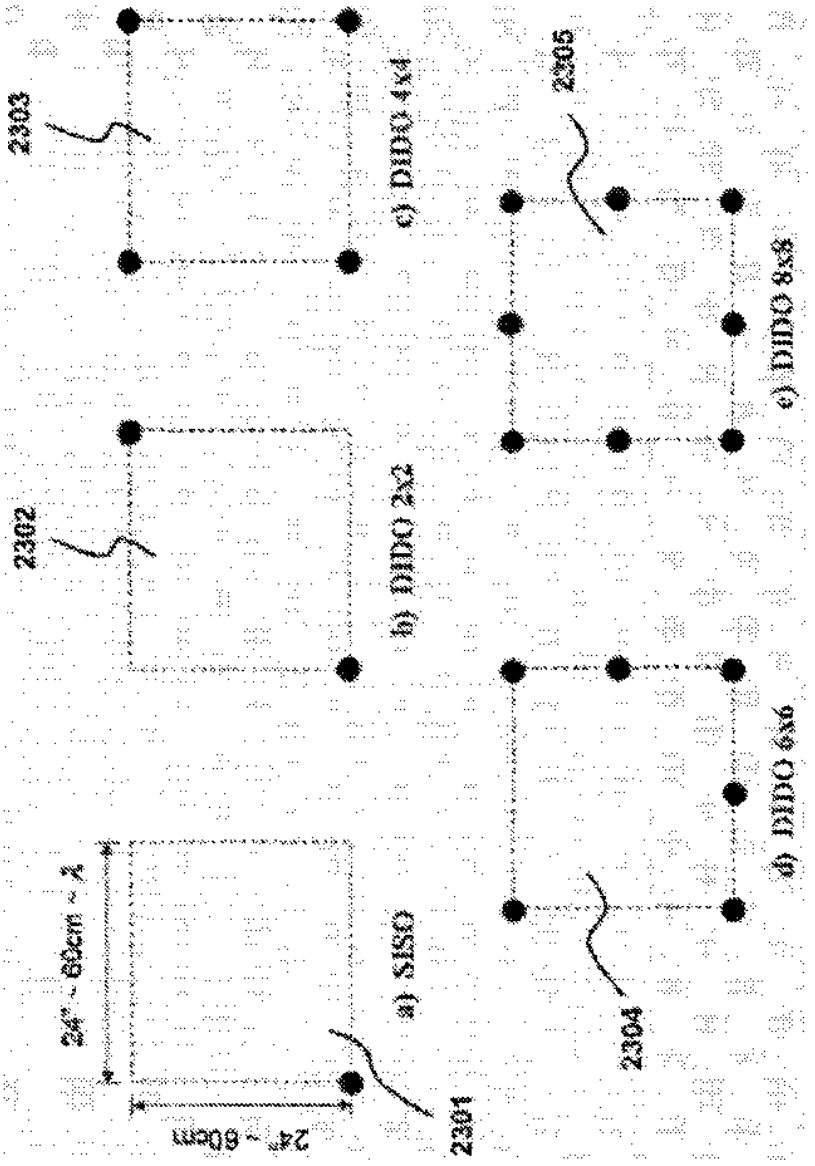

a) SISO
b) DIDO 2x2
c) DIDO 4x4
d) DIDO 6x6
e) DIDO 8x8
2301
2302
2303
2304
2305
24" ~ 60cm ~ λ
24" ~ 60cm

图23

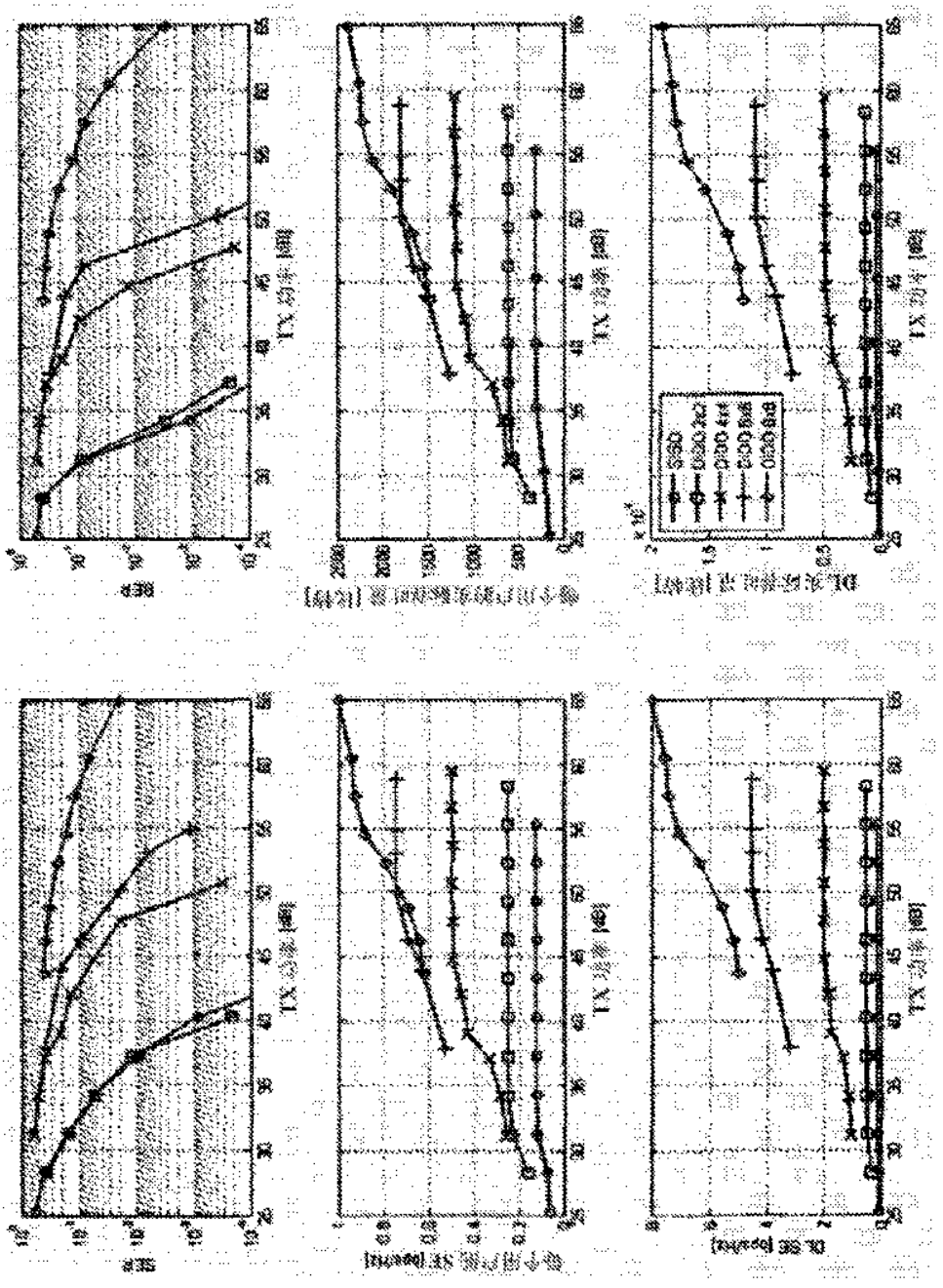


图24

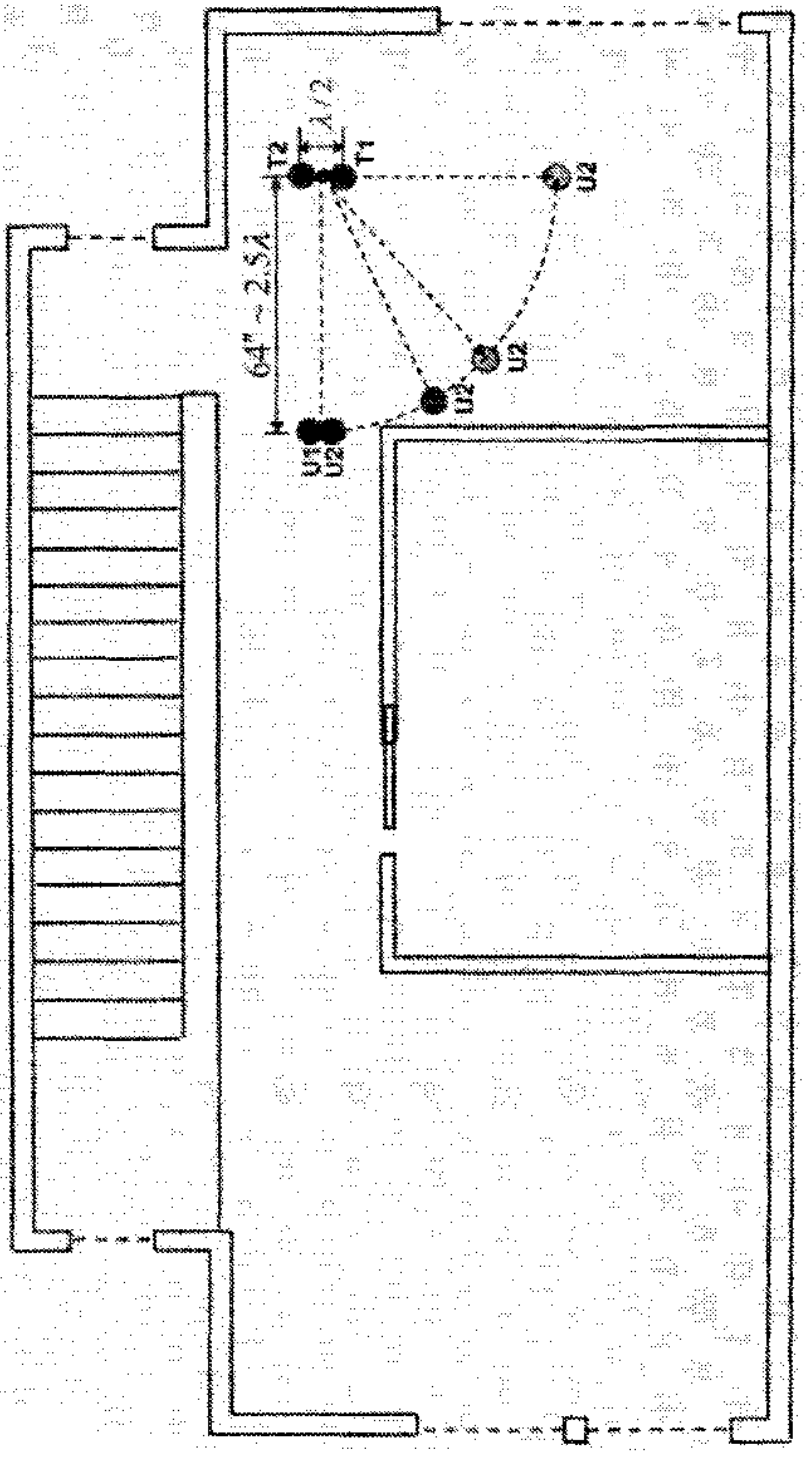


图25

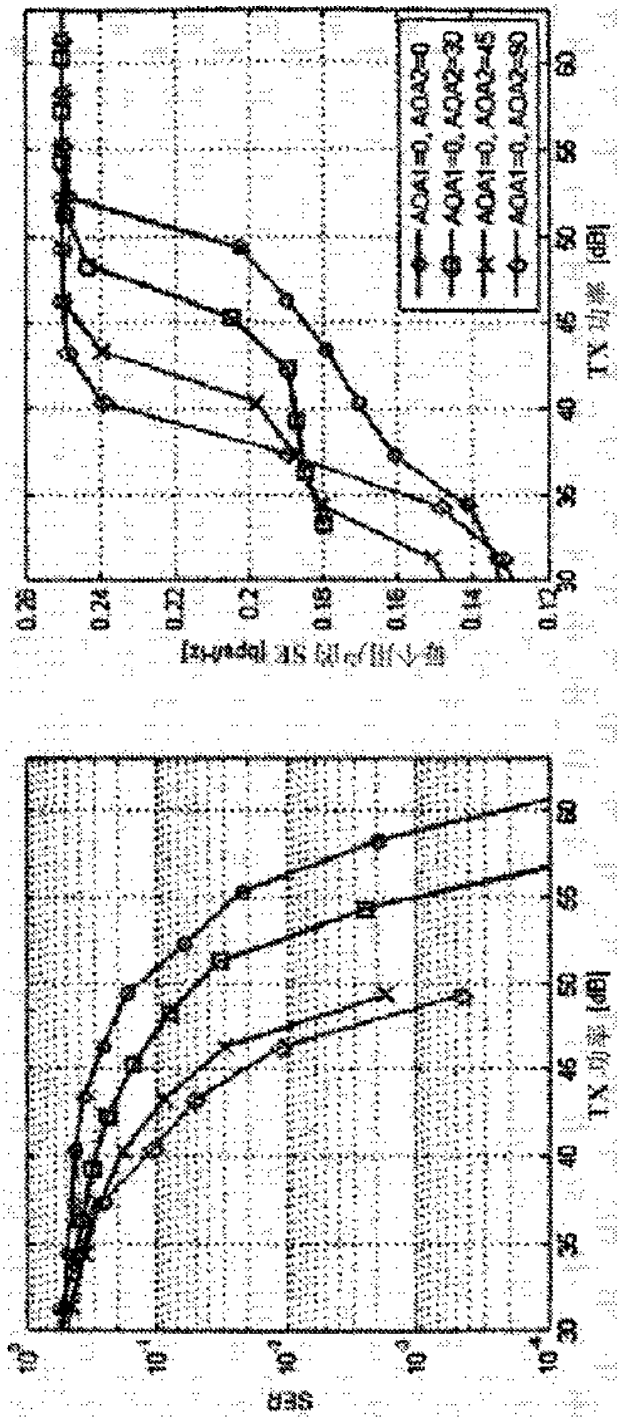

AOA1=0, AOA2=0
AOA1=0, AOA2=30
AOA1=0, AOA2=45
AOA1=0, AOA2=60
TX 功率 [dB]
每个用户的 SE [bps/Hz]
SER
TX 功率 [dB]

图26

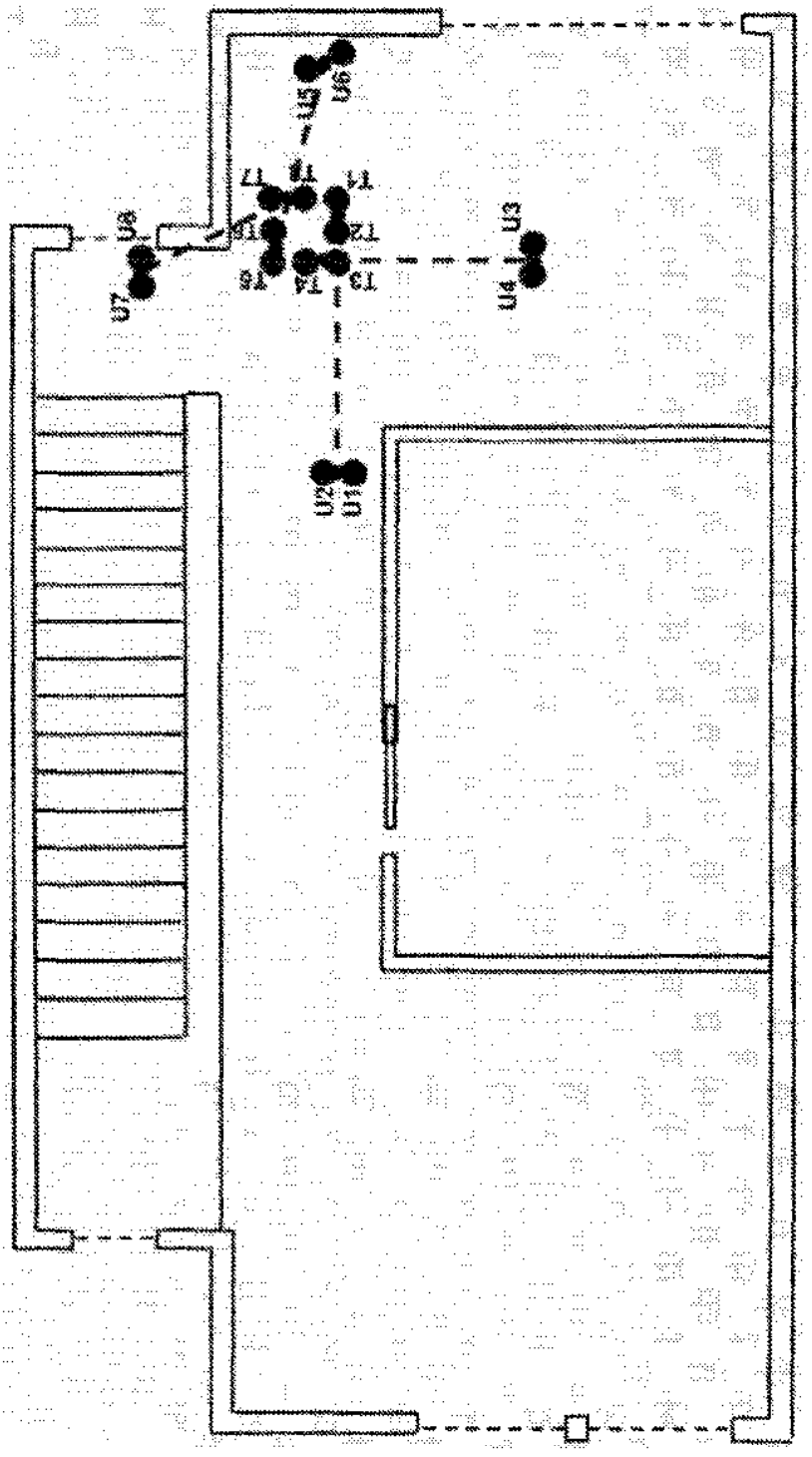


图27

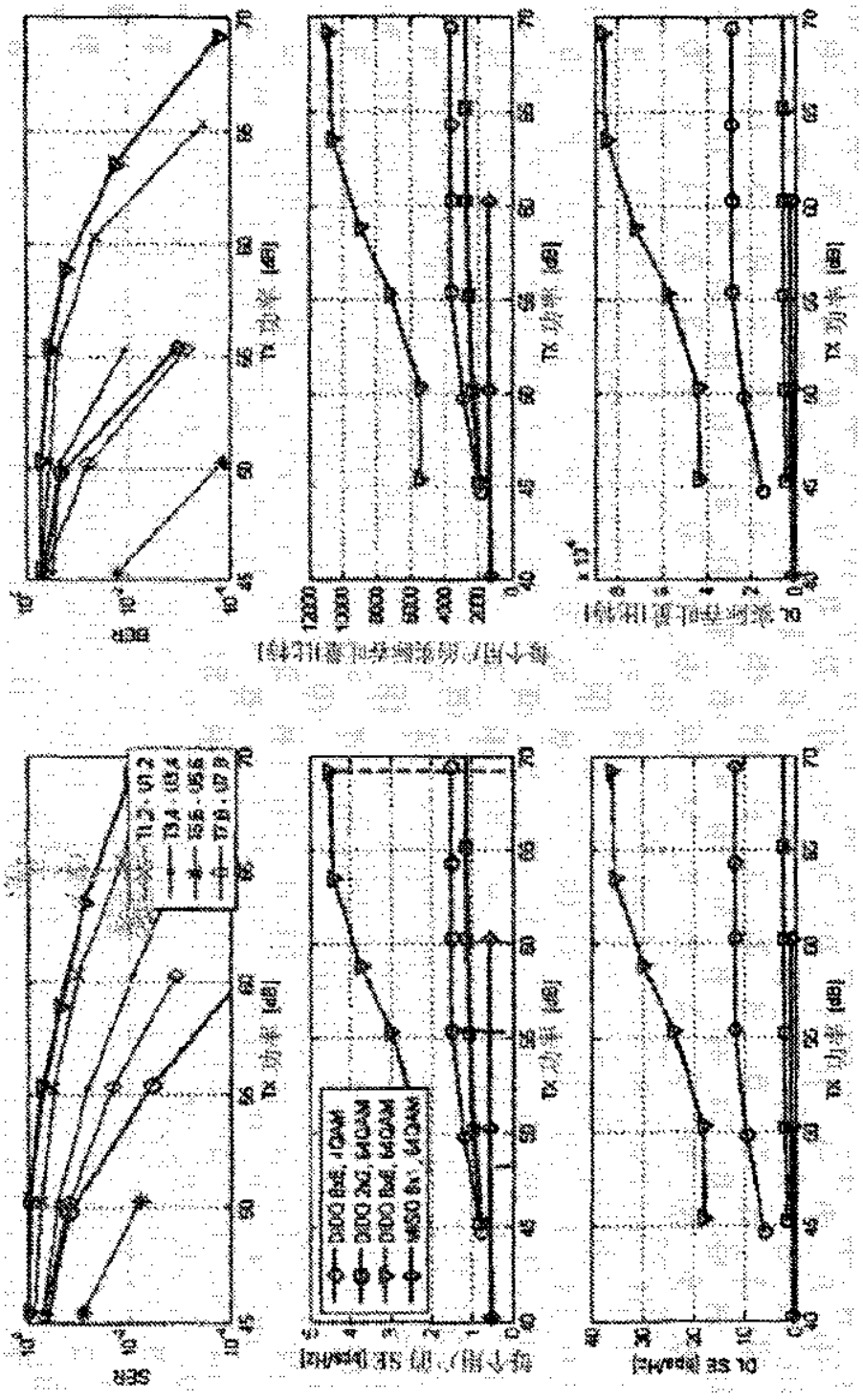

BER
SER
每个用户的 SE [bps/Hz]
每个用户的频谱效率比例
Tx 功率 [dB]
DSO BS, IGM
DSO NGM
DSO BS, CGD
NGM
WSO BS, IGM

图28

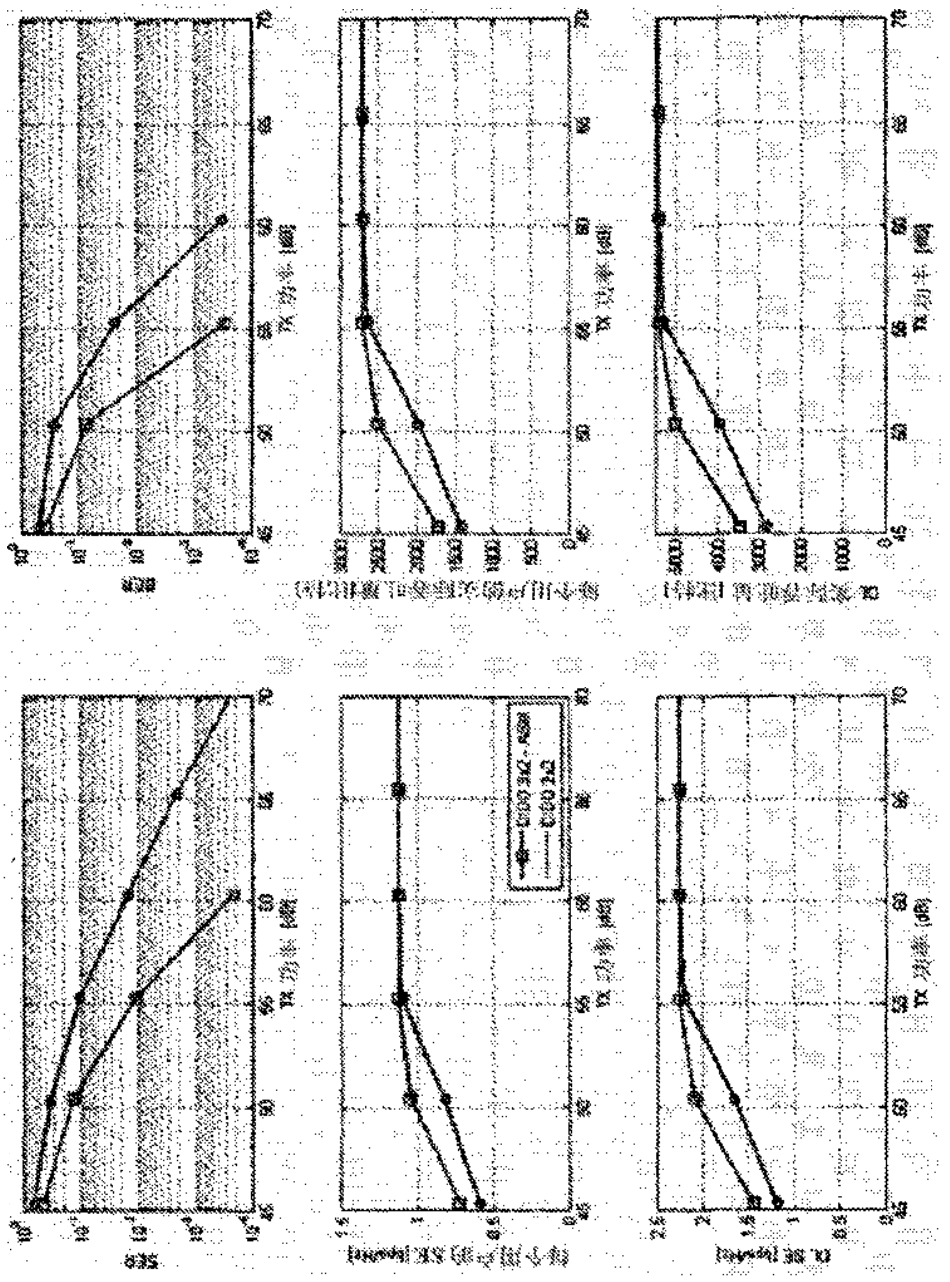


图29

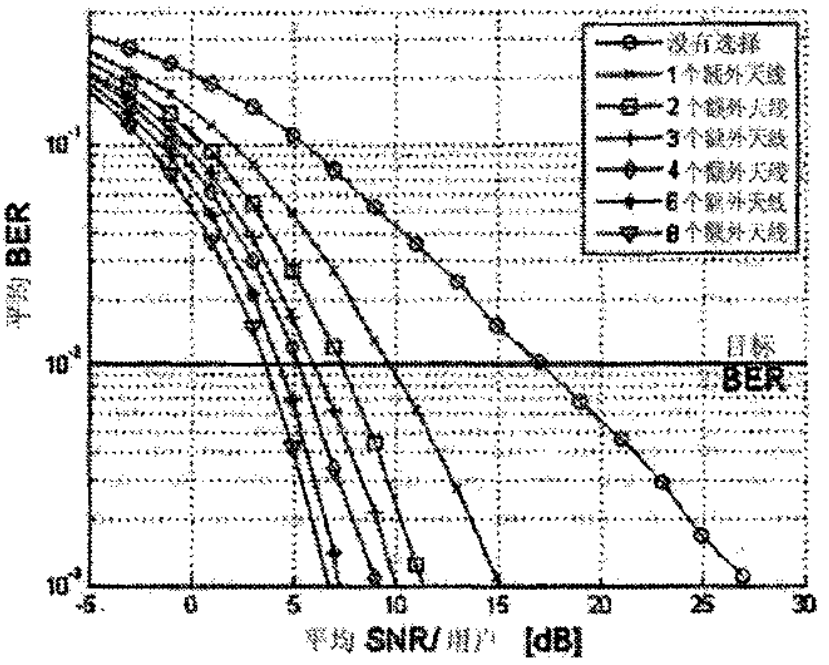

没有选择
1 个额外天线
2 个额外天线
3 个额外天线
4 个额外天线
6 个额外天线
8 个额外天线
目标
BER
平均 BER
平均 SNR/用户 [dB]

图30

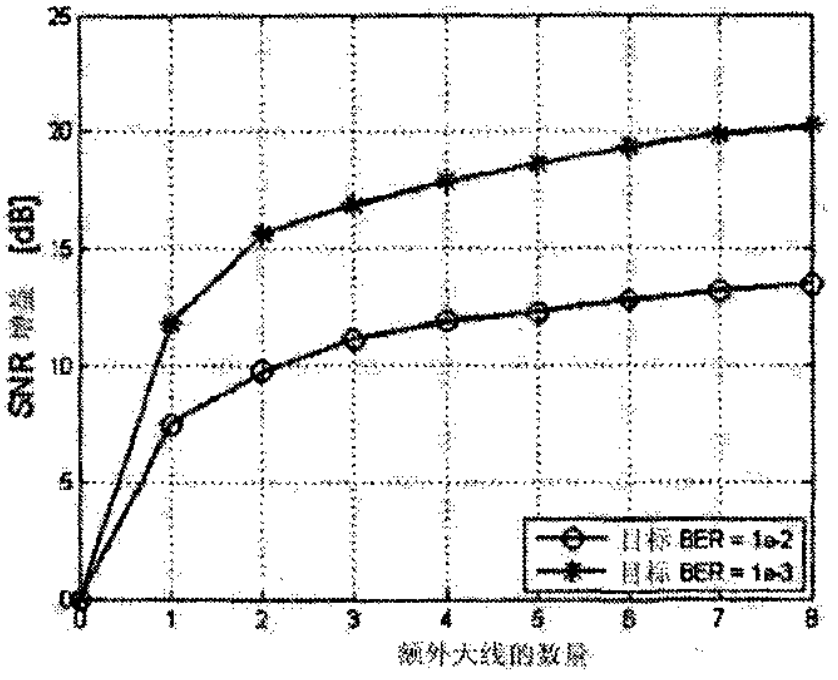

目标 BER = 1e-2
目标 BER = 1e-3
SNR 增益 [dB]
额外天线的数量

图31

图33

图32

说 明 书 附 图

CN 103501195 B

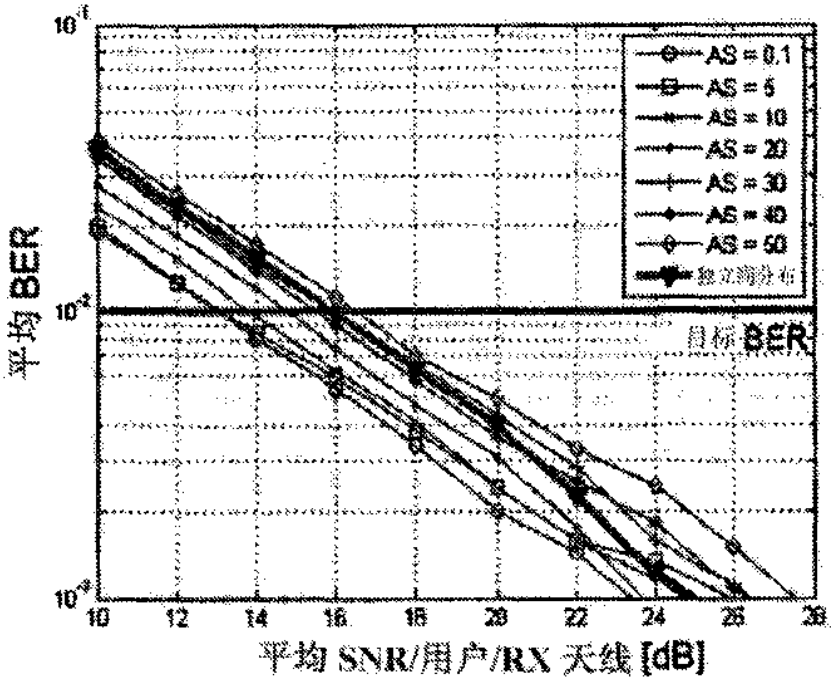

AS = 0.1
AS = 5
AS = 10
AS = 20
AS = 30
AS = 40
AS = 50
目标 BER
平均 BER
平均 SNR/用户/RX 天线 [dB]

图34

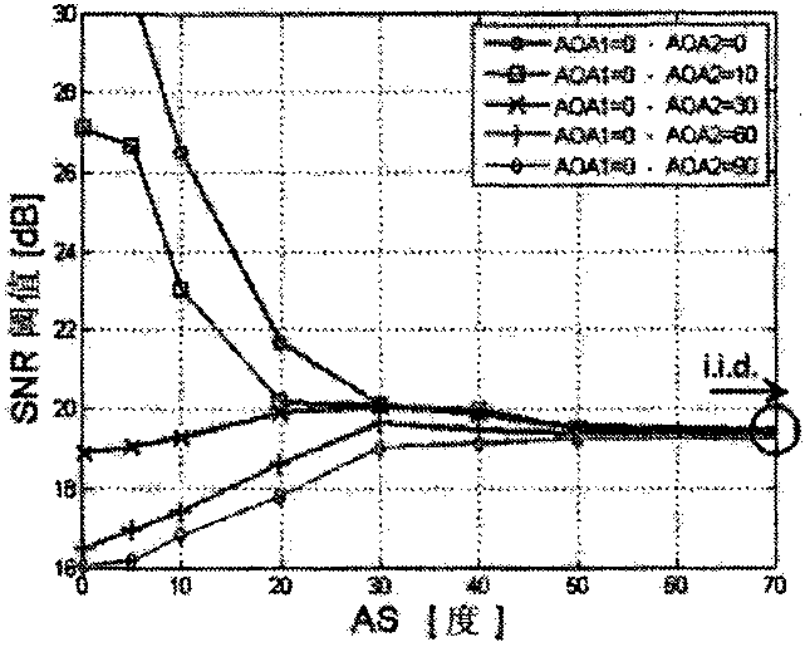

AOA1=0 · AOA2=0
AOA1=0 · AOA2=10
AOA1=0 · AOA2=30
AOA1=0 · AOA2=60
AOA1=0 · AOA2=90
i.i.d.
SNR 阈值 [dB]
AS [度]

图35

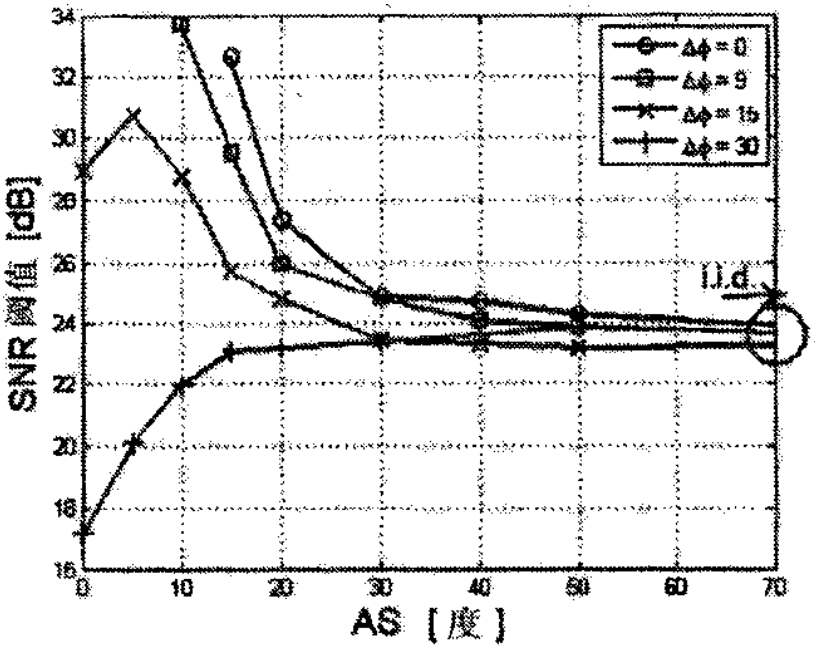

Δφ = 0
Δφ = 9
Δφ = 15
Δφ = 30
I.I.d.
SNR 阈值 [dB]
AS [度]

图36

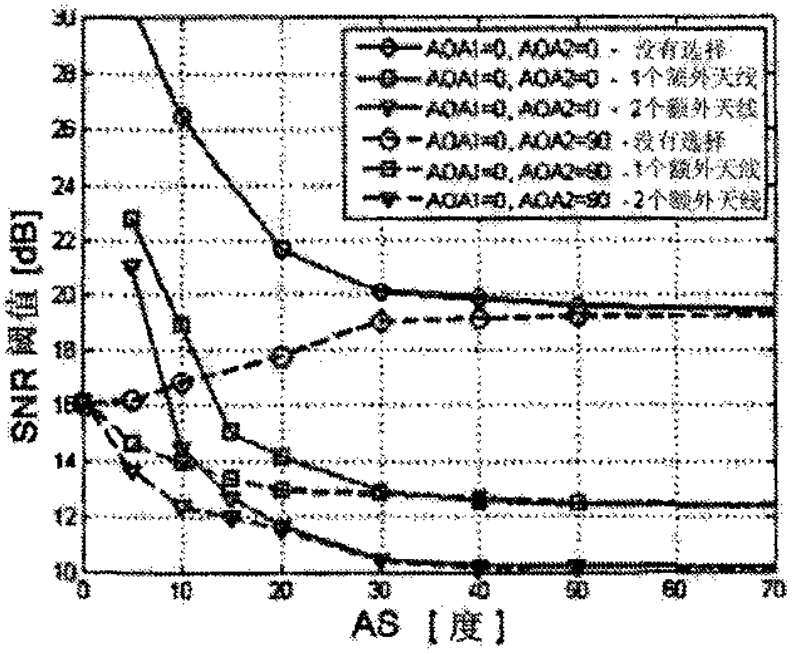

AOA1=0, AOA2=0 - 没有选择
AOA1=0, AOA2=0 - 1个额外天线
AOA1=0, AOA2=0 - 2个额外天线
AOA1=0, AOA2=90 - 没有选择
AOA1=0, AOA2=90 - 1个额外天线
AOA1=0, AOA2=90 - 2个额外天线
SNR 阈值 [dB]
AS [度]

图37

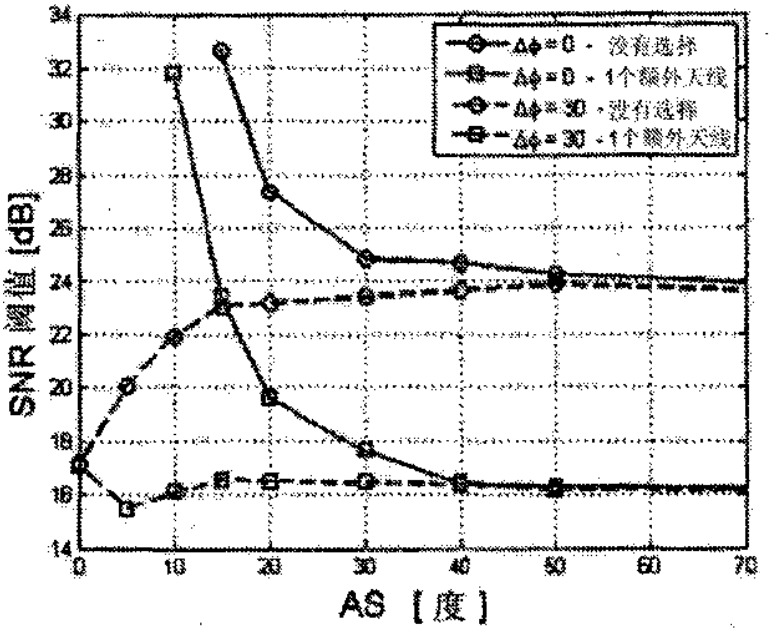

Δφ = 0 - 没有选择
Δφ = 0 - 1个额外天线
Δφ = 30 - 没有选择
Δφ = 30 - 1个额外天线
SNR 阈值 [dB]
AS [度]

图38

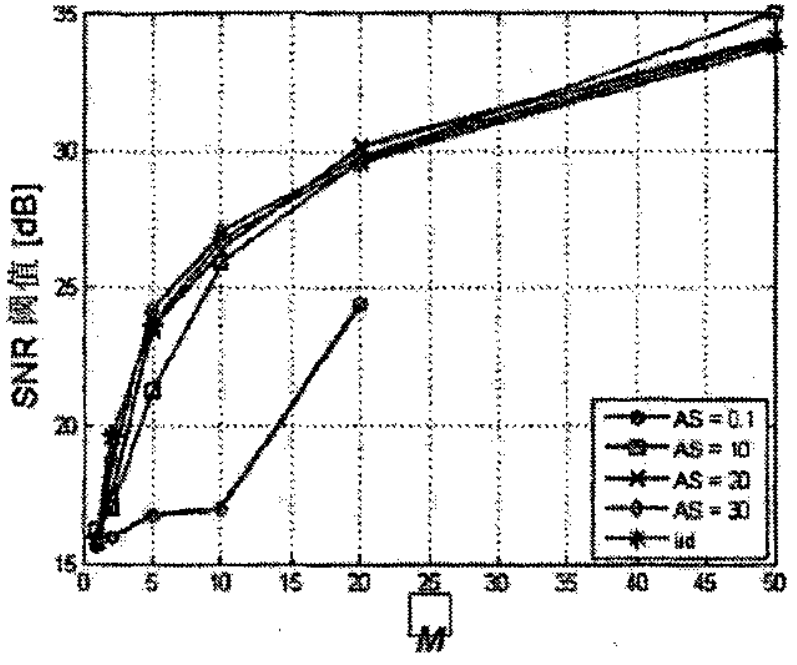

AS = 0.1
AS = 10
AS = 20
AS = 30
iid
SNR 阈值 [dB]
M

图39

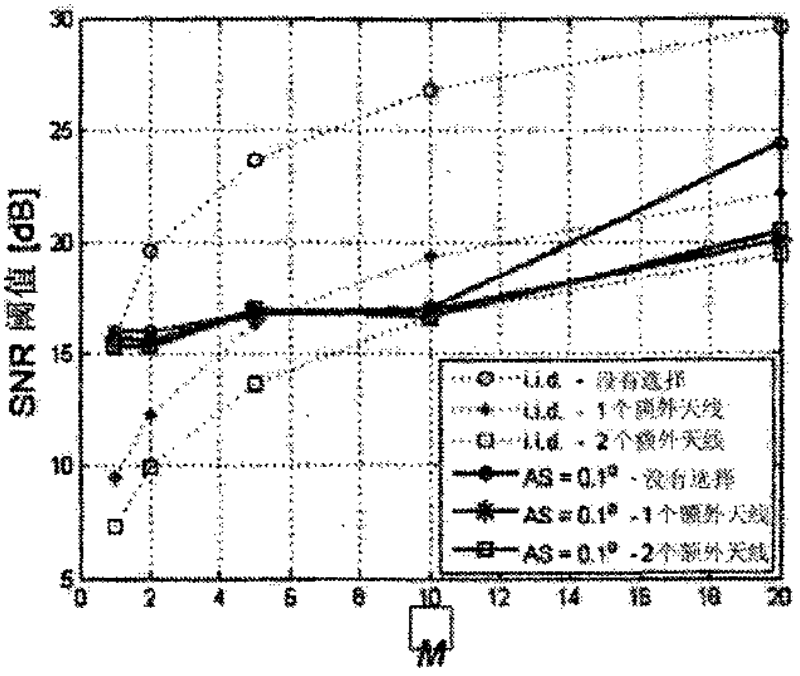

SNR 阈值 [dB]
M
i.i.d. - 没有选择
i.i.d. - 1个额外天线
i.i.d. - 2个额外天线
AS = 0.1° - 没有选择
AS = 0.1° -1个额外天线
AS = 0.1° -2个额外天线

图40

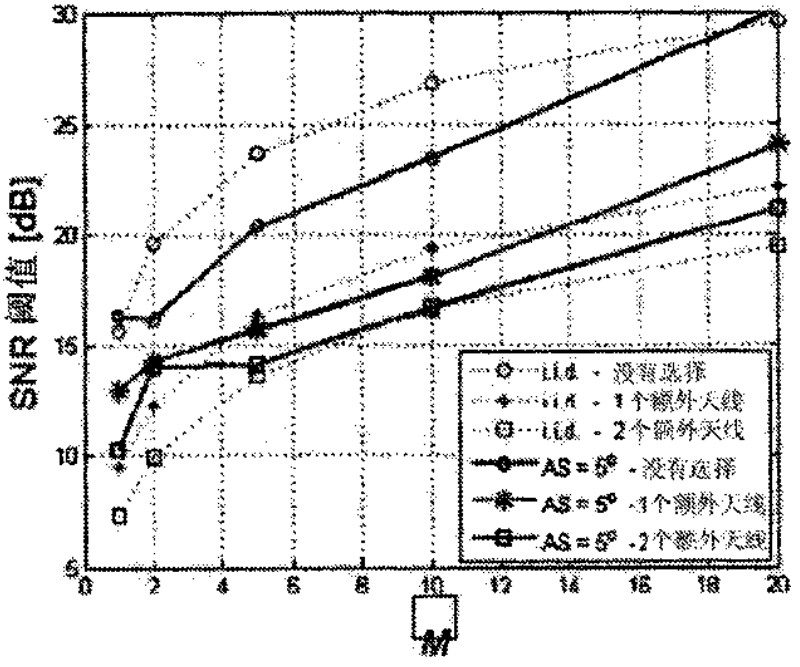

SNR 阈值 [dB]
M
i.i.d. - 没有选择
i.i.d. - 1个额外天线
i.i.d. - 2个额外天线
AS = 5° - 没有选择
AS = 5° -1个额外天线
AS = 5° -2个额外天线

图41

85

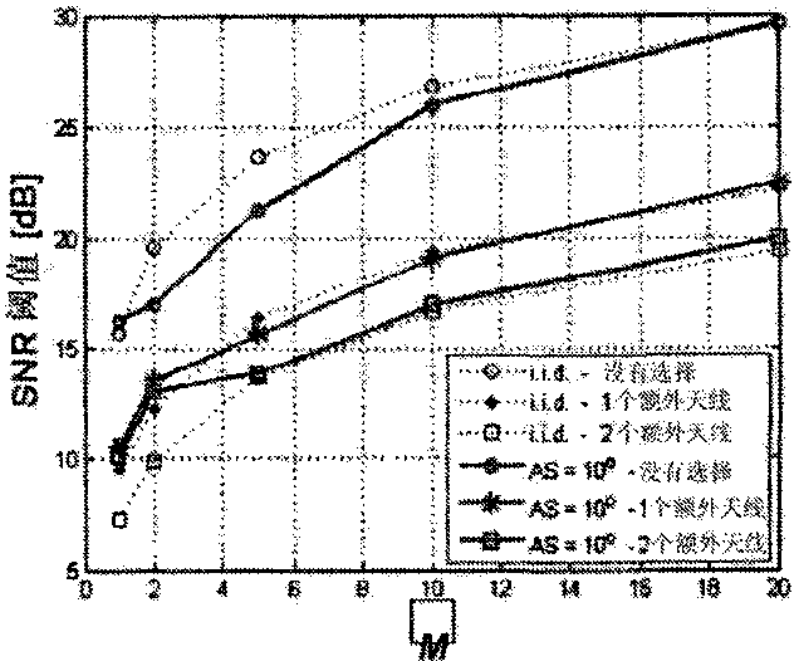

SNR 阈值 [dB]
M
i.i.d. - 没有选择
i.i.d. - 1个额外天线
i.i.d. - 2个额外天线
AS = 10° -没有选择
AS = 10° -1个额外天线
AS = 10° -2个额外天线

图42

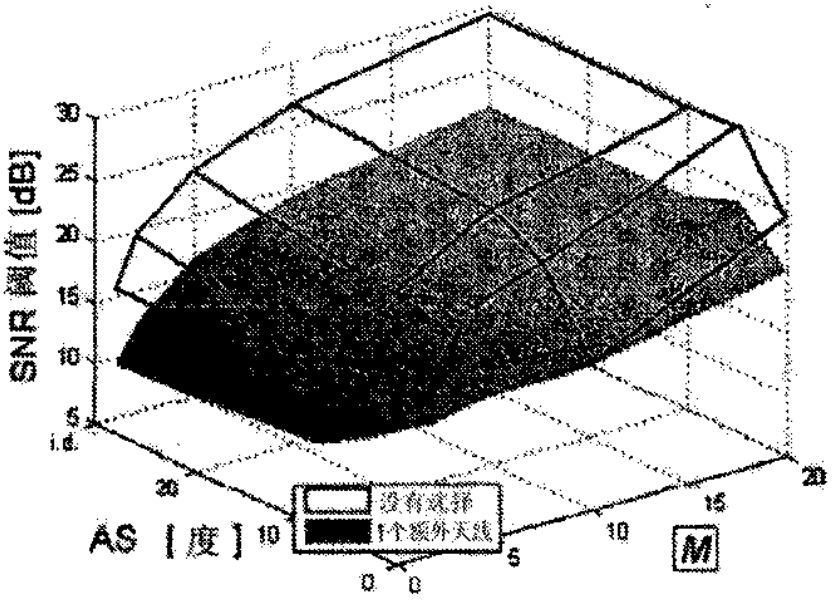

SNR 阈值 [dB]
AS [度]
M
没有选择
1个额外天线

图43

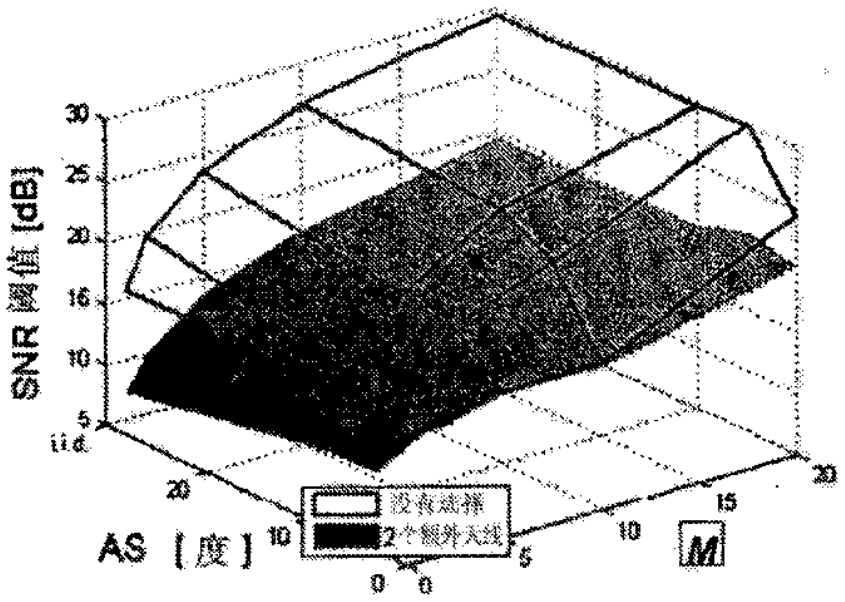

SNR 阈值 [dB]
30
25
20
15
10
5
AS [度]
20
10
0
没有选择
2个额外天线
0
5
10
15
20
M

图14

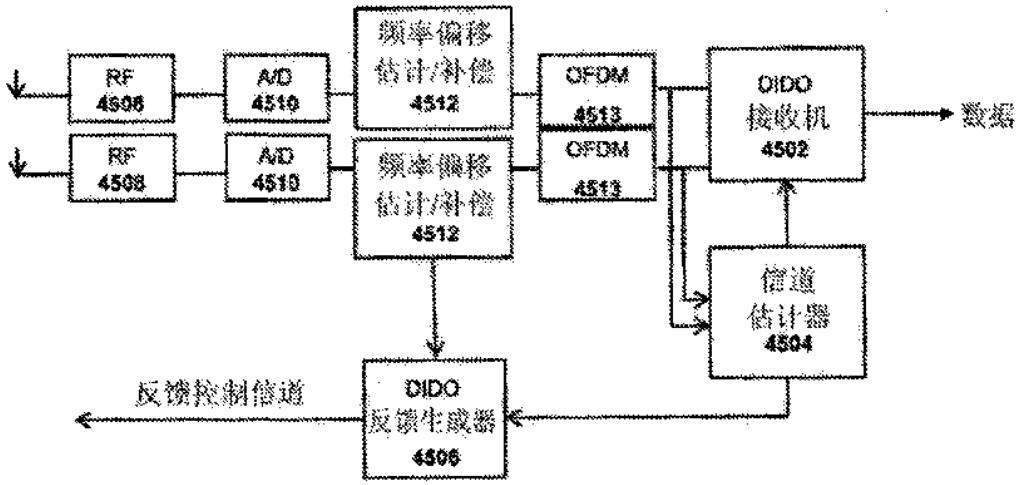

RF
4506
RF
4508
A/D
4510
A/D
4510
频率偏移
估计/补偿
4512
频率偏移
估计/补偿
4512
OFDM
4513
OFDM
4513
DIDO
接收机
4502
数据
信道
估计器
4504
DIDO
反馈生成器
4506
反馈控制信道

图15

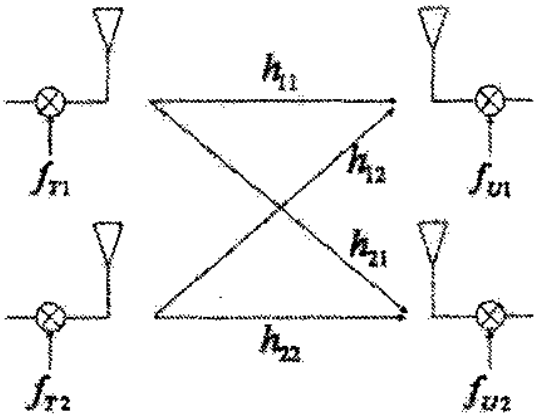

h_{11}
h_{12}
h_{21}
h_{22}
f_{T1}
f_{T2}
f_{U1}
f_{U2}

图46

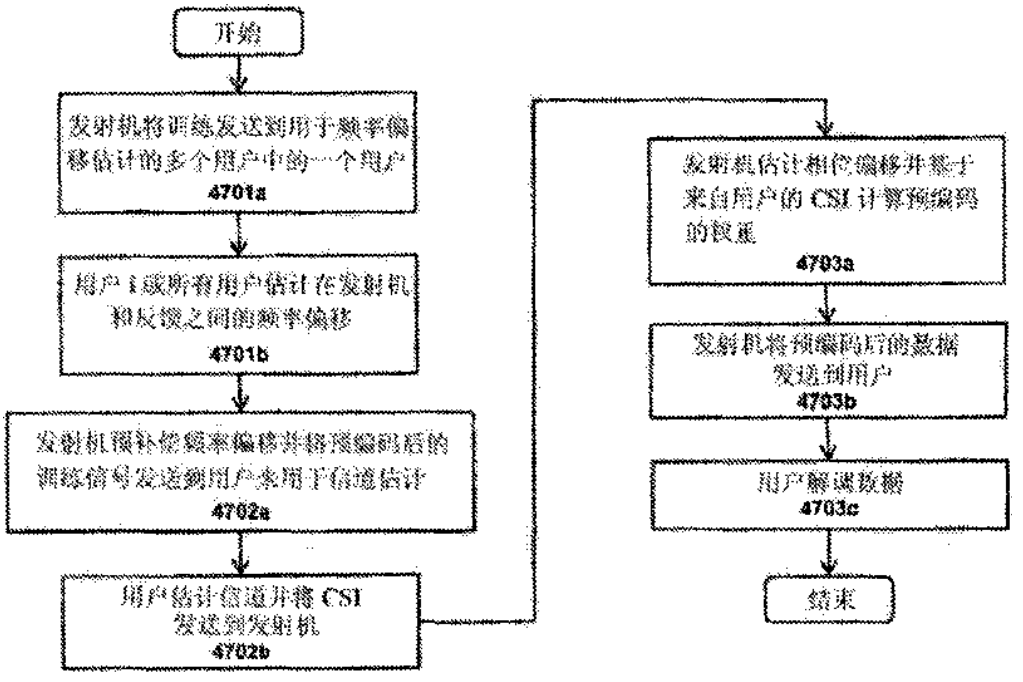

开始
发射机将训练发送到用于频率偏移估计的多个用户中的一个用户
4701a
用户 k 或所有用户估计在发射机和反馈之间的频率偏移
4701b
发射机预补偿频率偏移并将预编码后的训练信号发送到用户来用于信道估计
4702a
用户估计信道并将 CSI 发送到发射机
4702b
发射机估计相位偏移并基于来自用户的 CSI 计算预编码的权重
4703a
发射机将预编码后的数据发送到用户
4703b
用户解调数据
4703c
结束

图47

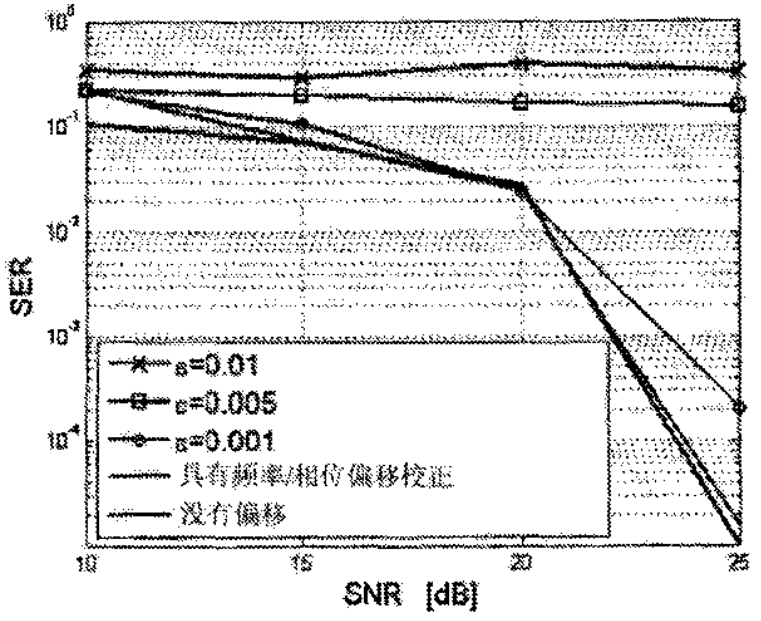

SER
SNR [dB]
ε=0.01
ε=0.005
ε=0.001
具有频率/相位偏移校正
没有偏移

图18

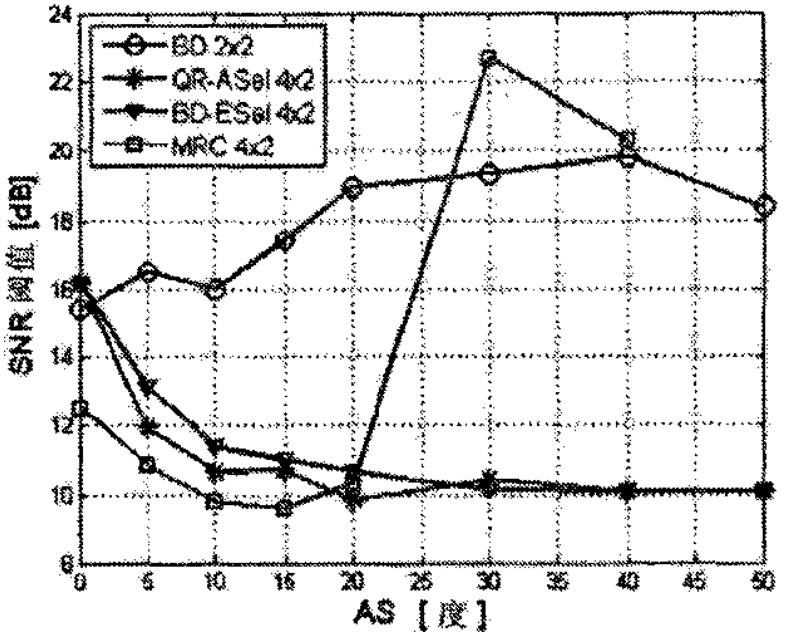

BD 2x2
QR-ASel 4x2
BD-ESel 4x2
MRC 4x2
SNR 阈值 [dB]
AS [度]

图19

方法	符号的数量	运行时间 (1,030 次迭代)	最大校正偏移
旧的	14208	600s	?
新的 ($N_t/\lvert M\rvert = 1$)	288	60s	1500Hz
新的 ($N_t/\lvert M\rvert = 4$ 长度 N_t)	1056	600s	1500Hz

图50

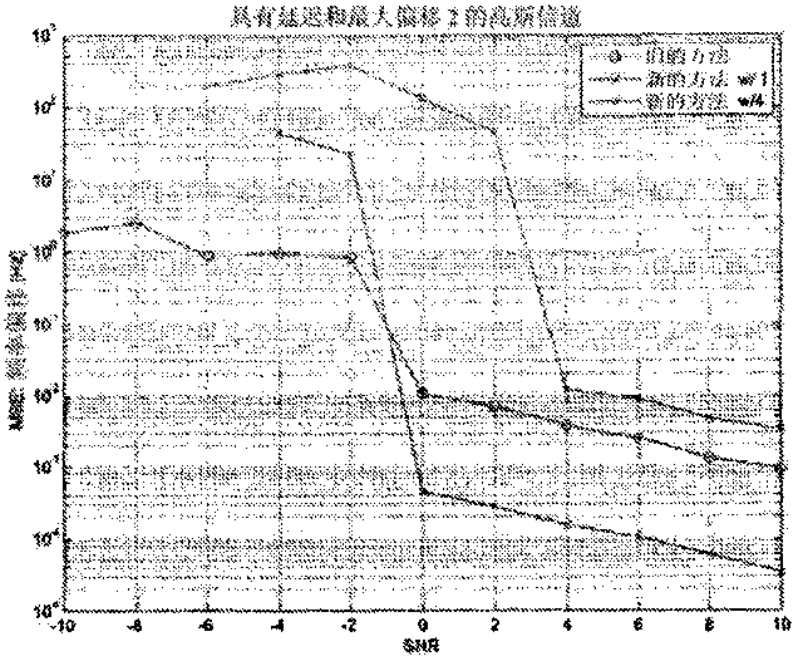

具有延迟和最大偏移 2 的瑞利信道
旧的方法
新的方法 w/1
新的方法 w/4
MSE: 频率偏移 (Hz)
SNR

图51

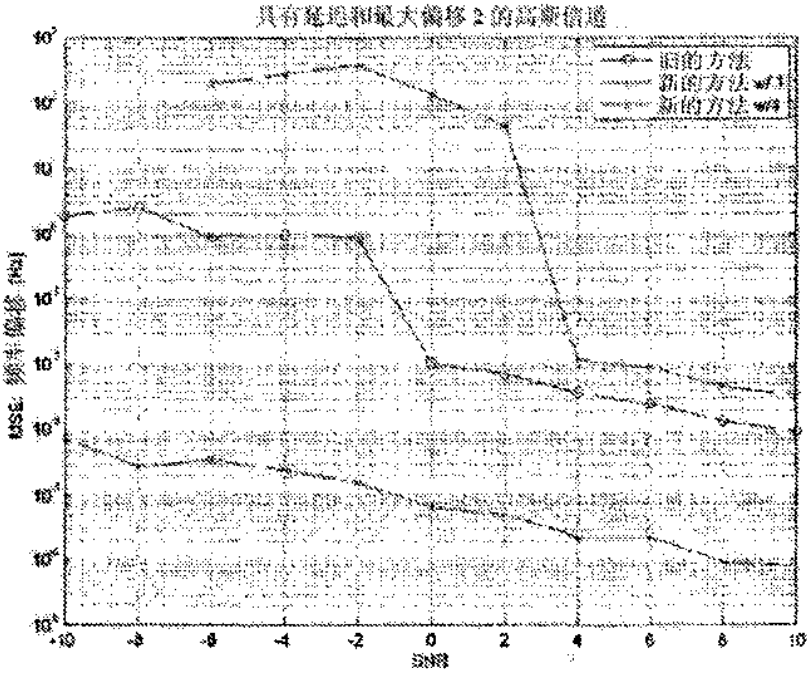

具有延迟和最大偏移 2 的高斯信道
旧的方法
新的方法 w/1
新的方法 w/4
MSE: 频率偏移 (Hz)
SNR

图52

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

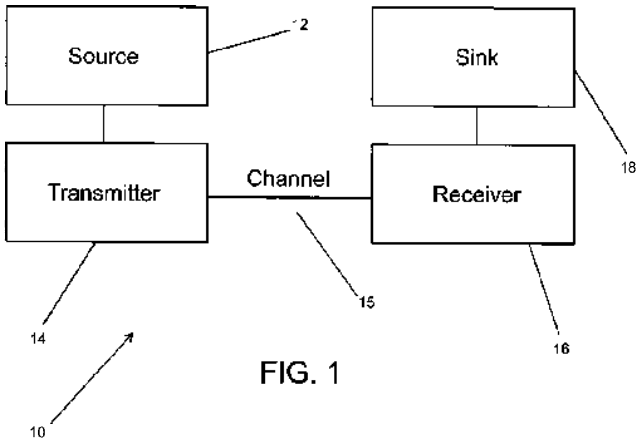


FIG. 1

(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_{mm}.

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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_{mm}) 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_{mm}, 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_{mm}. One approach to selecting a geometrically shaped constellation

having a greater capacity than that of a constellation maximized for d_{mn} 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, $_{i}V_{d\mathrm{Im}}$- 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:

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

where X_1 is the /th bit of the /-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:

$$CjOlNT = I(X\text{-}X)$$

[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 $C_0 \geq C_{J0INT} \geq C_{PD}$. 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 *apriori*. 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_{0Ut}) is determined (57). A determination (58) is made as to whether the SNR_{0Ut} and $SNR_{,n}$ have converged. In the illustrated embodiment convergence is indicated by SNR_{0Ut} equaling $SNR_{,n}$. In a number of embodiments, convergence can be determined based upon the difference between SNR_{0Ut} and $SNR_{,n}$ being less than a predetermined threshold. When SNR_{0Ut} and SNR_{In} have not converged, the process performs another iteration selecting SNR_{0ut} as the new SNR_{In} (55). When SNR_{0Ut} 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_{mn}.

[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_{mn} 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 = 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=l/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 capacaity 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 aper-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, $E[X^2]E[N^2]$. 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.

[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_{mn};

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_{mm} 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 geomtertically shaped symbol constellation; and

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

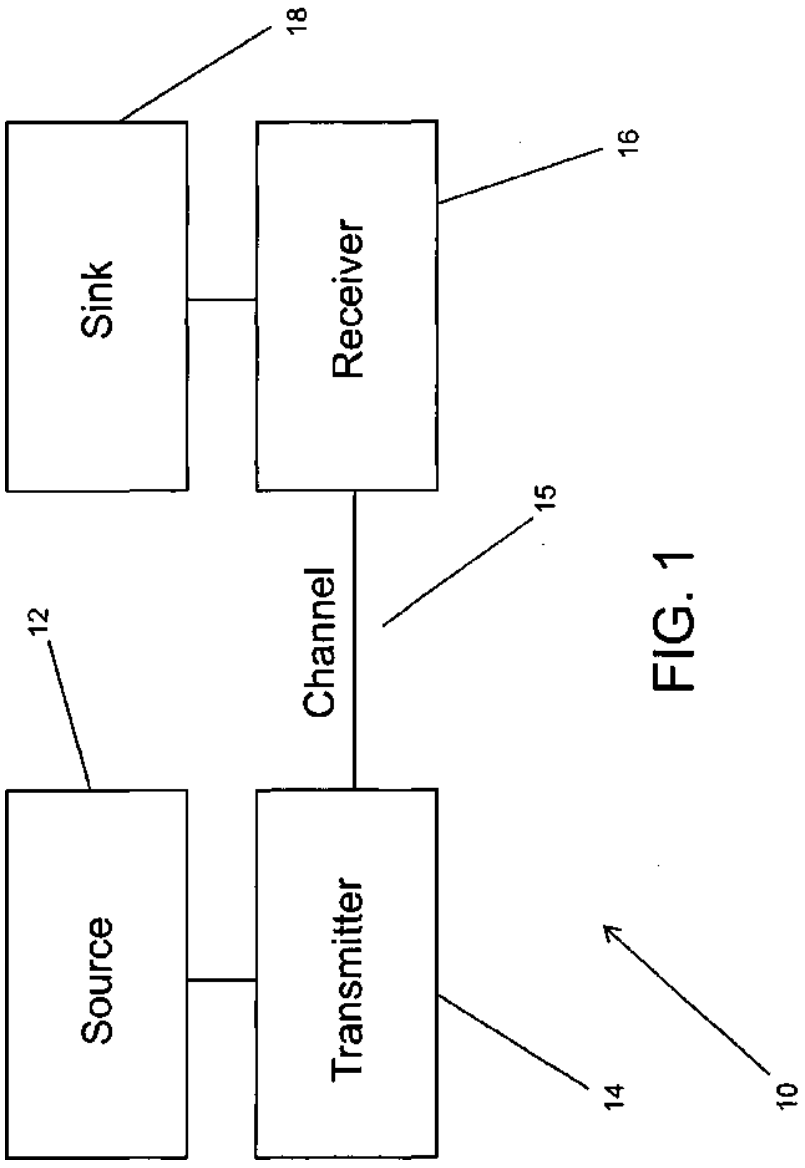


FIG. 1

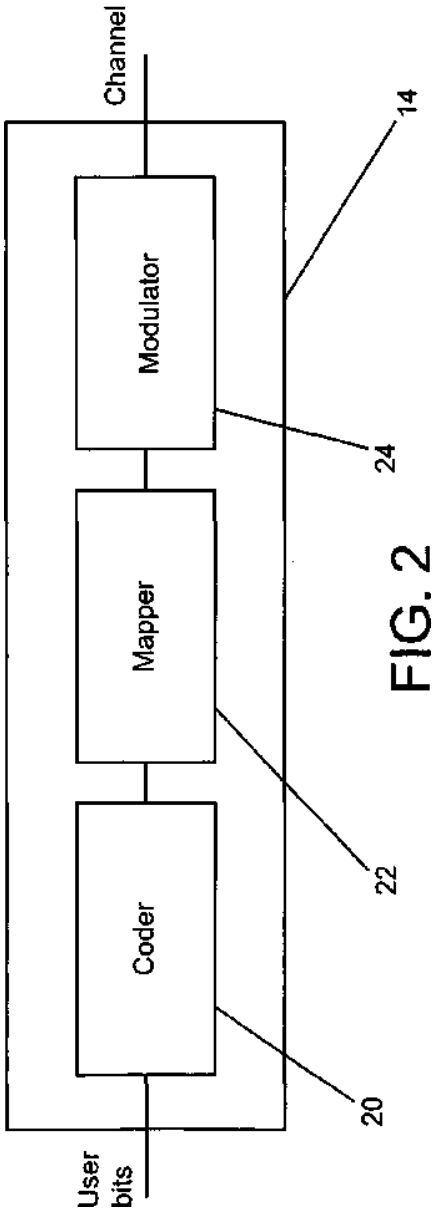

Channel
Modulator
Mapper
Coder
User bits
14
24
22
20

FIG. 2

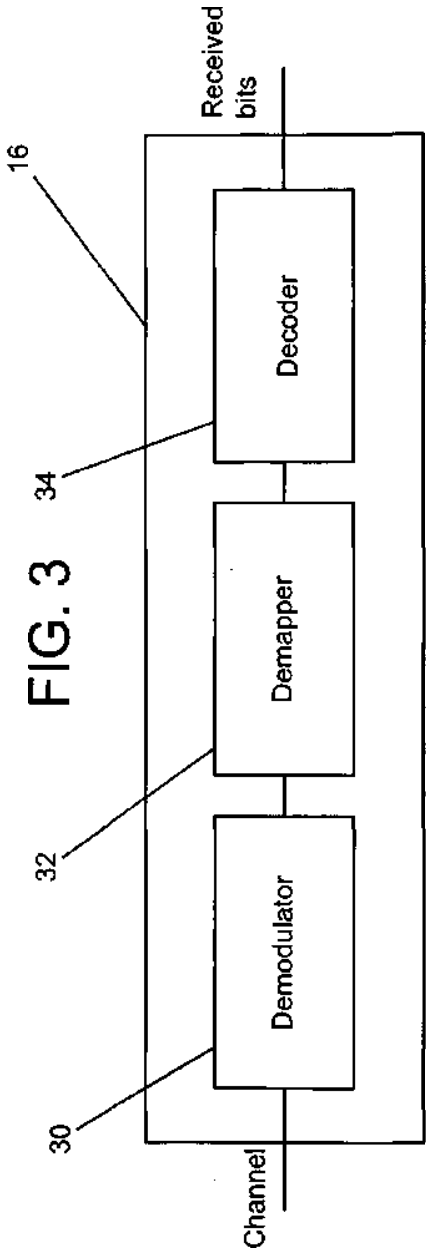

Received bits
Decoder
Demapper
Demodulator
Channel
16
34
32
30

FIG. 3

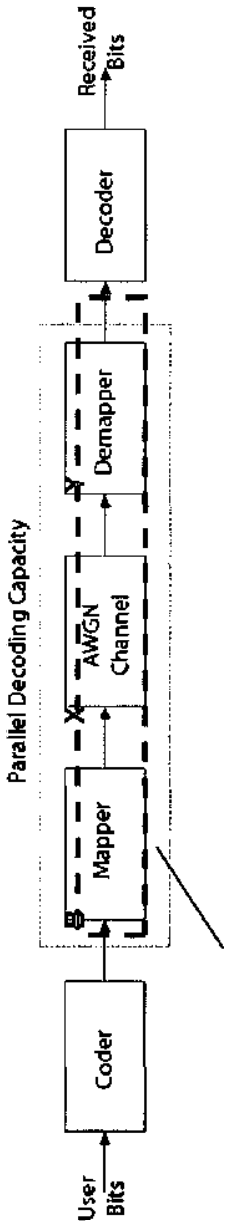

User Bits
Coder
B
Mapper
X
AWGN Channel
Y
Demapper
Decoder
Received Bits
Parallel Decoding Capacity
40

FIG. 4a

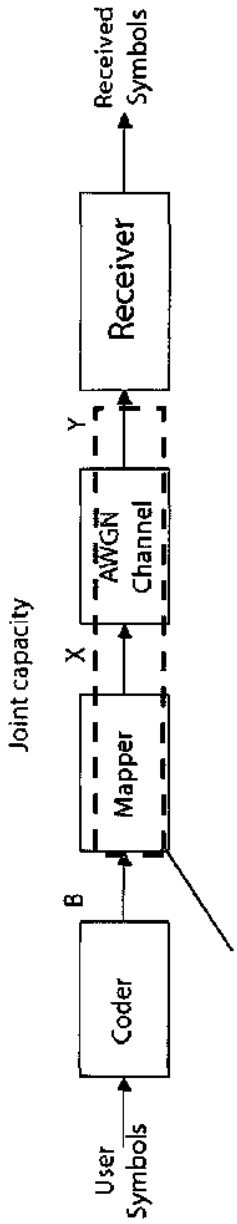

User Symbols
Coder
B
Mapper
X
AWGN Channel
Y
Receiver
Received Symbols
Joint capacity
42

FIG. 4b

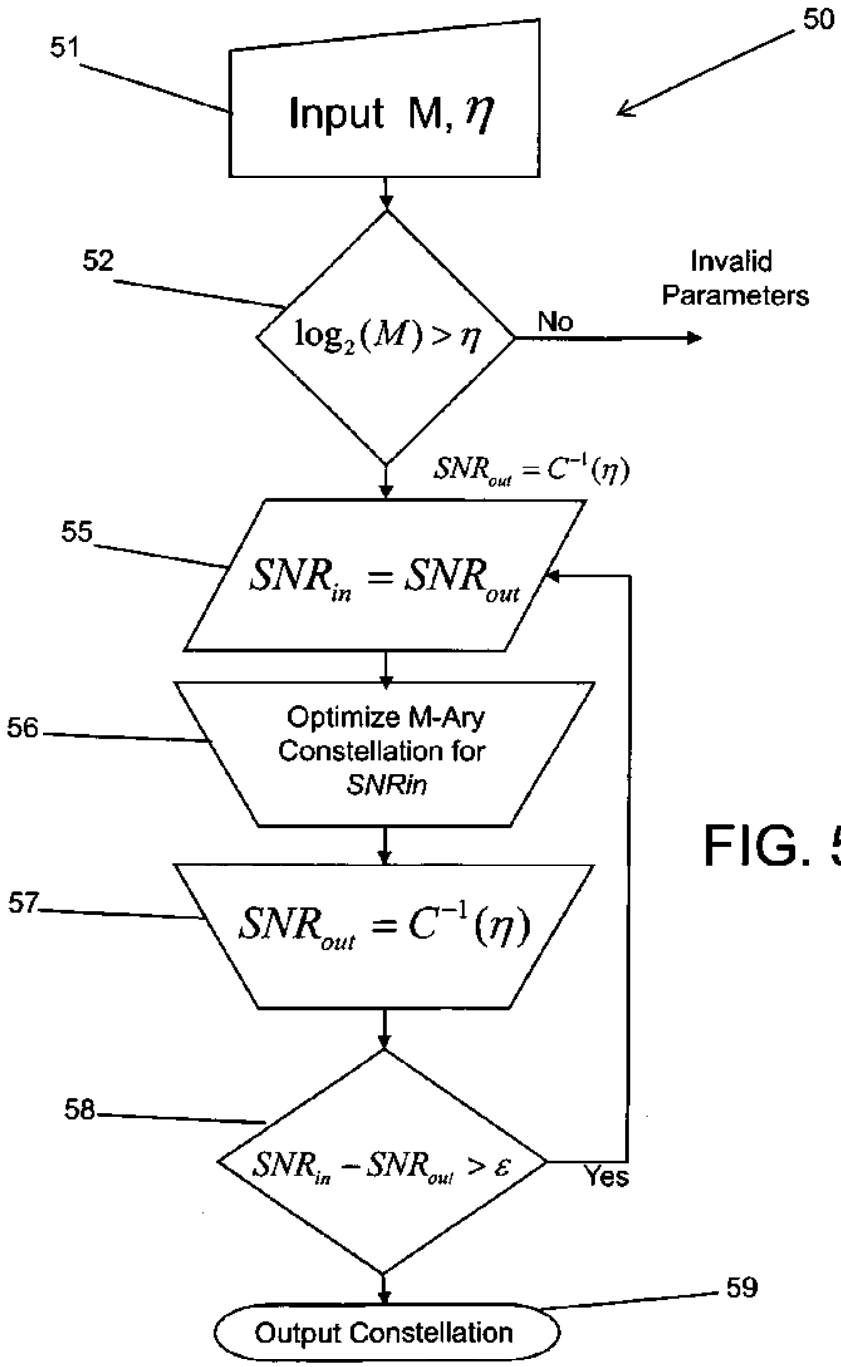

50
51
Input M, η
52
$\log_2(M) > \eta$
No
Invalid Parameters
$SNR_{out} = C^{-1}(\eta)$
55
$SNR_{in} = SNR_{out}$
56
Optimize M-Ary Constellation for SNRin
57
$SNR_{out} = C^{-1}(\eta)$
58
$SNR_{in} - SNR_{out} > \varepsilon$
Yes
59
Output Constellation

FIG. 5

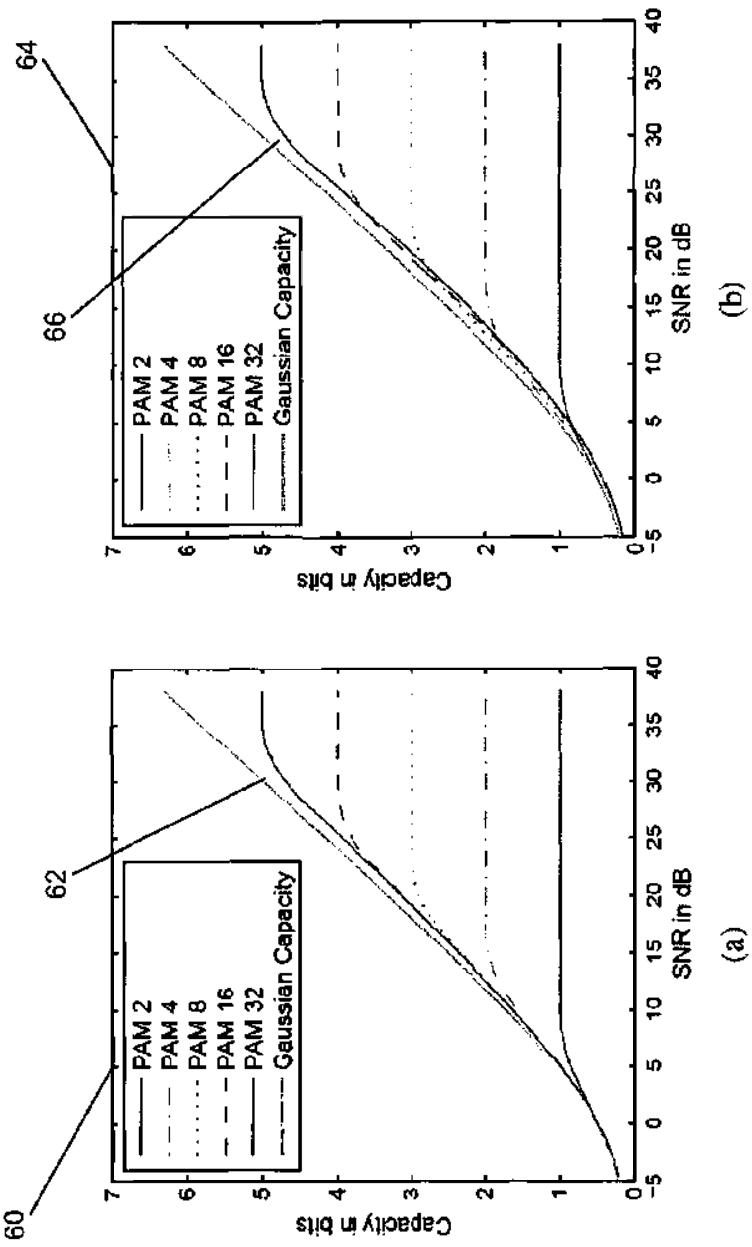

60
62
64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(a)
(b)

FIG. 6a

FIG. 6b

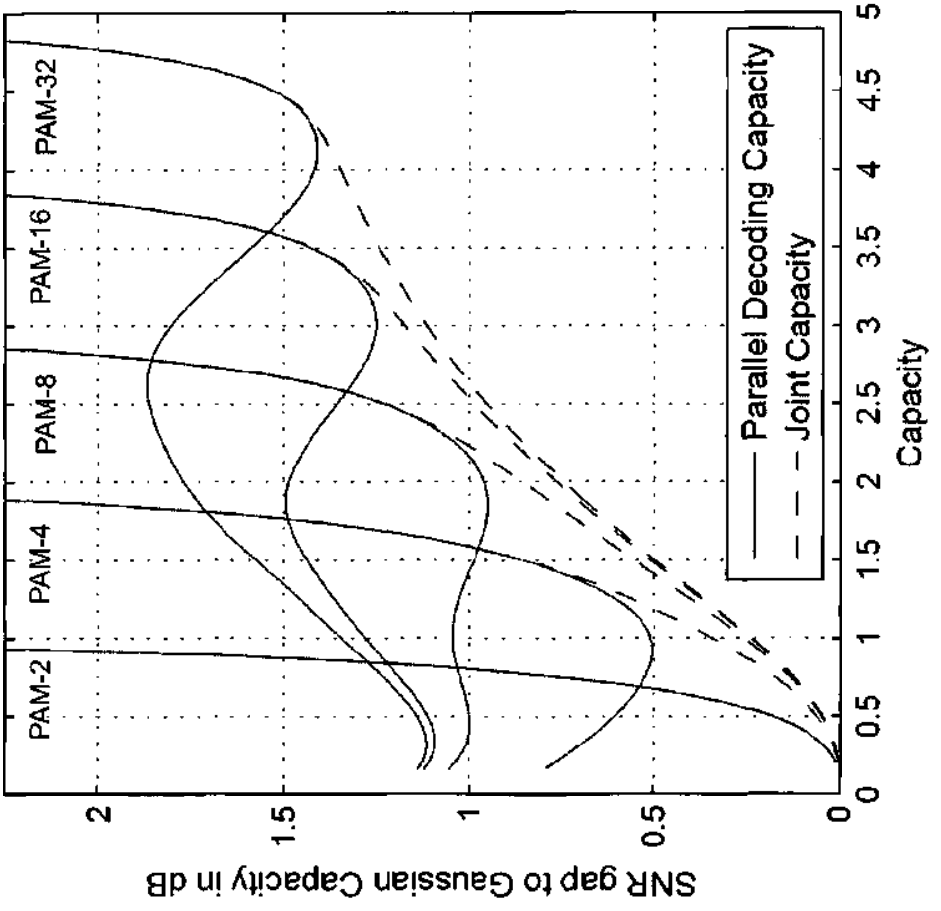

SNR gap to Gaussian Capacity in dB
Capacity
Parallel Decoding Capacity
Joint Capacity
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 7

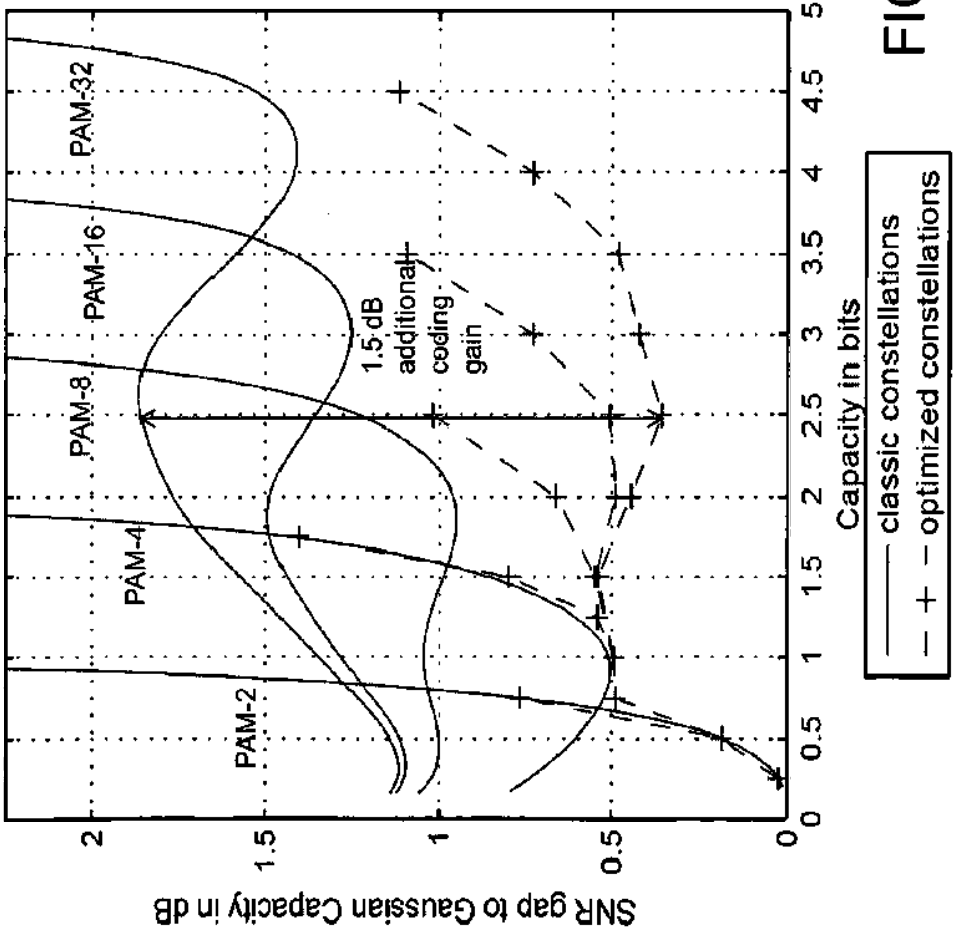

SNR gap to Gaussian Capacity in dB
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
1.5 dB additional coding gain
classic constellations
optimized constellations

FIG. 8a

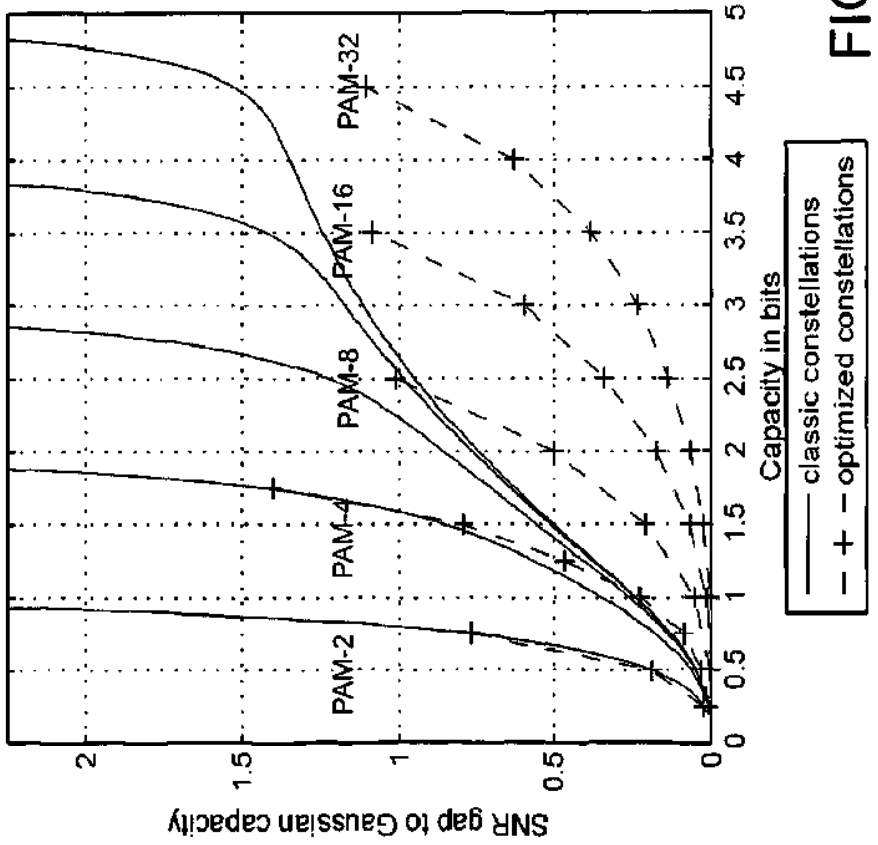

SNR gap to Gaussian capacity
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
classic constellations
optimized constellations
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8b

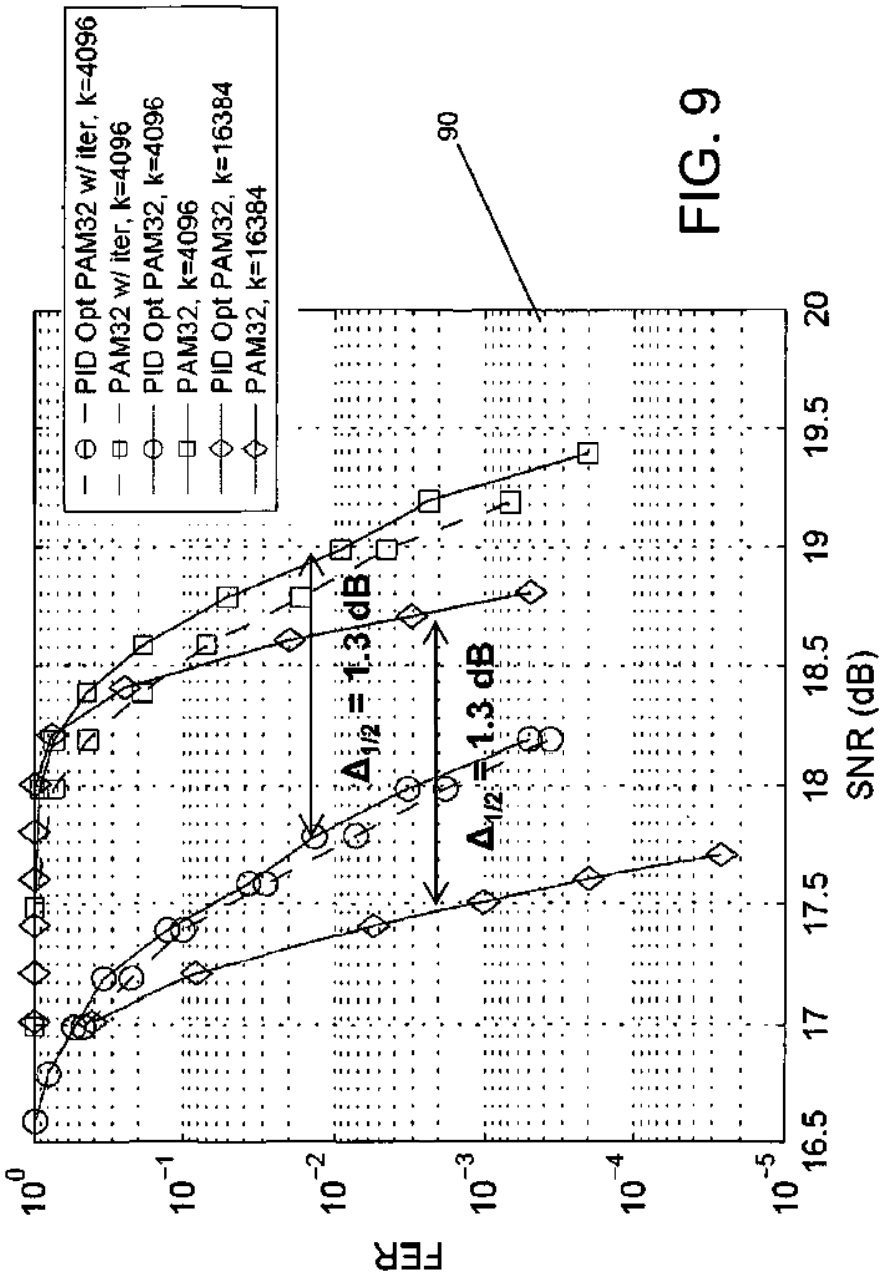

PID Opt PAM32 w/ iter, k=4096
PAM32 w/ iter, k=4096
PID Opt PAM32, k=4096
PAM32, k=4096
PID Opt PAM32, k=16384
PAM32, k=16384
90
$\Delta_{1/2}$ = 1.3 dB
$\Delta_{1/2}$ = 1.3 dB
FER
SNR (dB)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20

FIG. 9

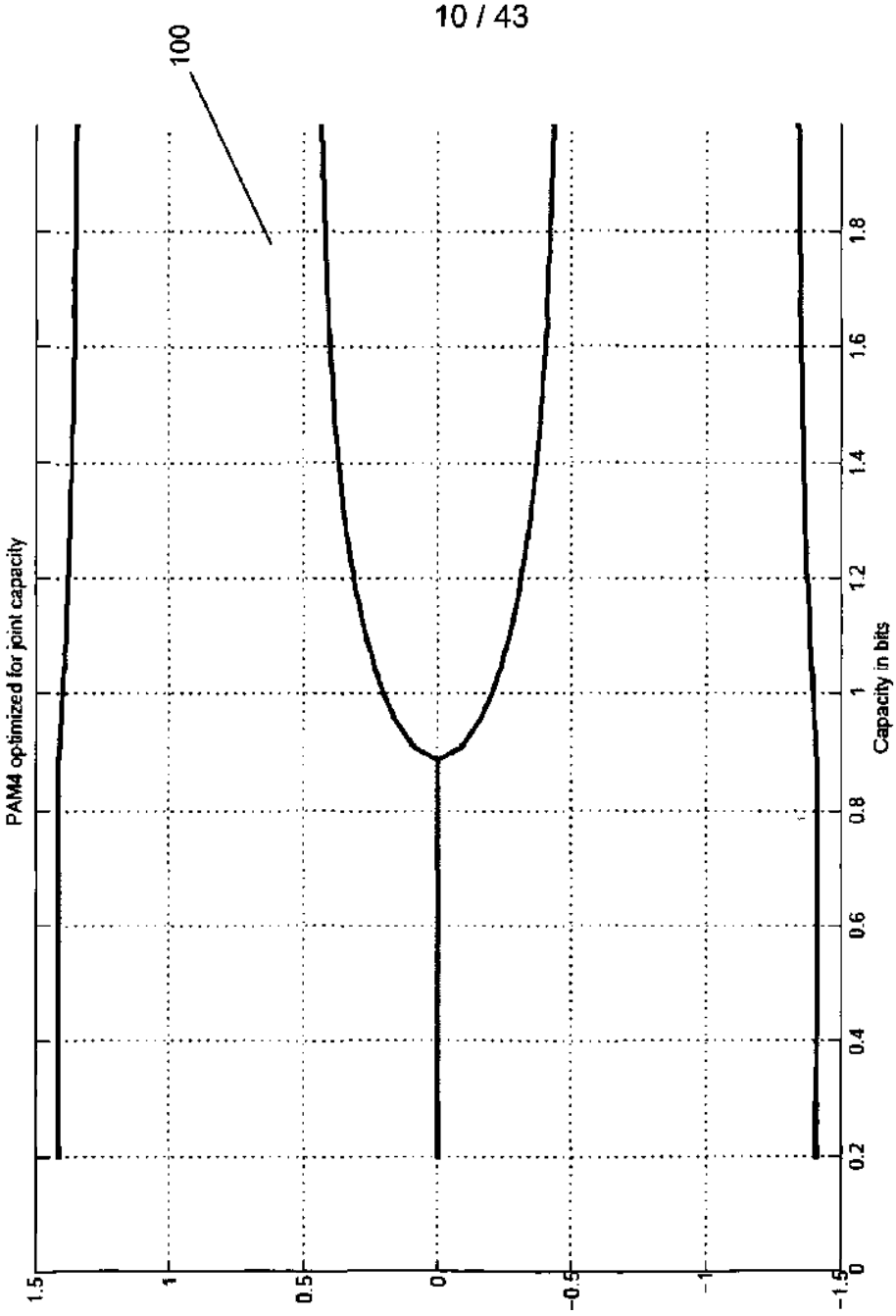

100
PAM4 optimized for joint capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10a

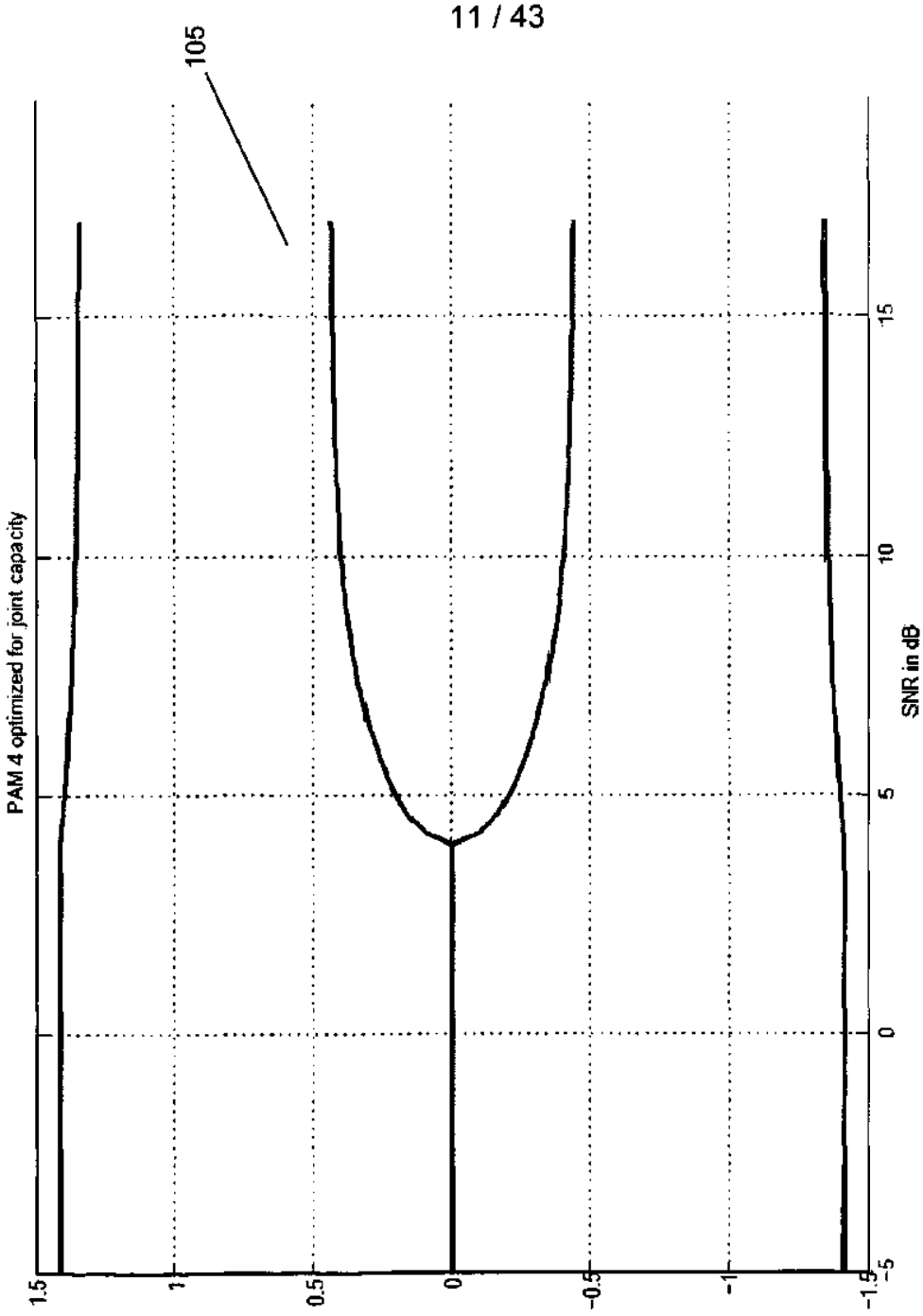

105
PAM 4 optimized for joint capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10b

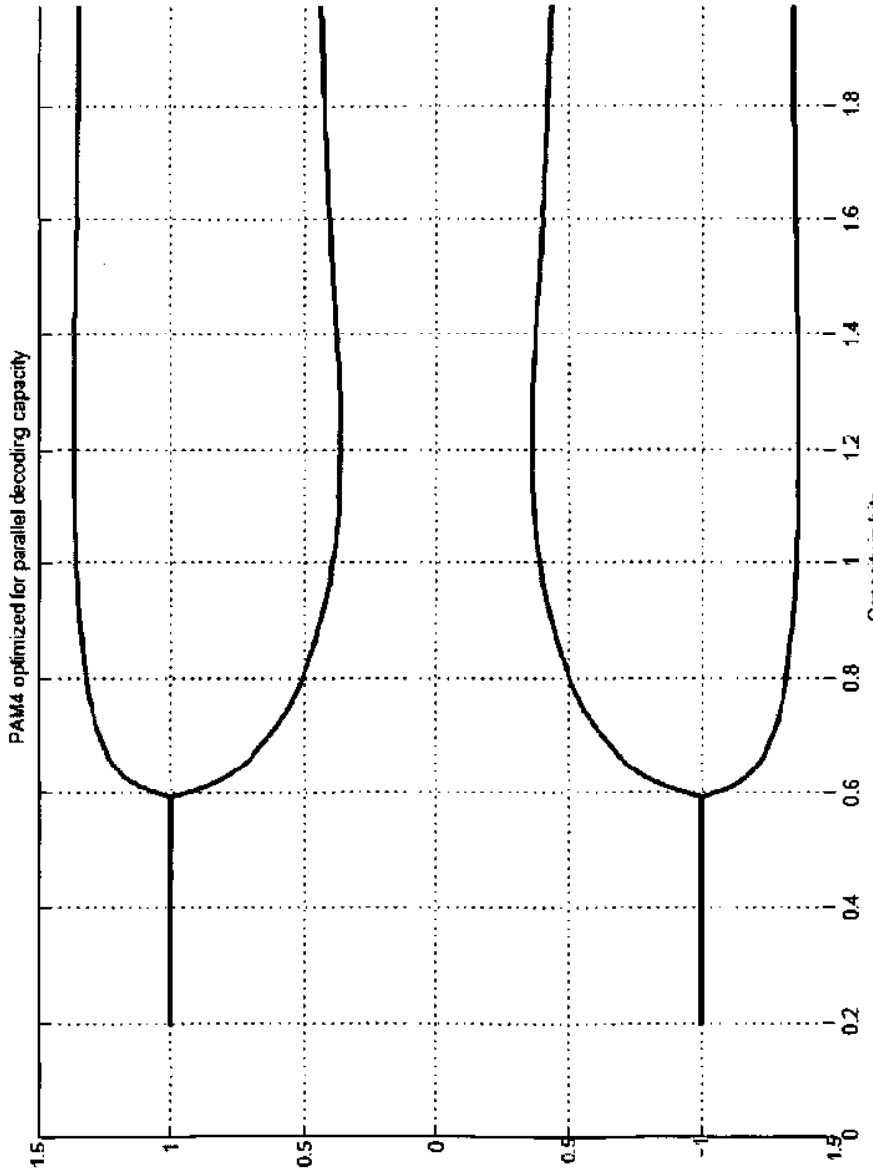

PAM4 optimized for parallel decoding capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10c

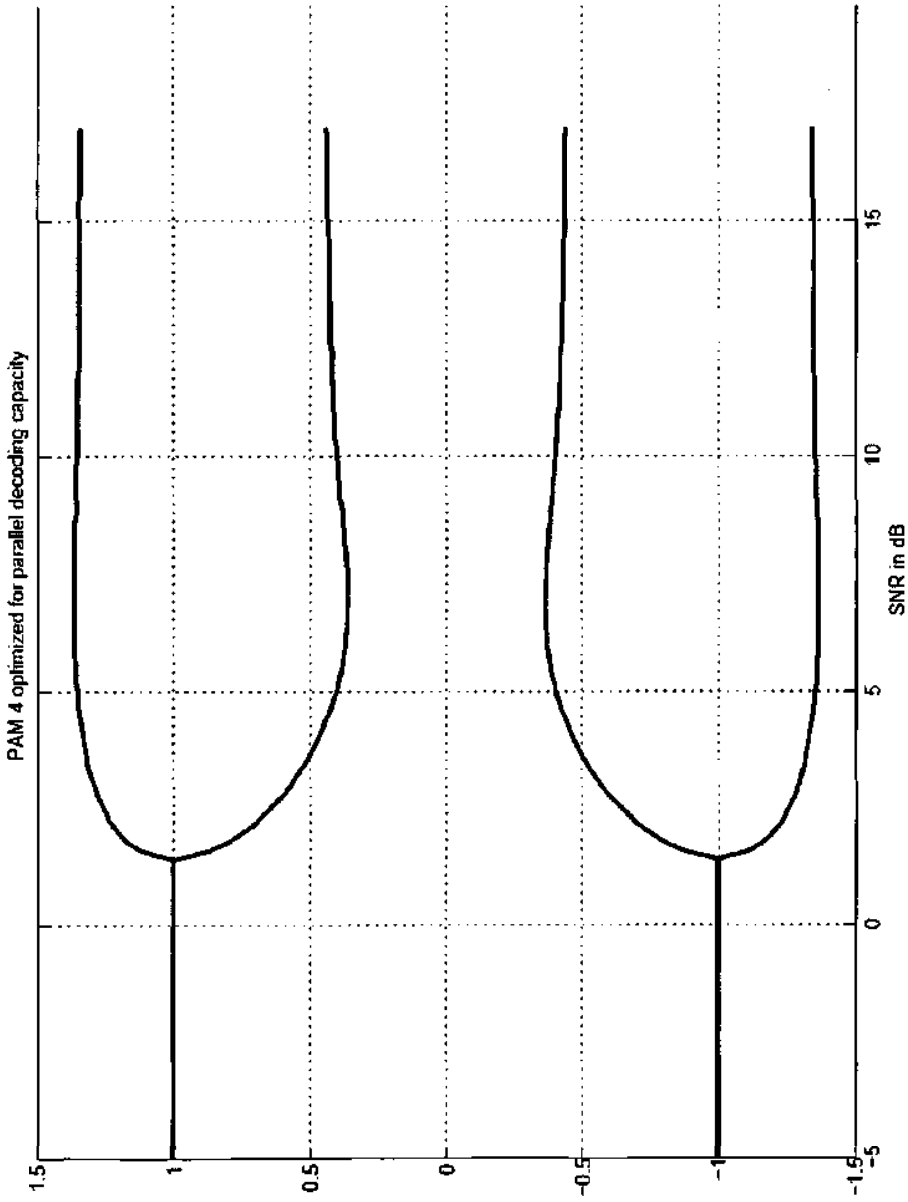

PAM 4 optimized for parallel decoding capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
15
10
5
0
-5

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

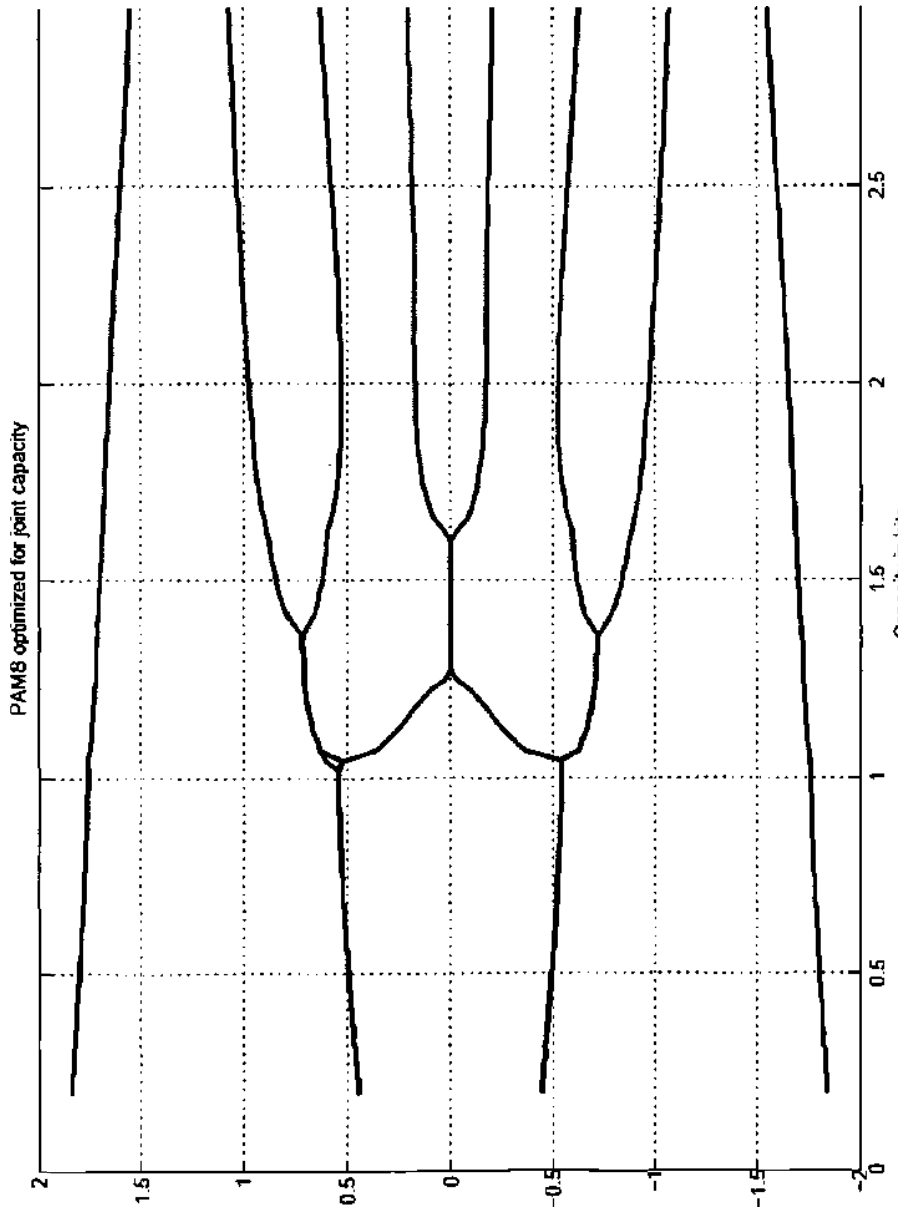

PAM8 optimized for joint capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12a

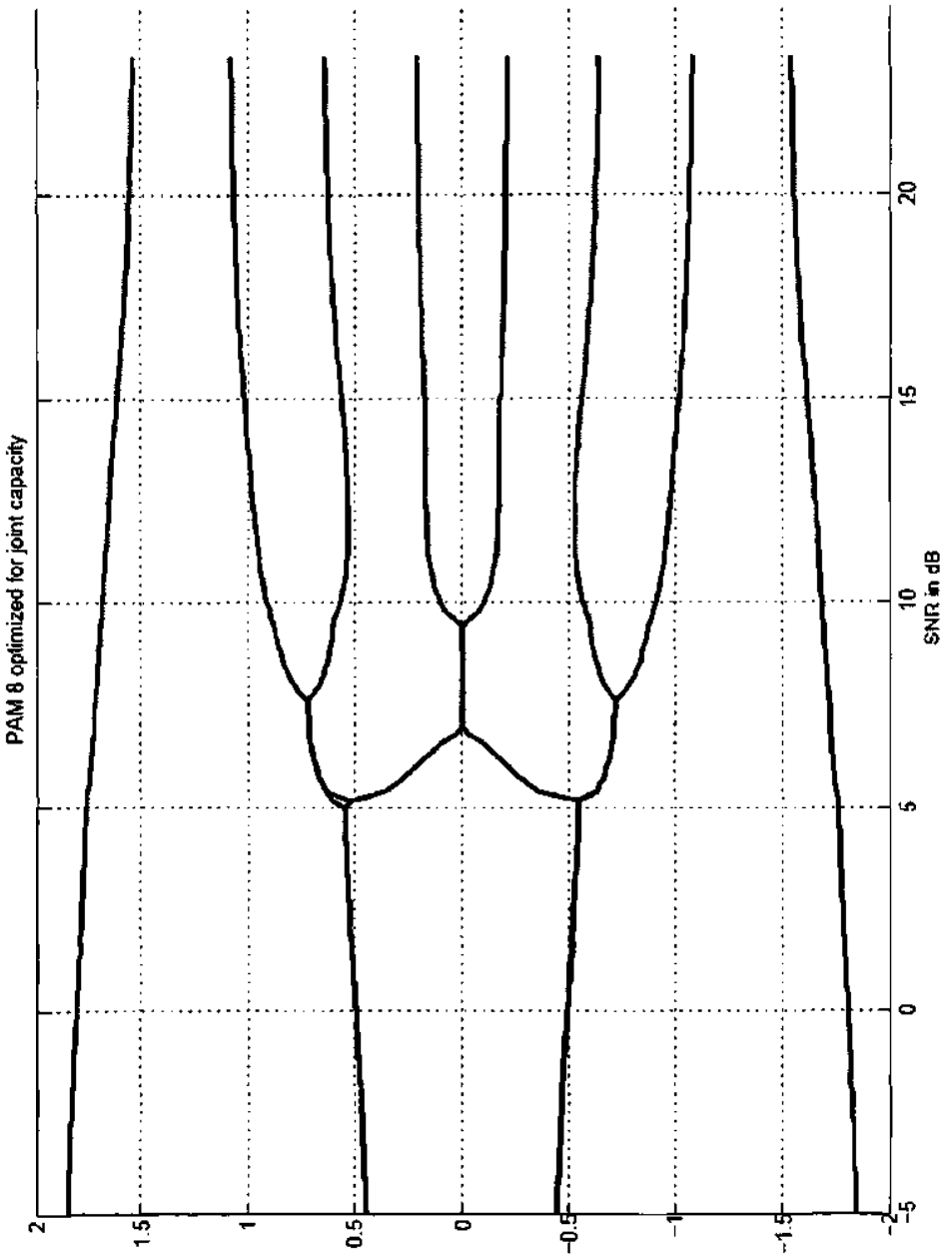

PAM 8 optimized for joint capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12b

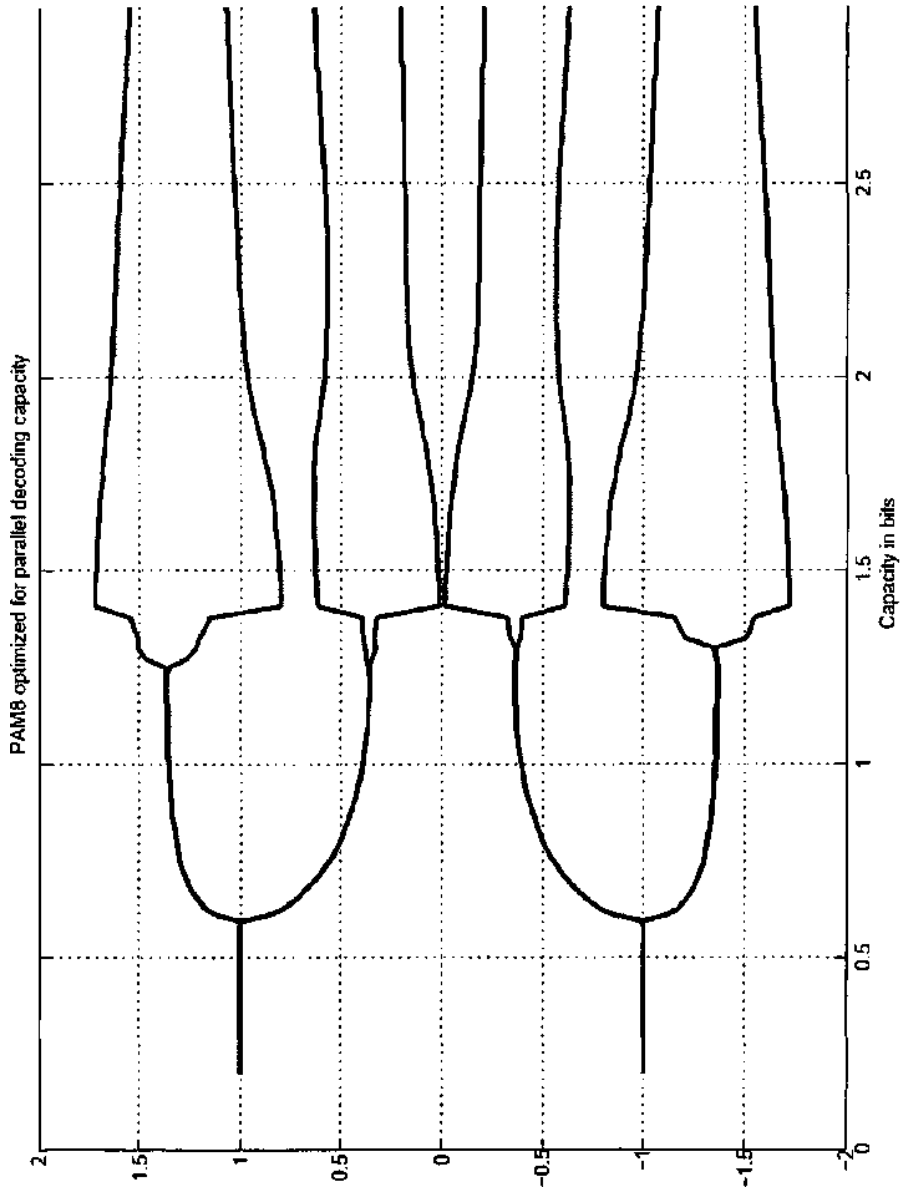

PAM8 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12c

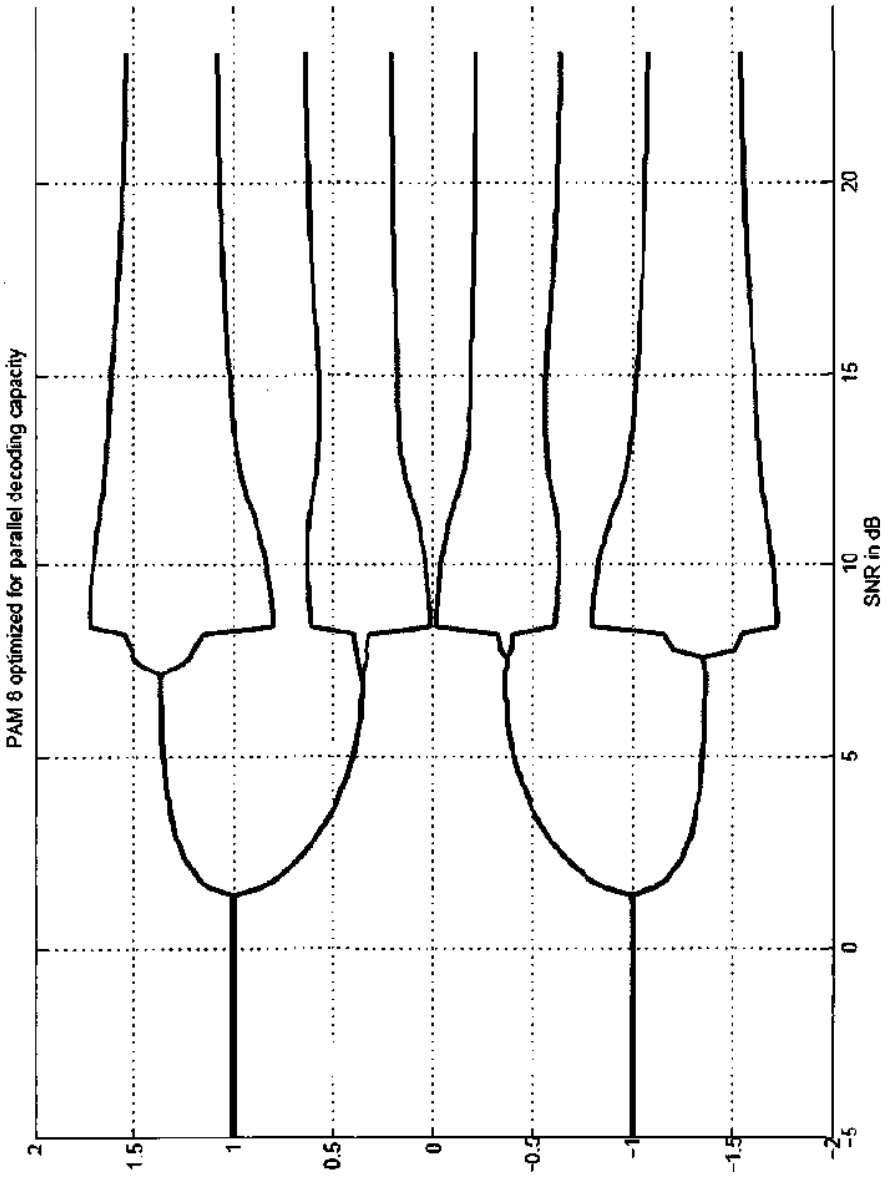

PAM 8 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

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

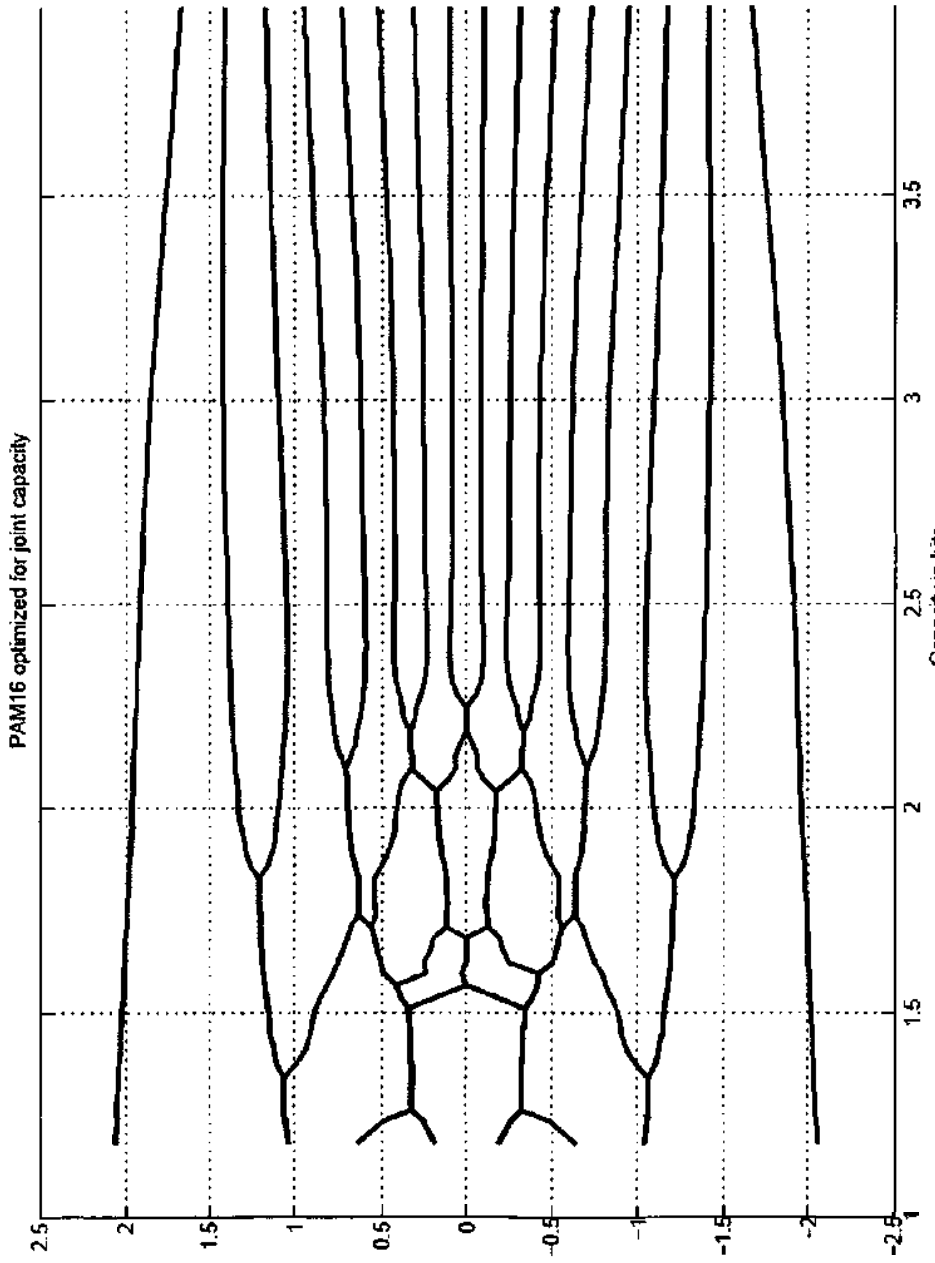

PAM16 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

FIG. 14a

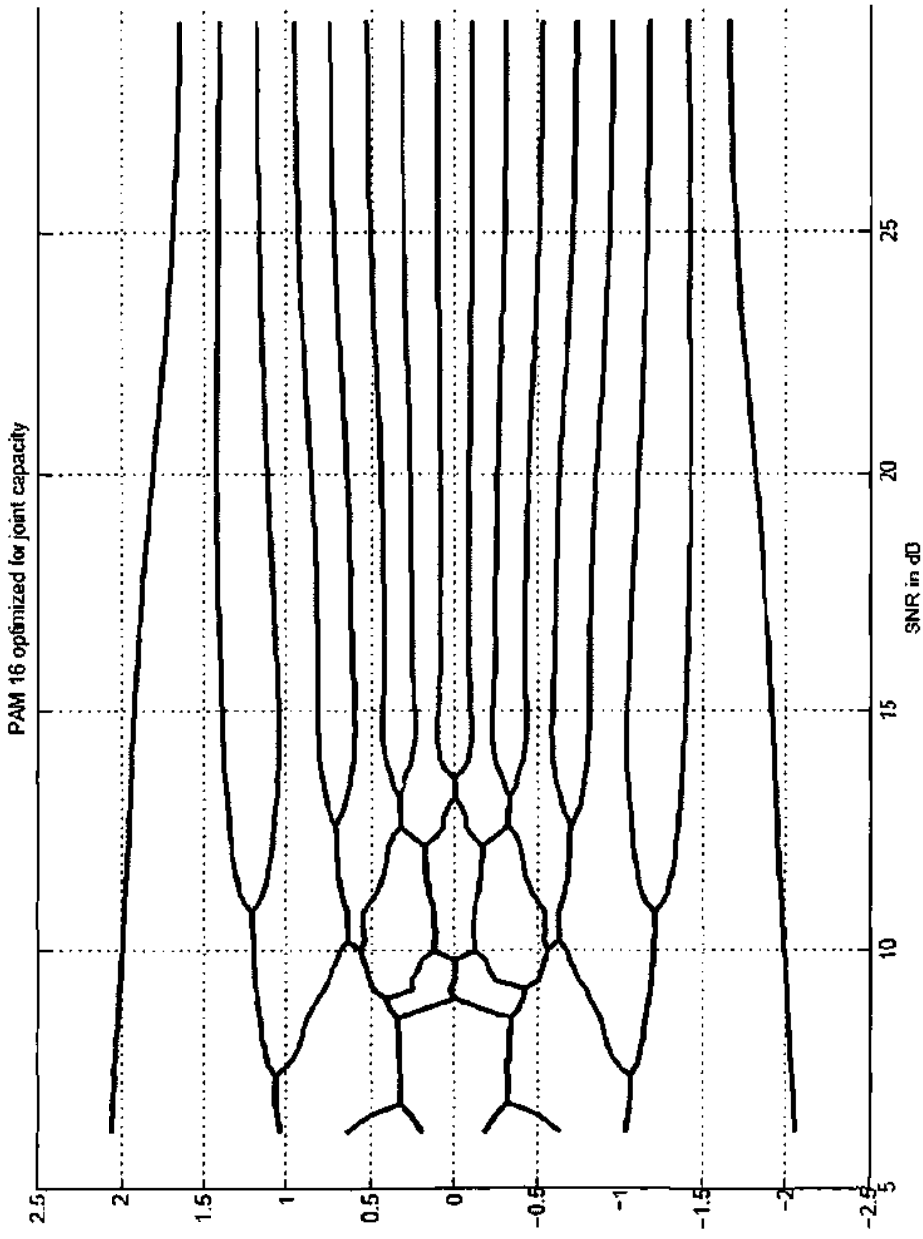

PAM 16 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

FIG. 14b

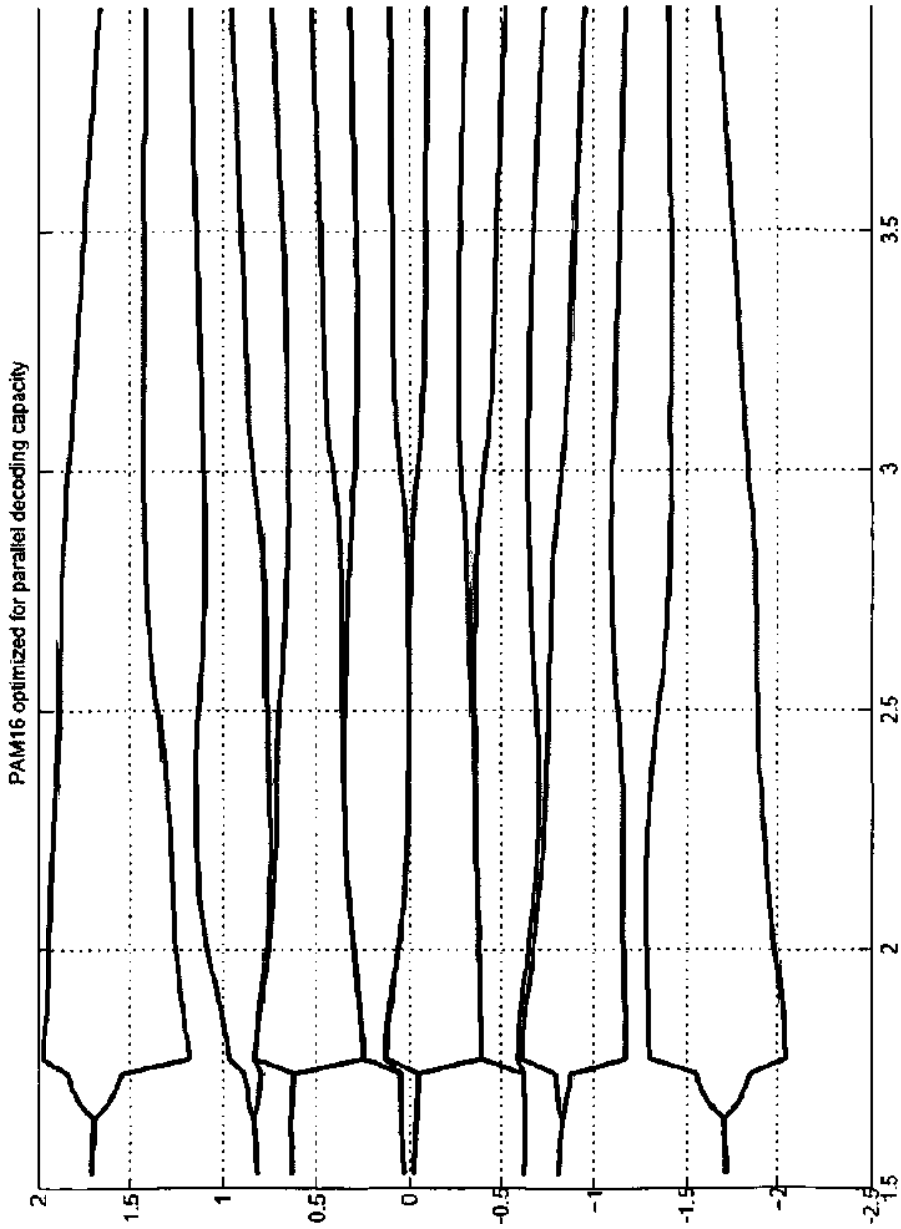

PAM16 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5

FIG. 14c

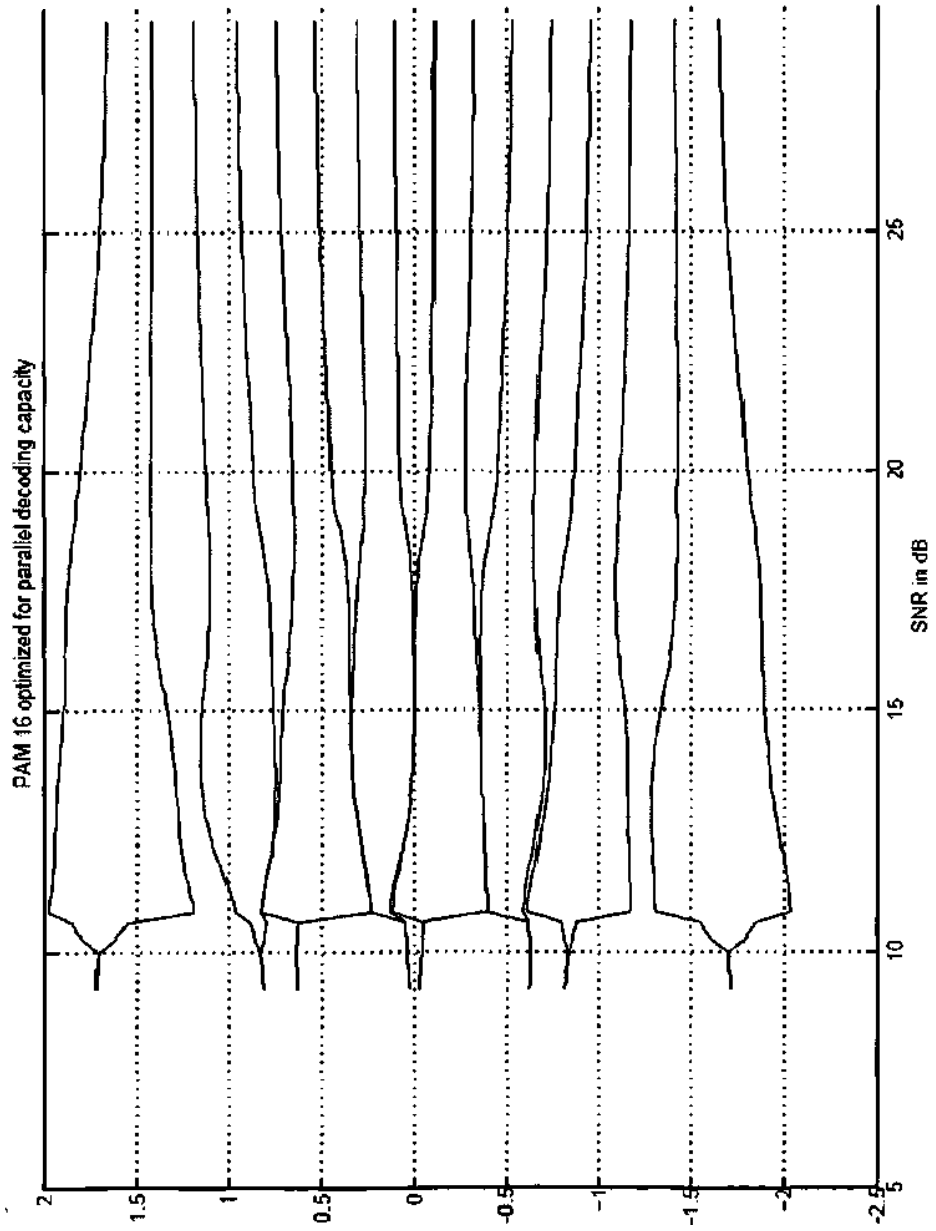

PAM 16 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

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

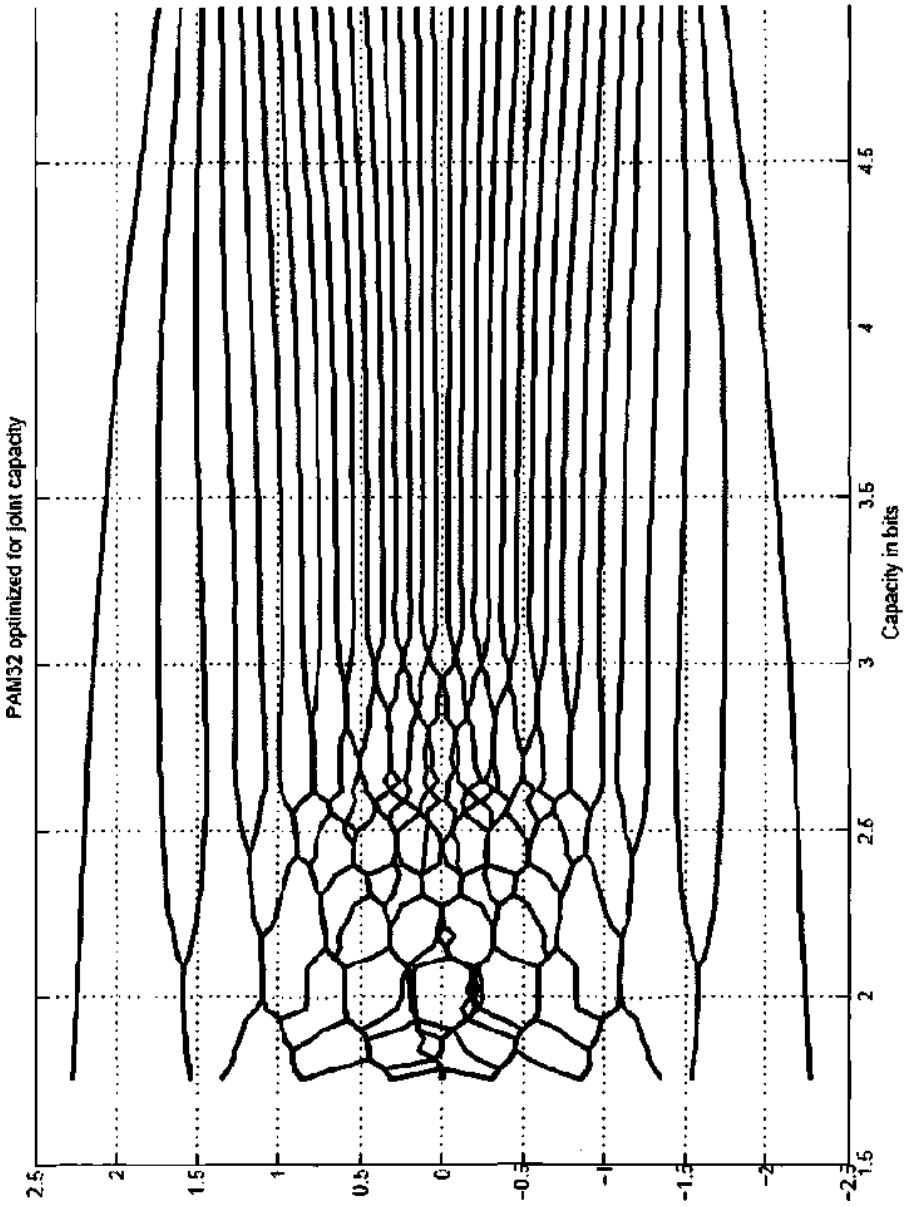

PAM32 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16a

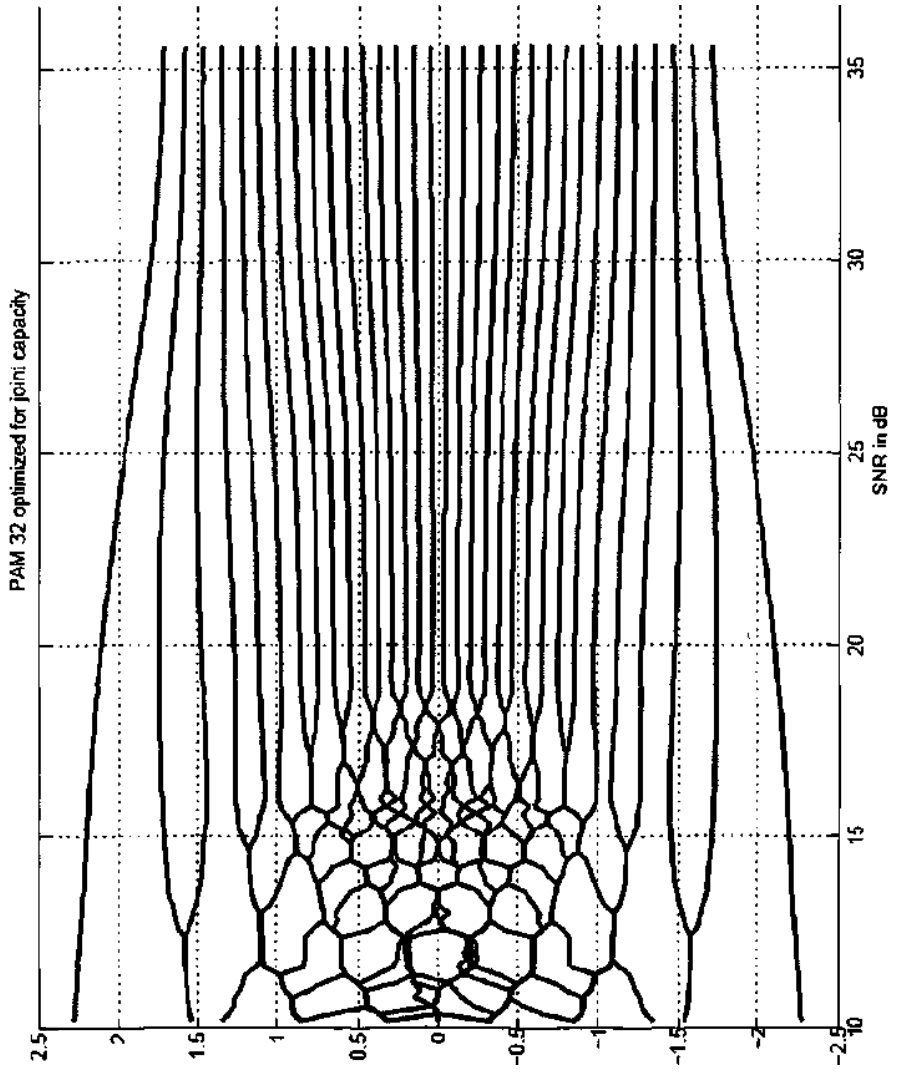

PAM 32 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

FIG. 16b

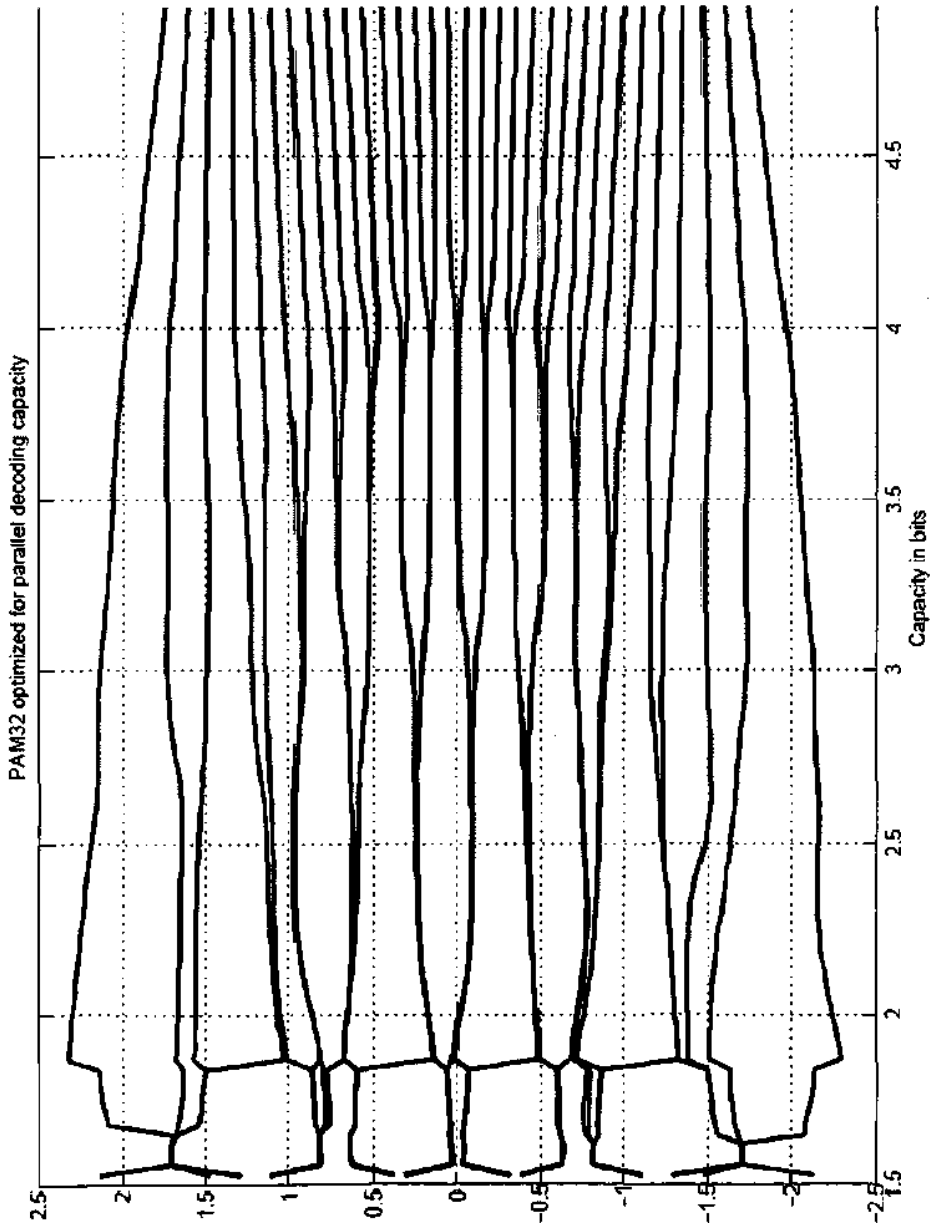

PAM32 optimized for parallel decoding capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16c

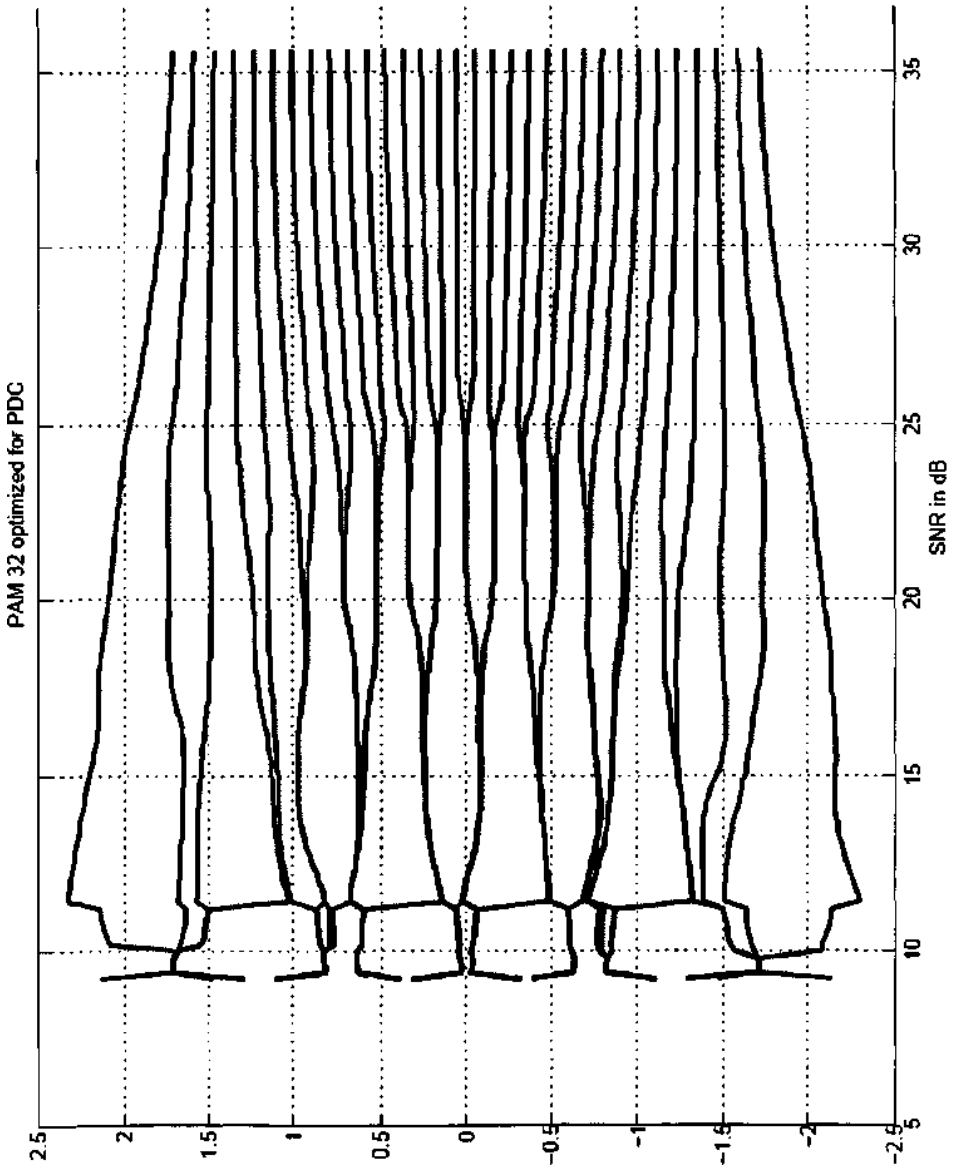

PAM 32 optimized for PDC
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

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

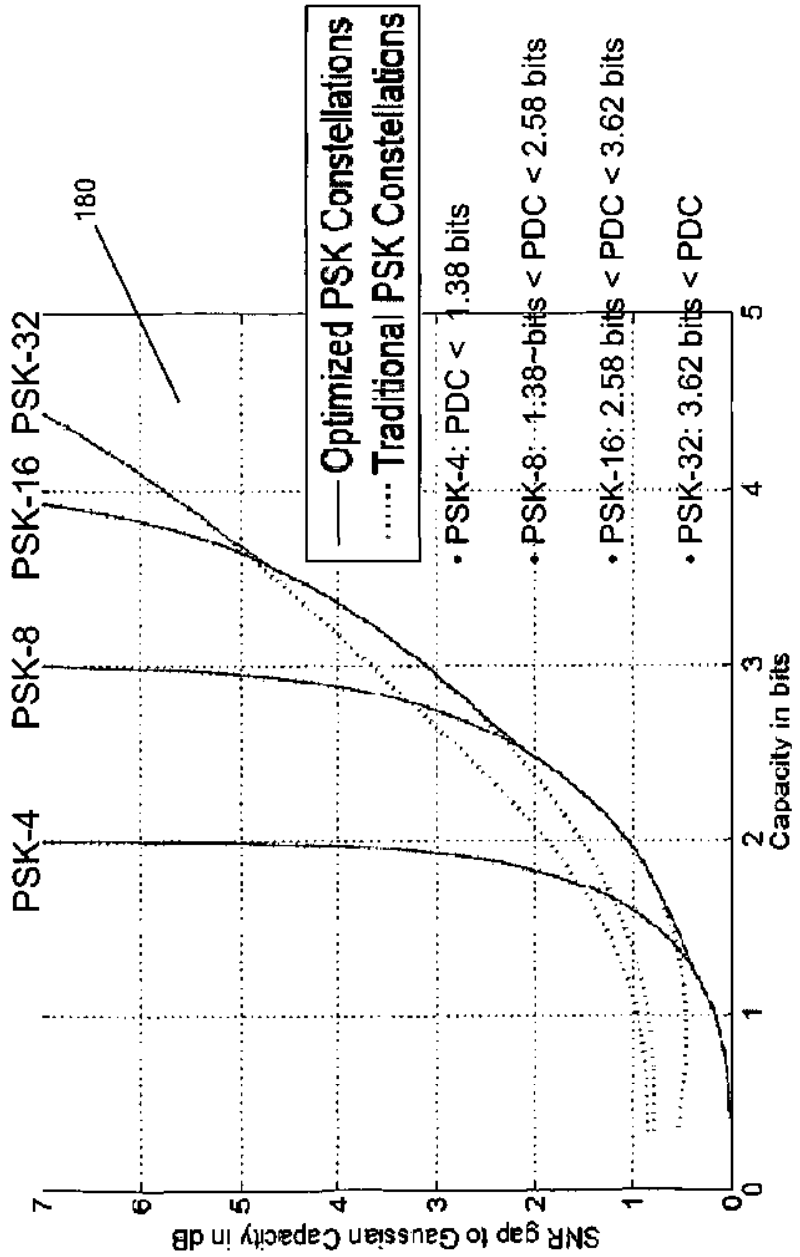

Optimized PSK Constellations
Traditional PSK Constellations
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38<bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
180
PSK-4
PSK-8
PSK-16
PSK-32
SNR gap to Gaussian Capacity in dB
Capacity in bits
0
1
2
3
4
5
6
7

FIG. 18

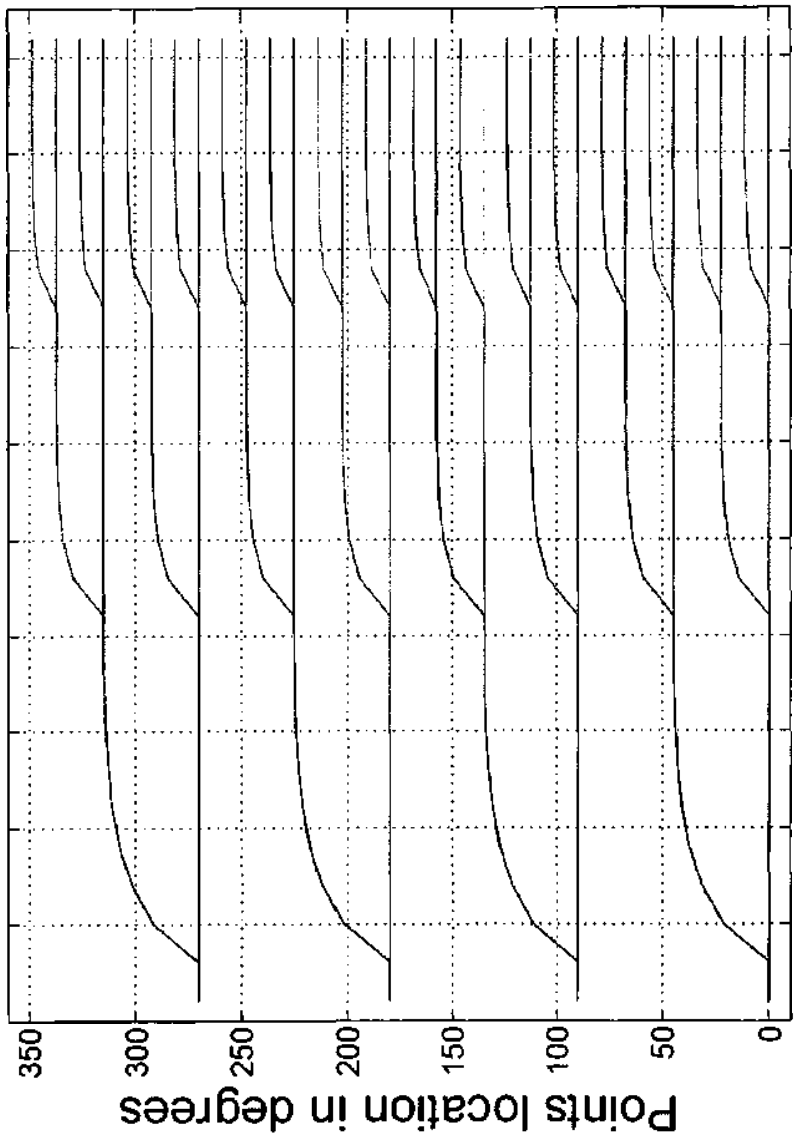

Points location in degrees
0
50
100
150
200
250
300
350
SNR in dB
0
2
4
6
8
10
12
14
16
18
20

FIG. 19

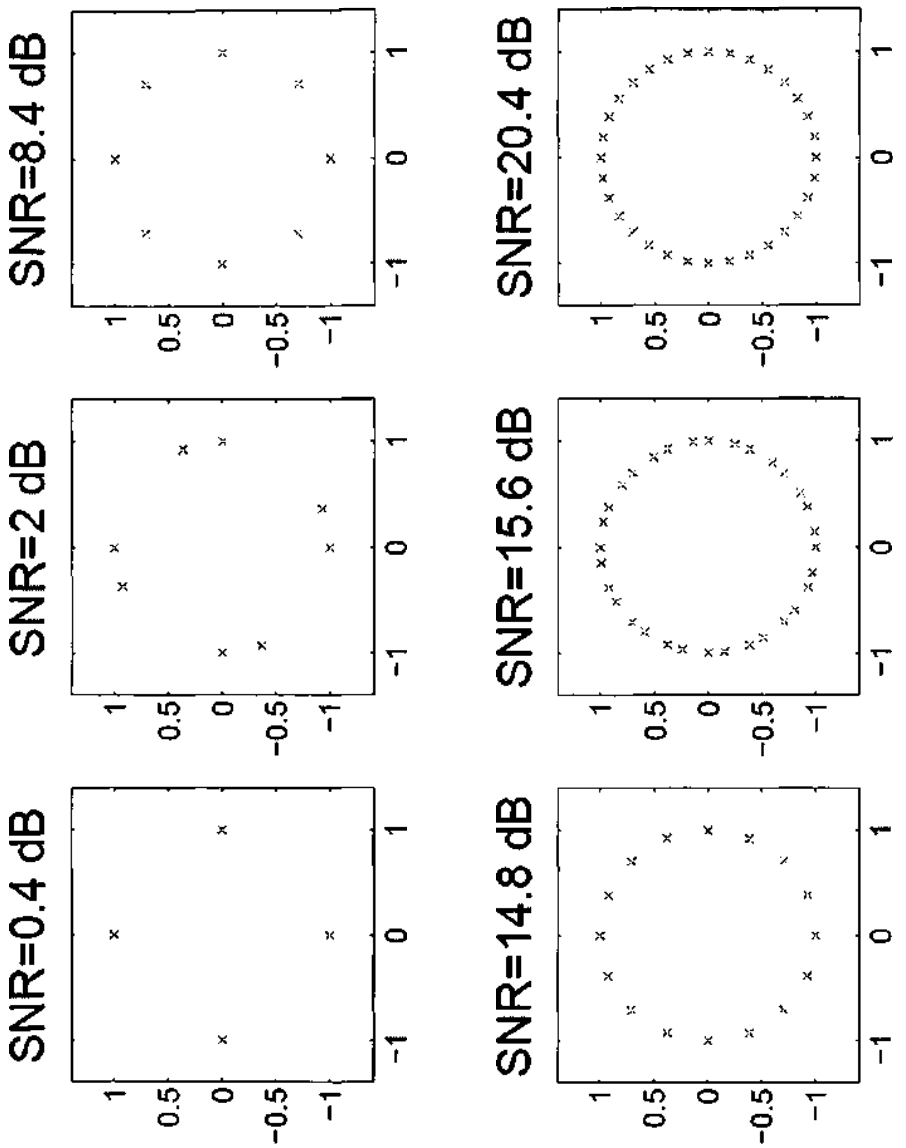


FIG. 20

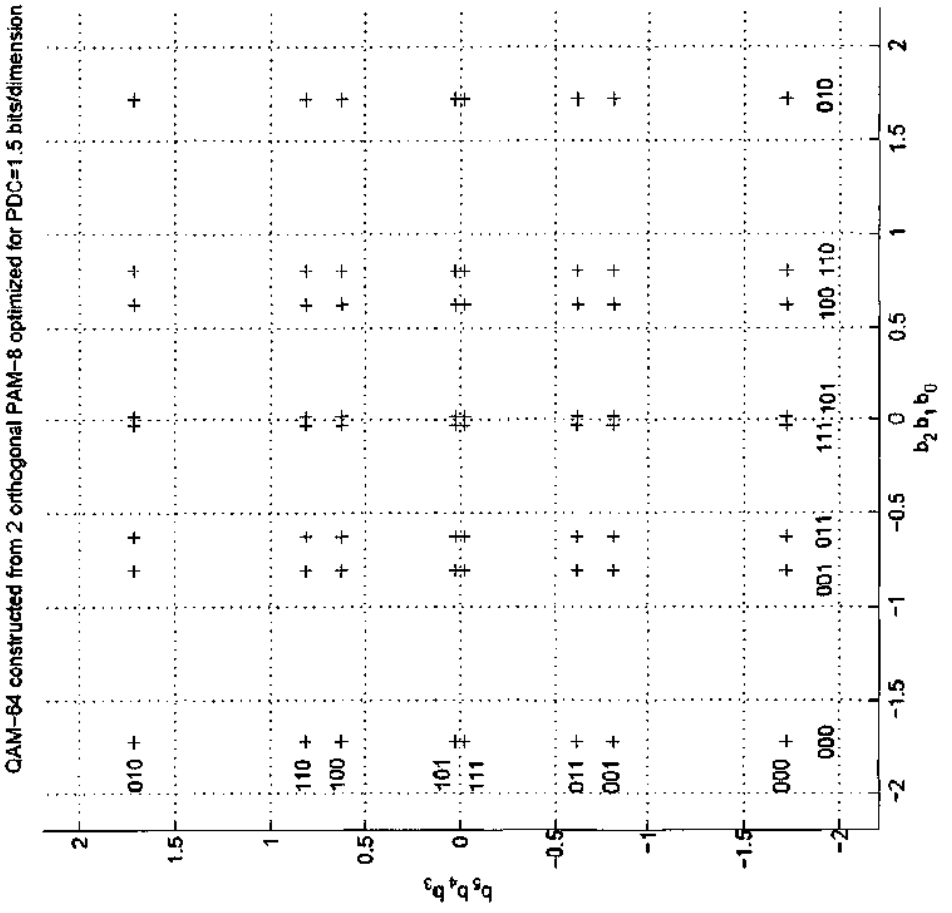

QAM-64 constructed from 2 orthogonal PAM-8 optimized for PDC=1.5 bits/dimension
b5b4b3
b2b1b0
010
110
100
101
111
011
001
000
000
001
011
111
101
100
110
010

FIG. 21

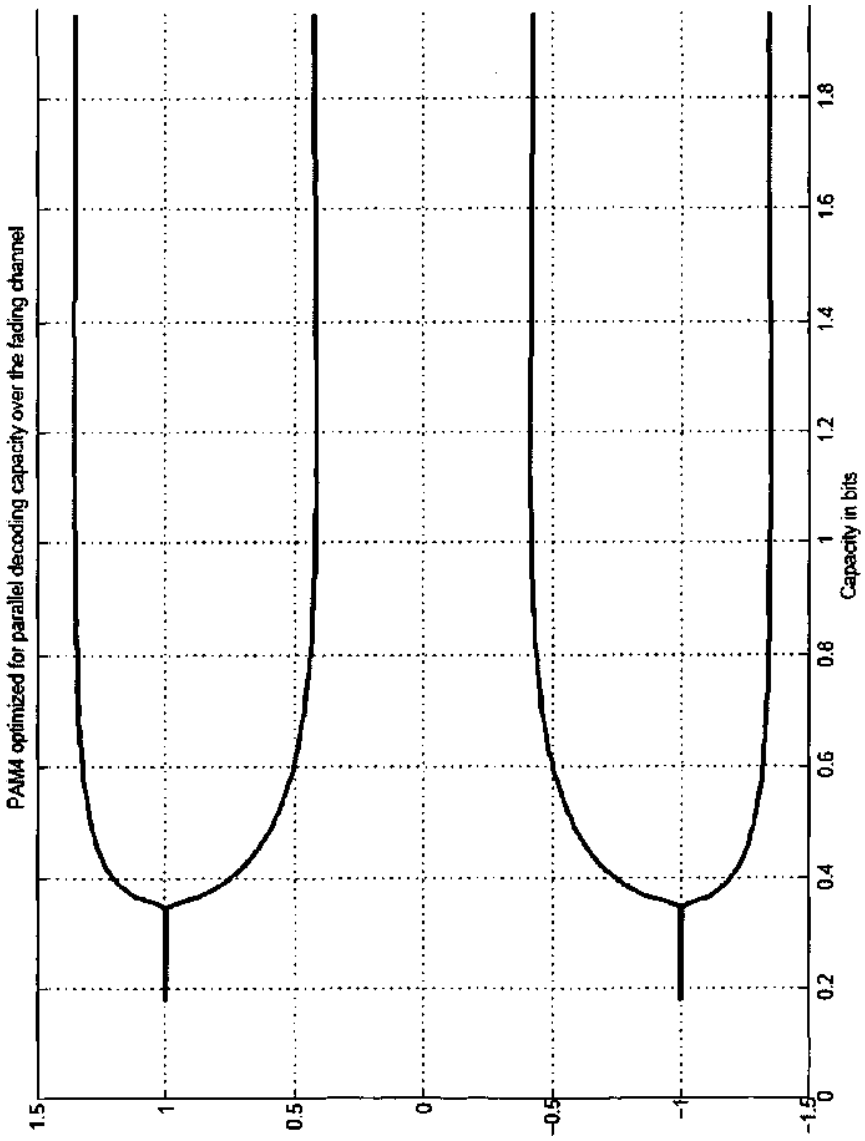

PAM4 optimized for parallel decoding capacity over the fading channel
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 22a

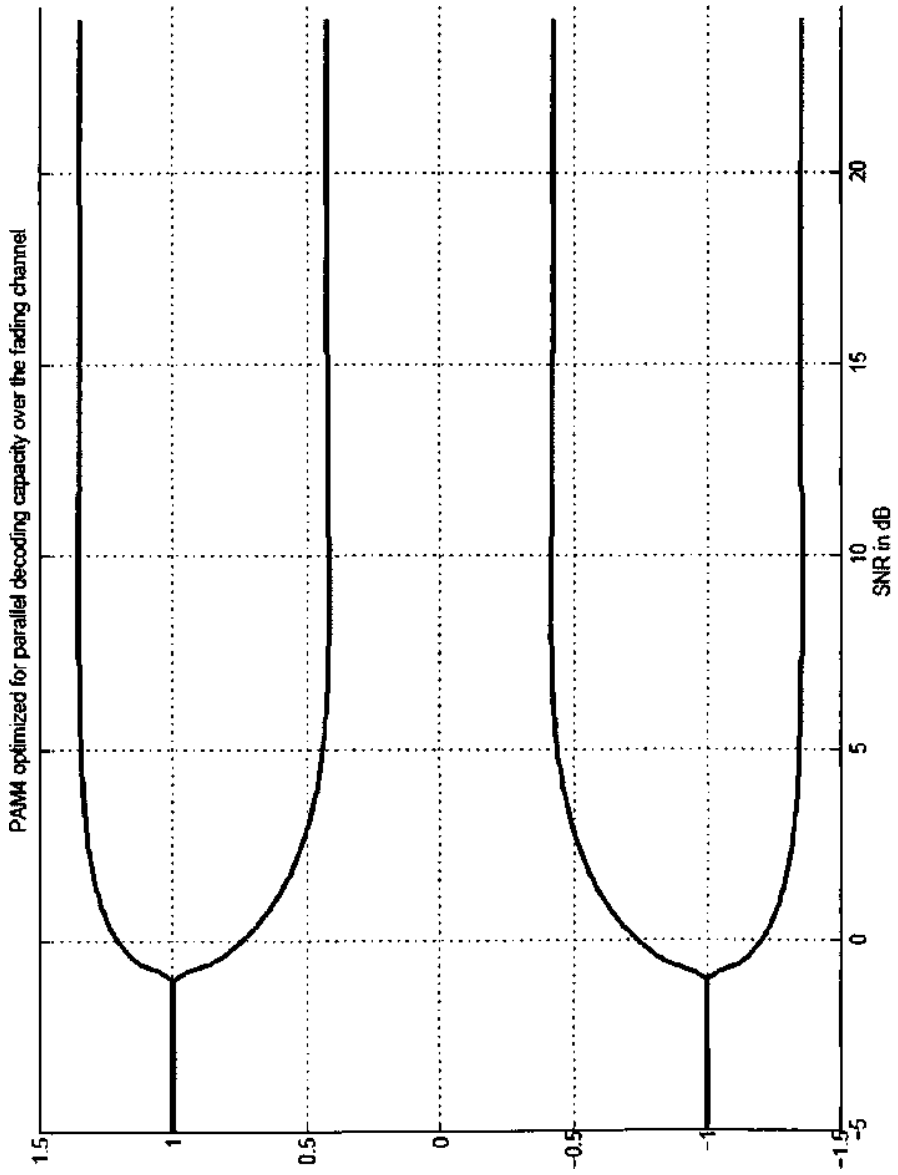

PAM4 optimized for parallel decoding capacity over the fading channel
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20

FIG. 22b

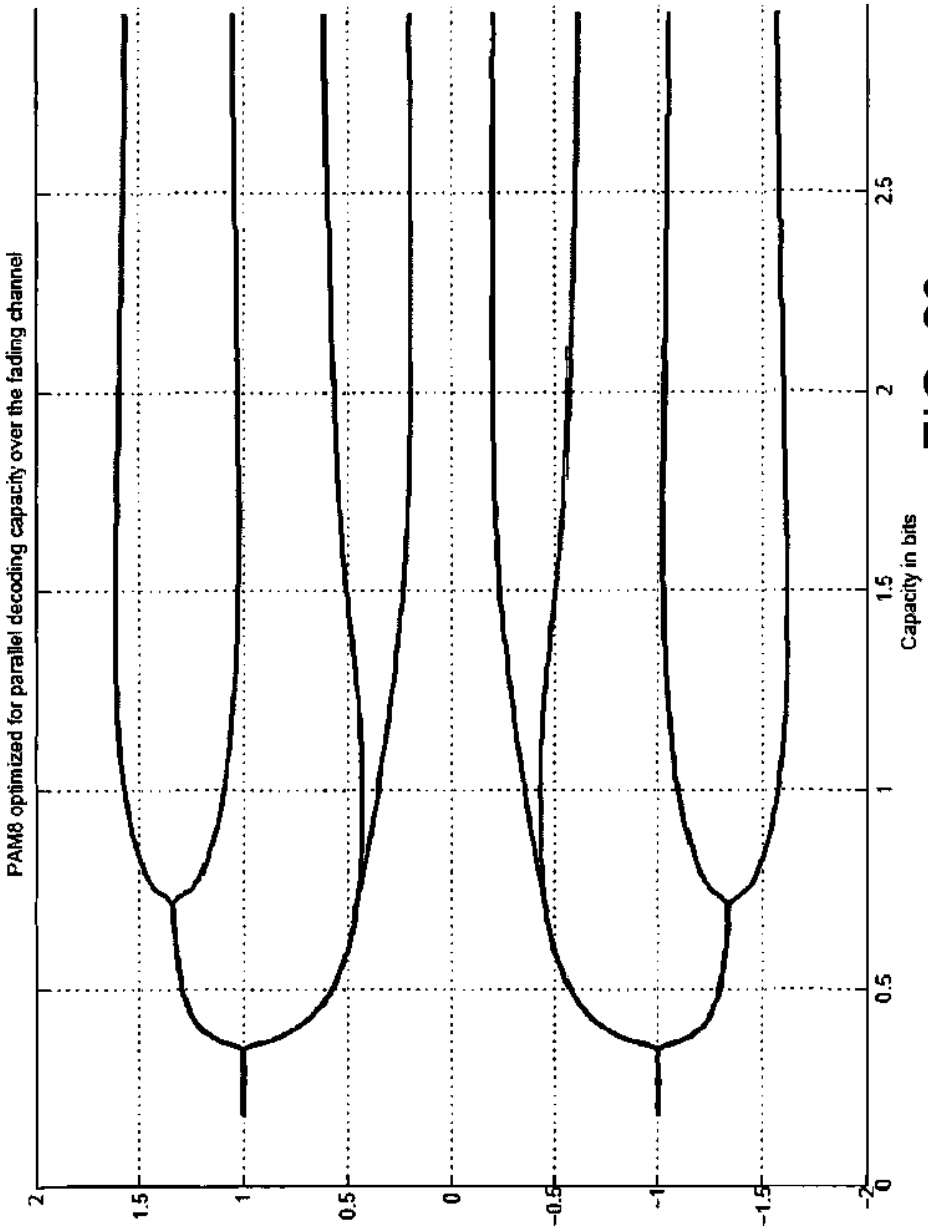

PAM8 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 23a

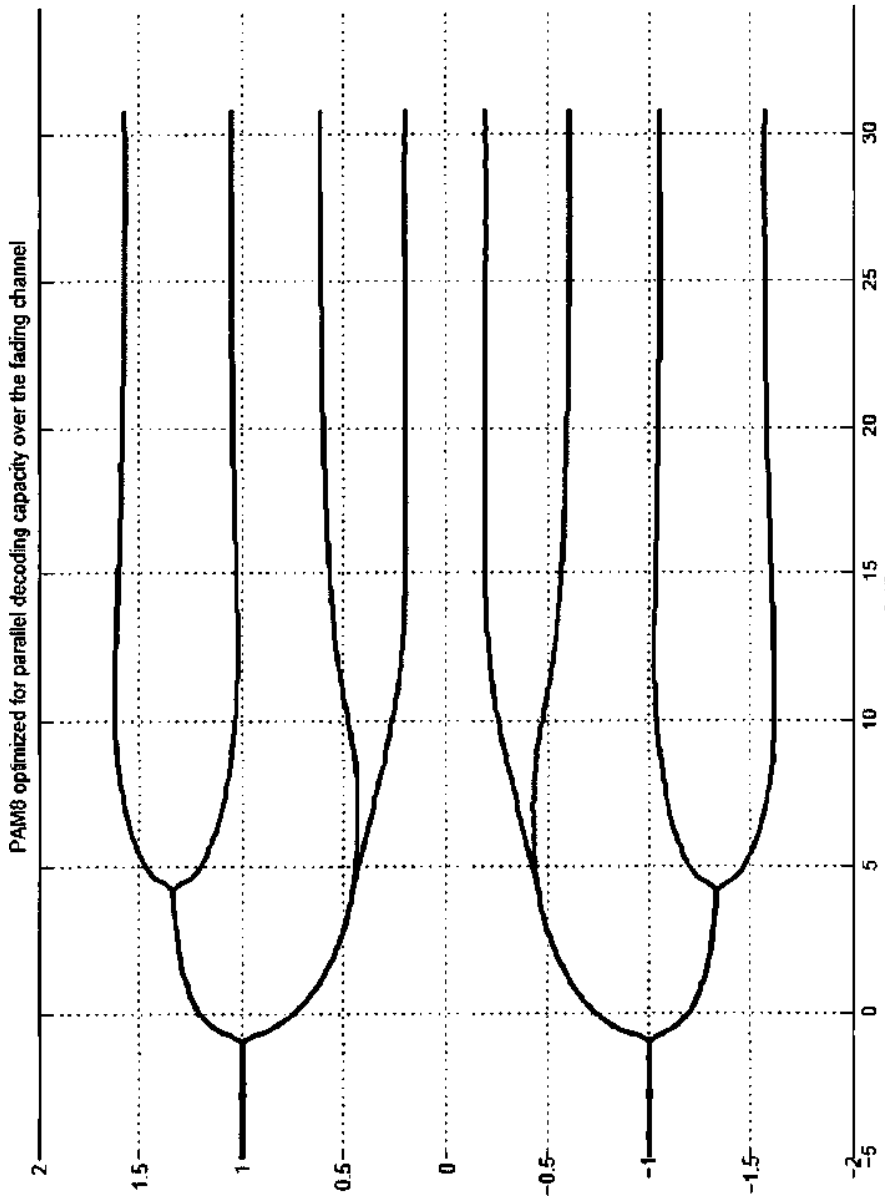

PAM8 optimized for parallel decoding capacity over the fading channel
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30
SNR in dB

FIG. 23b

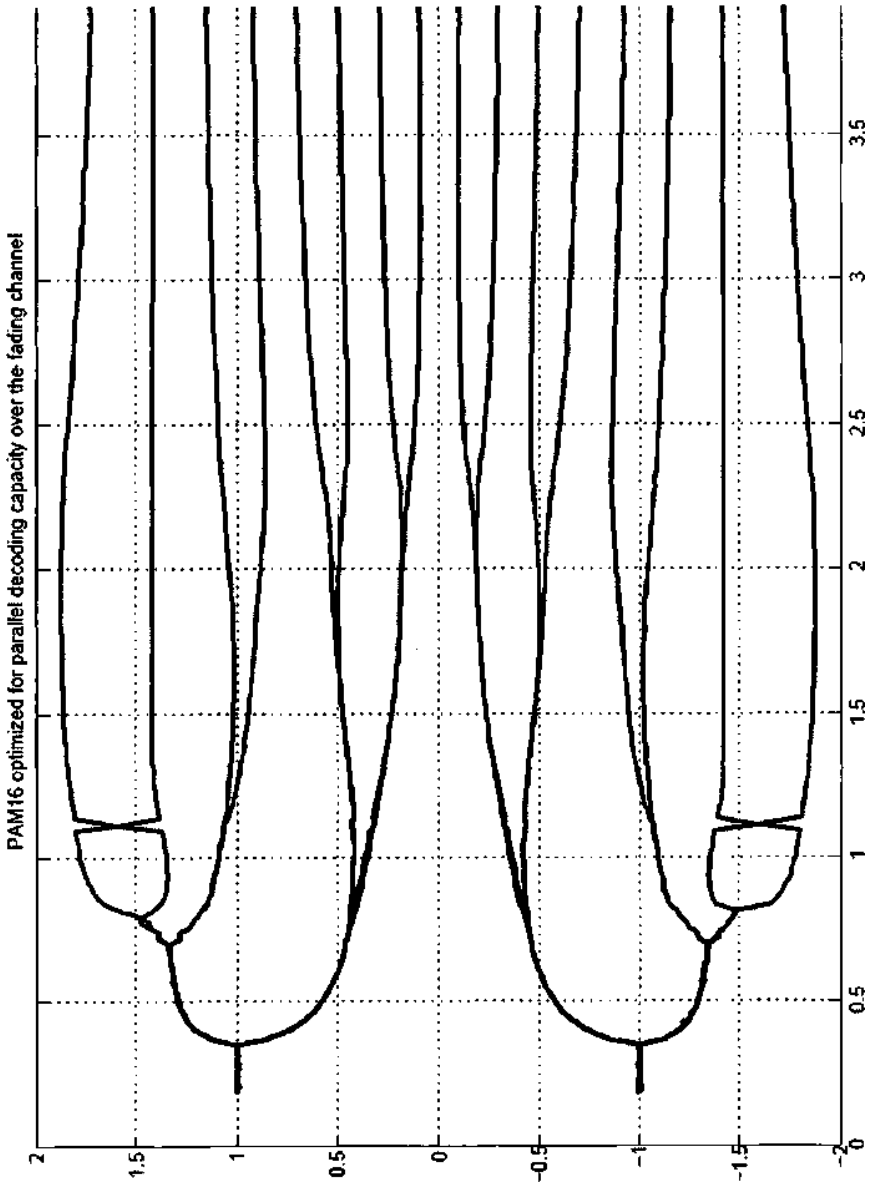

PAM16 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

FIG. 24a

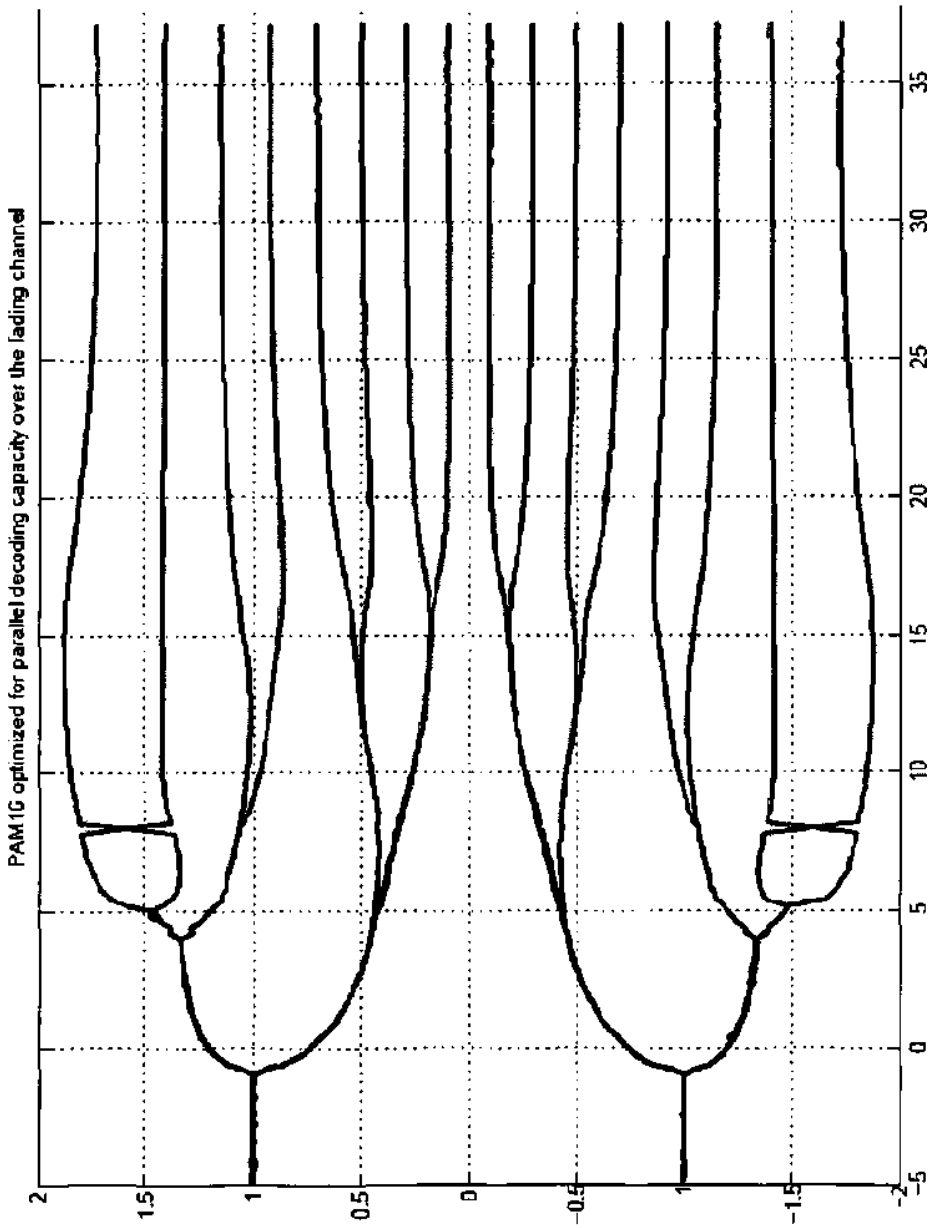

PAM16 optimized for parallel decoding capacity over the lading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
35
30
25
20
15
10
5
0
-5

FIG. 24b

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US2008/065994

A CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H04J1 1/00 (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 2007/0116181 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

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ES 2 712 914 T3

DESCRIPCIÓN

Método y aparato para la señalización con constelaciones de capacidad optimizada

Antecedentes

La presente invención se refiere generalmente a sistemas de transmisión digital de ancho de banda y/o eficientes energéticamente, y más específicamente al uso de constelaciones separadas desigualmente que tienen una mayor capacidad.

El término "constelación" se usa para describir los posibles símbolos que pueden transmitirse por un sistema de comunicación digital típico. Un receptor intenta detectar los símbolos que se transmitieron al asignar una señal recibida a la constelación. La distancia mínima (d_{min}) entre los puntos de constelación es indicativa de la capacidad de una constelación a altas relaciones de señal a ruido (SNR). Por lo tanto, las constelaciones usadas en muchos sistemas de comunicación se diseñan para maximizar d_{min}. El aumento de la dimensionalidad de una constelación permite una distancia mínima mayor para la energía constante de constelación por dimensión. Por lo tanto, se han diseñado varias constelaciones multidimensionales con buenas propiedades de distancia mínima.

Los sistemas de comunicación tienen una capacidad máxima teórica, que se conoce como el límite de Shannon. Muchos sistemas de comunicación intentan usar códigos para aumentar la capacidad de un canal de comunicación. Se han logrado importantes ganancias de codificación mediante el uso de técnicas de codificación tales como los códigos turbo y los códigos LDPC. Las ganancias de codificación que pueden lograrse mediante el uso de cualquier técnica de codificación se limitan por la constelación del sistema de comunicación. Puede pensarse que el límite de Shannon se basa en una constelación teórica conocida como una distribución gaussiana, que es una constelación infinita donde los símbolos en el centro de la constelación se transmiten con mayor frecuencia que los símbolos en el borde de la constelación. Las constelaciones prácticas son finitas y transmiten los símbolos con iguales probabilidades, y por lo tanto, tienen capacidades que son menores que la capacidad gaussiana. Se piensa que la capacidad de una constelación representa un límite en las ganancias que pueden lograrse mediante el uso de la codificación al usar esa constelación.

Se han hecho intentos previos para desarrollar constelaciones separadas desigualmente. Por ejemplo, se ha propuesto un sistema que usa constelaciones separadas desigualmente que se optimizan para minimizar la tasa de error de un sistema no codificado. Otro sistema propuesto usa una constelación con símbolos equiprobables pero separados desigualmente en un intento de imitar una distribución gaussiana.

Otros enfoques aumentan la dimensionalidad de una constelación o seleccionan un nuevo símbolo para transmitir al tomar en consideración los símbolos previamente transmitidos. Sin embargo, estas constelaciones aún se diseñaron según criterios de distancia mínima.

El uso de una constelación QAM no uniforme para mejorar la capacidad se ha descrito en el siguiente documento: D. Sommer y GP 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. La optimización de las constelaciones APSK se ha descrito en el siguiente documento: R. De Gaudenzi, A. Guillen i Fabregas y A. Martinez, "Performance analysis of turbo-coded APSK modulations over nonlinear satellite channels," en IEEE Transactions on Wireless Communications, vol. 5, núm. 9, pp. 2396-2407, septiembre de 2006.

Breve descripción de la invención

La invención se define y limita por el alcance de las reivindicaciones adjuntas 1-17. En la siguiente descripción, cualquier modalidad(es) a la(s) que se hace referencia y que no está(n) dentro del alcance de las reivindicaciones adjuntas, es(son) simplemente un ejemplo(s) útil(es) para la comprensión de la invención.

Los sistemas y métodos se describen para construir una modulación de manera que la capacidad restringida entre un transmisor y un receptor se aproxima al límite de capacidad del canal gaussiano descrito por primera vez por Shannon [ref Shannon 1948]. Los sistemas de comunicaciones tradicionales emplean modulaciones que dejan una brecha significativa en la capacidad gaussiana de Shannon. Las modulaciones de la presente invención reducen, y en algunos casos, casi eliminan esta brecha. La invención no requiere mecanismos de codificación especialmente diseñados que tienden a transmitir algunos puntos de una modulación con más frecuencia que otros, sino más bien proporciona un método para ubicar los puntos (en un espacio unidimensional o multidimensional) con el fin de maximizar la capacidad entre la entrada y la salida de un asignador y desasignador de bits o símbolos respectivamente. La aplicación práctica del método permite a los sistemas transmitir los datos a una velocidad determinada por menos energía o transmitir los datos a una velocidad mayor para la misma cantidad de energía.

Una modalidad de la invención incluye un transmisor configurado para transmitir señales a un receptor a través de un canal de comunicación, en donde el transmisor incluye un codificador configurado para recibir bits de usuario y bits de salida codificados a una tasa de bits de salida codificada expandida, un asignador configurado para asignar los bits codificados a los símbolos en una constelación de símbolos, un modulador configurado para generar una señal para la

transmisión a través del canal de comunicación mediante el uso de los símbolos generados por el asignador. Además, el receptor incluye un demodulador configurado para demodular la señal recibida a través del canal de comunicación, un desasignador configurado para estimar las probabilidades a partir de la señal demodulada, un descodificador que se configura para estimar los bits descodificados a partir de las probabilidades generadas por el desasignador. Además, la constelación de símbolos es una constelación de símbolos separada geométricamente de capacidad optimizada que proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza d_{min}.

Una modalidad adicional de la invención incluye codificar los bits de información del usuario mediante el uso de un esquema de codificación, asignar los bits codificados de información del usuario a una constelación de símbolos, en donde la constelación de símbolos es una constelación de símbolos separada geométricamente de capacidad optimizada que proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza d_{min}, modular los símbolos de acuerdo con un esquema de modulación, transmitir la señal modulada a través del canal de comunicación, recibir una señal modulada, demodular la señal modulada de acuerdo con el esquema de modulación, desasignar la señal demodulada mediante el uso de la constelación de señales conformada geométricamente para producir las probabilidades, y descodificar las probabilidades para obtener una estimación de los bits descodificados.

Otra modalidad de la invención incluye seleccionar un tamaño de la constelación apropiado y una capacidad deseada por dimensión, estimar una SNR inicial en la cual es probable que el sistema funcione, y optimizar iterativamente la ubicación de los puntos de la constelación para maximizar una medida de capacidad hasta que se haya alcanzado una mejora predeterminada en el rendimiento de la SNR de la constelación con relación a una constelación que maximiza d_{min}.

Aún una modalidad adicional de la invención incluye seleccionar un tamaño de la constelación apropiado y una capacidad deseada por dimensión, estimar una SNR inicial en la cual es probable que el sistema funcione, y optimizar iterativamente la ubicación de los puntos de la constelación para maximizar una medida de capacidad hasta que se haya alcanzado una mejora predeterminada en el rendimiento de la SNR de la constelación con relación a una constelación que maximiza d_{min}.

Aún otra modalidad de la invención incluye seleccionar un tamaño de constelación apropiado y una SNR deseada, y optimizar la ubicación de los puntos de la constelación para maximizar una medida de capacidad de la constelación.

Aún una modalidad adicional de la invención incluye obtener una constelación PAM conformada geométricamente con un tamaño de constelación que es la raíz cuadrada de dicho tamaño de constelación dado, donde la constelación PAM conformada geométricamente tiene una capacidad mayor que la de una constelación PAM que maximiza d_{min}, lo que crea una constelación PAM ortogonalizada que usa la constelación PAM conformada geométricamente, y combina la constelación PAM conformada geométricamente y la constelación PAM ortogonalizada para producir una constelación QAM conformada geométricamente.

Otra modalidad adicional de la invención incluye transmitir la información a través de un canal mediante el uso de una constelación de símbolos conformada geométricamente, y modificar la ubicación de los puntos dentro de la constelación de símbolos conformada geométricamente para cambiar la tasa de datos de usuario objetivo.

Breve descripción de los dibujos

La Figura 1 es una ilustración conceptual de un sistema de comunicación de acuerdo con una modalidad de la invención.
La Figura 2 es una ilustración conceptual de un transmisor de acuerdo con una modalidad de la invención.
La Figura 3 es una ilustración conceptual de un receptor de acuerdo con una modalidad de la invención.
La Figura 4a es una ilustración conceptual de la capacidad conjunta de un canal.
La Figura 4b es una ilustración conceptual de la capacidad de descodificación paralela de un canal.
La Figura 5 es un diagrama de flujo que muestra un proceso para obtener una constelación optimizada para la capacidad para su uso en un sistema de comunicación que tiene una tasa de código fija y un esquema de modulación de acuerdo con una modalidad de la invención.
La Figura 6a es un gráfico que muestra una comparación de la capacidad gaussiana y la capacidad de PD para las PAM-2,4,8,16,32 tradicionales.
La Figura 6b es un gráfico que muestra una comparación entre la capacidad gaussiana y la capacidad conjunta para las PAM-2,4,8,16,32 tradicionales.
La Figura 7 es un gráfico que muestra la brecha de la SNR a la capacidad gaussiana para la capacidad de PD y la capacidad conjunta de las constelaciones PAM-2,4,8,16,32 tradicionales.
La Figura 8a es un gráfico que compara la brecha de la SNR con la capacidad gaussiana de la capacidad de PD para las constelaciones PAM-2,4,8,16,32 tradicionales y optimizadas.
La Figura 8b es un gráfico que compara la brecha de la SNR con la capacidad gaussiana de la capacidad conjunta para las constelaciones PAM-2,4,8,16,32 tradicionales y optimizadas.
La Figura 9 es un gráfico que muestra el rendimiento de la tasa de error de trama de las constelaciones PAM-32 tradicionales y optimizadas para la capacidad de PD en simulaciones que involucran varios códigos LDPC de diferentes longitudes.

Las Figuras 10a - 10d son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-4 optimizada para la capacidad de PD y la capacidad conjunta versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 11a y 11b son tablas de diseño de constelaciones PAM-4 optimizadas para la capacidad de PD y la capacidad conjunta de acuerdo con modalidades de la invención.
Las Figuras 12a - 12d son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-8 optimizada para la capacidad de PD y la capacidad conjunta versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 13a y 13b son tablas de diseño de constelaciones PAM-8 optimizadas para la capacidad de PD y la capacidad conjunta de acuerdo con modalidades de la invención.
Las Figuras 14a - 14d son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-16 optimizada para la capacidad de PD y la capacidad conjunta versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 15a y 15b son tablas de diseño de constelaciones PAM-16 optimizadas para la capacidad de PD y la capacidad conjunta de acuerdo con modalidades de la invención.
Las Figuras 16a - 16d son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-32 optimizada para la capacidad de PD y la capacidad conjunta versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 17a y 17b son tablas de diseño de constelaciones PAM-32 optimizadas para la capacidad de PD y la capacidad conjunta de acuerdo con modalidades de la invención.
La Figura 18 es un gráfico que muestra la brecha de la SNR a la capacidad gaussiana para las constelaciones PSK de capacidad optimizada y tradicionales.
La Figura 19 es un gráfico que muestra la ubicación de los puntos de constelación de las constelaciones PSK-32 optimizadas para la capacidad de PD.
La Figura 20 es una serie de constelaciones PSK-32 optimizadas para la capacidad de PD en diferentes SNR.
La Figura 21 ilustra una QAM-64 construida a partir del producto cartesiano ortogonal de dos constelaciones PAM-8 optimizadas PD de acuerdo con una modalidad de la invención.
Las Figuras 22a y 22b son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-4 optimizada para la capacidad de PD a través de un canal de desvanecimiento versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 23a y 23b son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-8 optimizada para la capacidad de PD a través de un canal de desvanecimiento versus la tasa de bits del usuario por dimensión y versus la SNR.
Las Figuras 24a y 24b son gráficos de lugar que muestran la ubicación de los puntos de constelación de una constelación PAM-16 optimizada para la capacidad de PD a través de un canal de desvanecimiento versus la tasa de bits del usuario por dimensión y versus la SNR.

Descripción detallada de la invención

Con referencia ahora a los dibujos, se describen los sistemas de comunicación de acuerdo con las modalidades de la invención que usan las constelaciones de señales, que tienen puntos separados desigualmente (es decir, conformadas 'geométricamente'). En varias modalidades, las ubicaciones de los puntos conformados geométricamente se diseñan para proporcionar una medida de capacidad dada a una relación de señal a ruido (SNR) reducida en comparación con la SNR requerida por una constelación que maximiza d_{min}. En muchas modalidades, las constelaciones se seleccionan para proporcionar una mayor capacidad en un intervalo predeterminado de relaciones de señal a ruido (SNR) de canal. Las medidas de capacidad que pueden usarse en la selección de la ubicación de los puntos de constelación incluyen, pero no se limitan a, la capacidad de descodificación paralela (PD) y la capacidad conjunta.

En muchas modalidades, los sistemas de comunicación utilizan códigos próximos a la capacidad que incluyen, pero no se limitan a, códigos LDPC y Turbo. Como se describe adicionalmente más abajo, la optimización directa de los puntos de constelación de un sistema de comunicación que utiliza un código de canal próximo a la capacidad, puede producir diferentes constelaciones en dependencia de la SNR para la cual se optimizan. Por lo tanto, es poco probable que la misma constelación logre las mismas ganancias de codificación aplicadas en todas las tasas de código; es decir, la misma constelación no permitirá el mejor rendimiento posible en todas las tasas. En muchos casos, una constelación a una tasa de código puede lograr ganancias que no pueden lograrse a otra tasa de código. Los procesos para seleccionar las constelaciones de capacidad optimizada para lograr mayores ganancias de codificación basadas en una tasa de codificación específica de acuerdo con las modalidades de la invención se describen más abajo. En una serie de modalidades, los sistemas de comunicación pueden adaptar la ubicación de los puntos en una constelación en respuesta a las condiciones del canal, los cambios en la tasa de código y/o para cambiar la tasa de datos de usuario objetivo.

Sistemas de comunicación

Un sistema de comunicación de acuerdo con una modalidad de la invención se muestra en la Figura 1. El sistema de comunicación 10 incluye una fuente 12 que proporciona bits de usuario a un transmisor 14. El transmisor transmite símbolos a través de un canal a un receptor 16 mediante el uso de un esquema de modulación predeterminado. El receptor

usa el conocimiento del esquema de modulación para descodificar la señal recibida desde el transmisor. Los bits descodificados se proporcionan a un dispositivo sumidero que se conecta al receptor.

Un transmisor de acuerdo con una modalidad de la invención se muestra en la Figura 2. El transmisor 14 incluye un codificador 20 que recibe los bits de usuario desde una fuente y codifica los bits de acuerdo con un esquema de codificación predeterminado. En varias modalidades, se usa un código próximo a la capacidad, tal como un código turbo o un código LDPC. En otras modalidades, pueden usarse otros esquemas de codificación para proporcionar una ganancia de codificación dentro del sistema de comunicación. Un asignador 22 se conecta al codificador. El asignador asigna la salida de bits por el codificador a un símbolo dentro de una constelación de señales distribuidas geométricamente almacenada dentro del asignador. El asignador proporciona los símbolos a un modulador 24, que modula los símbolos para la transmisión a través del canal.

Un receptor de acuerdo con una modalidad de la invención se ilustra en la Figura 3. El receptor 16 incluye un demodulador 30 que demodula una señal recibida a través del canal para obtener probabilidades de símbolos o bits. El desasignador usa el conocimiento de la constelación de símbolos conformada geométricamente usada por el transmisor para determinar estas probabilidades. El desasignador 32 proporciona las probabilidades a un descodificador 34 que descodifica el flujo de bits codificado para proporcionar una secuencia de bits recibidos a un sumidero.

Constelaciones conformadas geométricamente

Los transmisores y receptores de acuerdo con las modalidades de la invención utilizan constelaciones de símbolos conformadas geométricamente. En varias modalidades, se usa una constelación de símbolos conformada geométricamente que optimiza la capacidad de la constelación. Más abajo se describen diversas constelaciones de símbolos conformadas geométricamente que pueden usarse de acuerdo con las modalidades de la invención y las técnicas para obtener las constelaciones de símbolos conformadas geométricamente.

Selección de constelaciones conformadas geométricamente

La selección de una constelación conformada geométricamente para su uso en un sistema de comunicación de acuerdo con una modalidad de la invención puede depender de una variedad de factores que incluyen si la tasa de código es fija. En muchas modalidades, una constelación conformada geométricamente se usa para reemplazar una constelación convencional (es decir, una constelación maximizada para d_{min}) en el asignador de transmisores y el desasignador de receptores dentro de un sistema de comunicación. La actualización de un sistema de comunicación implica la selección de una constelación y, en muchos casos, la actualización puede lograrse a través de una simple actualización de firmware. En otras modalidades, una constelación conformada geométricamente se selecciona junto con una tasa de código para cumplir con los requisitos de rendimiento específicos, que pueden incluir, por ejemplo, factores tales como una tasa de bits específica, una potencia de transmisión máxima. Los procesos para seleccionar una constelación geométrica cuando se actualizan los sistemas de comunicación existentes y cuando se diseñan nuevos sistemas de comunicación se describen adicionalmente más abajo.

Actualización de los sistemas de comunicación existentes

Una constelación conformada geométricamente que proporciona una capacidad, que es mayor que la capacidad de una constelación maximizada para d_{min}, puede usarse en lugar de una constelación convencional en un sistema de comunicación de acuerdo con las modalidades de la invención. En muchos casos, la sustitución de la constelación conformada geométricamente puede lograrse mediante una actualización de firmware o software de los transmisores y receptores dentro del sistema de comunicación. No todas las constelaciones conformadas geométricamente tienen una capacidad mayor que la de una constelación maximizada para d_{min}. Un enfoque para seleccionar una constelación conformada geométricamente que tenga una capacidad mayor que la de una constelación maximizada para d_{min}, es optimizar la forma de la constelación con respecto a una medida de la capacidad de la constelación para una SNR dada. Las medidas de capacidad que pueden usarse en el proceso de optimización pueden incluir, pero no se limitan a, la capacidad conjunta o la capacidad de descodificación paralela.

Capacidad conjunta y capacidad de descodificación paralela

Una constelación puede parametrizarse por el número total de puntos de constelación, M, y el número de dimensiones reales, N_{dim}. En los sistemas donde no hay iteraciones de propagación de creencias entre el descodificador y el desasignador de constelaciones, el desasignador de constelaciones puede considerarse como parte del canal. Un diagrama que ilustra conceptualmente las porciones de un sistema de comunicación que pueden considerarse parte del canal con el fin de determinar la capacidad de PD se muestra en la Figura 4a. Las porciones del sistema de comunicación que se consideran parte del canal se indican por la línea fantasma 40. La capacidad del canal definido como tal es la capacidad de descodificación paralela (PD), dada por:

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

donde Xi es el bit i del símbolo transmitido /-bits, e Y es el símbolo recibido, e I(A;B) denota la información mutua entre las variables aleatorias A y B.

Expresado de otra manera, la capacidad de PD de un canal puede observarse en términos de la información mutua entre los bits de salida del codificador (tal como un codificador LDPC) en el transmisor y las probabilidades calculadas por el desasignador en el receptor. La capacidad de PD se influencia tanto por la colocación de los puntos dentro de la constelación como por las asignaciones de etiquetado.

Con las iteraciones de propagación de creencias entre el desasignador y el descodificador, el desasignador ya no puede observarse como parte del canal, y la capacidad conjunta de la constelación se convierte en el límite conocido más estrecho en el rendimiento del sistema. Un diagrama que ilustra conceptualmente las porciones de un sistema de comunicación que se consideran parte del canal con el propósito de determinar la capacidad conjunta de una constelación se muestra en la Figura 4b. Las porciones del sistema de comunicación que se consideran parte del canal se indican por la línea fantasma 42. La capacidad conjunta del canal está dada por:

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

La capacidad conjunta es una descripción de la capacidad alcanzable entre la entrada del asignador en el lado de transmisión del enlace y la salida del canal (que incluye, por ejemplo, los canales AWGN y de desvanecimiento). Los sistemas prácticos a menudo deben 'desasignar' las observaciones del canal antes de la descodificación. En general, la etapa provoca alguna pérdida de capacidad. De hecho, puede probarse que $C_G \geq C_{CONJUNTA} \geq C_{PD}$. Es decir, $C_{CONJUNTA}$ limita por arriba la capacidad alcanzable por C_{PD}. Los métodos de la presente invención se motivan al considerar el hecho de que los límites prácticos para una capacidad de sistema de comunicación dada se limitan por $C_{CONJUNTA}$ y C_{PD}. En varias modalidades de la invención, se seleccionan constelaciones conformadas geométricamente que maximizan estas medidas.

Selección de una constelación que tiene una capacidad óptima

Las constelaciones conformadas geométricamente de acuerdo con las modalidades de la invención pueden diseñarse para optimizar las medidas de capacidad que incluyen, pero no se limitan a, la capacidad de PD o la capacidad conjunta. Un proceso para seleccionar los puntos, y el etiquetado, de una constelación conformada geométricamente para su uso en un sistema de comunicación que tiene una tasa de código fija de acuerdo con una modalidad de la invención se muestra en la Figura 5. El proceso 50 comienza con la selección (52) de un tamaño de constelación apropiado M y una capacidad deseada por dimensión η. En la modalidad ilustrada, el proceso implica una verificación (52) para asegurar que el tamaño de la constelación puede soportar la capacidad deseada. En el caso de que el tamaño de la constelación pueda soportar la capacidad deseada, entonces el proceso optimiza iterativamente la constelación M-aria para la capacidad especificada. La optimización de una constelación para una capacidad especificada a menudo implica un proceso iterativo, ya que la constelación óptima depende de la SNR en la que opera el sistema de comunicación. La SNR de la constelación óptima para dar una capacidad requerida no se conoce a priori. A lo largo de la descripción de la presente invención, la SNR se define como la relación entre la energía de constelación promedio por dimensión y la energía de ruido promedio por dimensión. En la mayoría de los casos, la capacidad puede configurarse para igualar la tasa de bits de usuario objetivo por símbolo por dimensión. En algunos casos, la adición de cierto margen de implementación en la parte superior de la tasa de bits de usuario objetivo podría resultar en un sistema práctico que puede proporcionar la tasa de usuario requerida a una velocidad menor. El margen depende del código. El siguiente procedimiento podría usarse para determinar la capacidad objetivo que incluye cierto margen en la parte superior de la tasa de usuario. En primer lugar, el código (por ejemplo, LDPC o Turbo) puede simularse junto con una constelación convencional separada igualmente. En segundo lugar, a partir de los resultados de la simulación puede encontrarse la SNR real de operación a la tasa de error requerida. En tercer lugar, puede calcularse la capacidad de la constelación convencional en esa SNR. Finalmente, una constelación conformada geométricamente puede optimizarse para esa capacidad.

En la modalidad ilustrada, el lazo de optimización iterativo implica seleccionar una estimación inicial de la SNR en la cual es probable que el sistema funcione (es decir, $SNR_{entrada}$). En varias modalidades, la estimación inicial es la SNR requerida mediante el uso de una constelación convencional. En otras modalidades, pueden usarse otras técnicas para seleccionar la SNR inicial. Luego, se obtiene una constelación M-aria al optimizar (56) la constelación para maximizar una medida de capacidad seleccionada en la estimación inicial de la $SNR_{entrada}$. A continuación, se describen varias técnicas para obtener una constelación optimizada para una estimación de la SNR dada.

Se determina (57) la SNR en la cual la constelación M-aria optimizada proporciona la capacidad deseada por dimensión η (SNR_{salida}). Se determina (58) si la SNR_{salida} y la $SNR_{entrada}$ han convergido. En la modalidad ilustrada, la convergencia se indica por la SNR_{salida} igual a la $SNR_{entrada}$. En varias modalidades, la convergencia puede determinarse en base a la diferencia entre la SNR_{salida} y la SNR que es menor que un umbral predeterminado. Cuando la SNR_{salida} y la $SNR_{entrada}$ no han convergido, el proceso realiza otra iteración que selecciona la SNR_{salida} como la nueva $SNR_{entrada}$ (55). Cuando la SNR_{salida} y la $SNR_{entrada}$ han convergido, se ha optimizado la medida de capacidad de la constelación. Como se explica en más detalle a continuación, las constelaciones de capacidad optimizada a bajas SNR son constelaciones conformadas geométricamente que pueden lograr ganancias de rendimiento significativamente más altas (medidas como la reducción en la SNR mínima requerida) que las constelaciones que maximizan d_{min}.

El proceso ilustrado en la Figura 5 puede maximizar la capacidad de PD o la capacidad conjunta de una constelación M-aria para una SNR dada. Aunque el proceso ilustrado en la Figura 5 muestra la selección de una constelación M-aria optimizada para la capacidad, podría usarse un proceso similar que termina tras la generación de una constelación M-aria donde la brecha de la SNR a la capacidad gaussiana en una capacidad dada es un margen predeterminado menor que la brecha de la SNR de una constelación convencional, por ejemplo 0.5 dB. Alternativamente, pueden usarse otros procesos que identifican las constelaciones M-arias que tienen una capacidad mayor que la capacidad de una constelación convencional de acuerdo con las modalidades de la invención. Una constelación conformada geométricamente de acuerdo con las modalidades de la invención puede lograr una capacidad mayor que la capacidad de una constelación que maximiza d_{min} sin tener la capacidad óptima para el intervalo de la SNR dentro del cual opera el sistema de comunicación.

Notamos que las constelaciones diseñadas para maximizar la capacidad conjunta también pueden ser particularmente adecuadas para códigos con símbolos sobre GF(q), o con descodificación de múltiples etapas. Por el contrario, las constelaciones optimizadas para la capacidad de PD podrían adaptarse mejor al caso más común de códigos con símbolos sobre GF(2).

Optimización de la capacidad de una constelación m-aria en una SNR dada

Los procesos para obtener una constelación de capacidad optimizada a menudo implican determinar la ubicación óptima para los puntos de una constelación M-aria en una SNR dada. Un proceso de optimización, tal como el proceso de optimización 56 mostrado en la Figura 5, típicamente implica una optimización no lineal no restringida o restringida. Las posibles funciones objetivas que deben maximizarse son las funciones de capacidad conjunta o de PD. Estas funciones pueden dirigirse a canales que incluyen, pero no se limitan a, canales de ruido gaussiano blanco aditivo (AWGN) o canales de desvanecimiento de Rayleigh. El proceso de optimización proporciona la ubicación de cada punto de constelación identificado por las etiquetas de sus símbolos. En el caso donde el objetivo es la capacidad conjunta, las etiquetas de los bits de puntos son irrelevantes, lo que significa que el cambio de las etiquetas de los bits no cambia la capacidad conjunta siempre y cuando el conjunto de ubicaciones de los puntos permanezca sin cambios.

El proceso de optimización típicamente encuentra la constelación que proporciona la mayor capacidad de PD o capacidad conjunta en una SNR dada. El proceso de optimización en sí, a menudo implica un proceso numérico iterativo que, entre otras cosas, considera varias constelaciones y selecciona la constelación que proporciona la mayor capacidad en una SNR dada. En otras modalidades, también puede encontrarse la constelación que requiere que la SNR mínima proporcione una capacidad de PD o capacidad conjunta requerida. Esto requiere ejecutar el proceso de optimización de forma iterativa como se muestra en la Figura 5.

Las restricciones de optimización en las ubicaciones de los puntos de constelación pueden incluir, pero no se limitan a, los límites inferior y superior en la ubicación de los puntos, la energía pico a promedio de la constelación resultante, y la media cero en la constelación resultante. Esto puede mostrar fácilmente que una constelación óptima a nivel mundial tendrá una media cero (sin componente DC). La inclusión explícita de una restricción de media cero ayuda a la rutina de optimización a converger más rápidamente. Excepto en los casos donde la búsqueda exhaustiva de todas las combinaciones de ubicaciones de puntos y etiquetas sea posible, este no siempre será necesariamente el caso de que las soluciones sean demostrablemente óptimas a nivel mundial. En los casos donde es posible realizar una búsqueda exhaustiva, la solución proporcionada por el optimizador no lineal es, de hecho, óptima a nivel mundial.

Los procesos descritos anteriormente proporcionan ejemplos de la manera en que puede obtenerse una constelación conformada geométricamente que tiene una mayor capacidad con relación a una capacidad convencional para su uso en un sistema de comunicación que tiene una tasa de código fija y un esquema de modulación. Las ganancias reales que pueden lograrse mediante el uso de constelaciones que se optimizan para la capacidad en comparación con las constelaciones convencionales que maximizan d_{min} se consideran a continuación.

Ganancias logradas por las constelaciones optimizadas separadas geométricamente

Se cree que la capacidad teórica máxima que puede lograrse mediante cualquier método de comunicación es la capacidad gaussiana, CG, que se define como:

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

Donde la relación de señal a ruido (SNR) es la relación entre la energía de la señal esperada y la energía del ruido esperado. La brecha que queda entre la capacidad de una constelación y la CG puede considerarse como una medida de la calidad de un diseño de constelación dado.

La brecha en la capacidad entre un esquema de modulación convencional en combinación con un codificador teóricamente óptimo puede observarse con referencia a las Figuras 6a y 6b. La Figura 6a incluye un gráfico 60 que muestra una comparación entre la capacidad gaussiana y la capacidad de PD de las constelaciones PAM-2, 4, 8, 16 y 32

convencionales que maximizan d_{min}, las brechas 62 existen entre el trazo de la capacidad gaussiana y la capacidad de PD de las diversas constelaciones PAM. La Figura 6b incluye un gráfico 64 que muestra una comparación entre la capacidad gaussiana y la capacidad conjunta de las constelaciones PAM-2, 4, 8, 16 y 32 convencionales que maximizan d_{min}, las brechas 66 existen entre el trazo de la capacidad gaussiana y la capacidad conjunta de las diversas constelaciones PAM. Estas brechas en la capacidad representan la medida en que las constelaciones PAM convencionales no llegan a obtener la capacidad teórica máxima, es decir, la capacidad gaussiana.

Con el fin de obtener una mejor vista de las diferencias entre las curvas mostradas en las Figuras 6a y 6b en puntos cercanos a la capacidad gaussiana, la brecha de la SNR a la capacidad gaussiana para diferentes valores de capacidad para cada constelación se representa en la Figura 7. Es interesante observar a partir del gráfico 70 en la Figura 7 que (a diferencia de la capacidad conjunta) en la misma SNR, la capacidad de PD no aumenta necesariamente con el número de puntos de constelación. Como se describe adicionalmente más abajo, este no es el caso con las constelaciones PAM optimizadas para la capacidad de PD.

Las Figuras 8a y 8b resumen el rendimiento de las constelaciones para PAM-4, 8, 16 y 32 optimizadas para la capacidad de PD y la capacidad conjunta (debe señalarse que BPSK es la constelación PAM-2 óptima en todas las tasas de código). Las constelaciones se optimizan para la capacidad de PD y la capacidad conjunta para diferentes bits de usuario objetivo por dimensión (es decir, tasas de código). Las constelaciones optimizadas son diferentes en dependencia de los bits de usuario objetivo por dimensión, y además dependen de si se han diseñado para maximizar la capacidad de PD o la capacidad conjunta. Todas las constelaciones PAM optimizadas de PD se etiquetan mediante el uso de un etiquetado gris que no siempre es el etiquetado gris reflectivo binario. Debe señalarse que no todas las etiquetas grises alcanzan la capacidad de PD máxima posible, incluso al tener la libertad de colocar los puntos de constelación en cualquier lugar en la línea real. La Figura 8a muestra la brecha de la SNR para cada constelación optimizada para la capacidad de PD. La Figura 8b muestra la brecha de la SNR a la capacidad gaussiana para cada constelación optimizada para la capacidad conjunta. Nuevamente, debe enfatizarse que cada ' ' en el gráfico representa una constelación diferente.

Con referencia a la Figura 8a, la ganancia de codificación lograda mediante el uso de una constelación optimizada para la capacidad de PD puede apreciarse mediante la comparación de la brecha de la SNR a una tasa de bits de usuario por dimensión de 2.5 bits para la PAM-32. Una tasa de bits de usuario por dimensión de 2.5 bits para un sistema que transmite 5 bits por símbolo constituye una tasa de código de 1/2. A esa tasa de código, la constelación optimizada para la capacidad de PD proporciona una ganancia de codificación adicional de aproximadamente 1.5 dB en comparación con la constelación PAM-32 convencional.

Las ganancias de la SNR que pueden lograrse mediante el uso de constelaciones que se optimizan para la capacidad de PD pueden verificarse a través de la simulación. Los resultados de una simulación realizada mediante el uso de un código LDPC de tasa 1/2 junto con una constelación PAM-32 convencional y junto con una constelación PAM-32 optimizada para la capacidad de PD se ilustran en la Figura 9. Un gráfico 90 incluye trazos del rendimiento de la tasa de error de trama de las diferentes constelaciones con respecto a la SNR y el uso de códigos de diferentes longitudes (es decir, k=4,096 y k=16,384). Independientemente del código que se usa, la constelación optimizada para la capacidad de PD alcanza una ganancia de aproximadamente 1.3 dB, que se acerca mucho a la ganancia prevista de la Figura 8A.

Constelaciones PAM de capacidad optimizada

Mediante el uso de los procesos descritos anteriormente, pueden generarse gráficos de lugar de las constelaciones PAM optimizadas para la capacidad que muestran la ubicación de los puntos dentro de las constelaciones PAM versus la SNR. Los gráficos de lugar de las constelaciones PAM-4, 8, 16 y 32 optimizadas para la capacidad de PD y la capacidad conjunta, y las tablas de diseño correspondientes a varias tasas de bits de usuario típicas por dimensión se ilustran en las Figuras 10a - 17b. Los gráficos de lugar y las tablas de diseño muestran las ubicaciones de los puntos de constelación PAM-4,8,16,32 y las etiquetas de la SNR de baja a alta correspondientes a un intervalo de eficiencia espectral de baja a alta.

En la Figura 10a, un gráfico de lugar 100 muestra la ubicación de los puntos de las constelaciones PAM-4 optimizadas para la capacidad conjunta representada contra la capacidad alcanzada. Un gráfico de lugar similar 105 que muestra la ubicación de los puntos de las constelaciones PAM-4 optimizadas para la capacidad conjunta representadas contra la SNR se incluye en la Figura 10b. En la Figura 10c se grafica la ubicación de los puntos para la PAM-4 optimizada para la capacidad de PD contra la capacidad alcanzable, y en la Figura 10d se grafica la ubicación de los puntos para la PAM-4 para la capacidad de PD contra la SNR. A bajas SNR, las constelaciones PAM-4 optimizadas para la capacidad de PD tienen sólo 2 puntos únicos, mientras que las constelaciones optimizadas para la capacidad conjunta tienen 3. A medida que aumenta la SNR, cada optimización proporciona finalmente 4 puntos únicos. Este fenómeno se describe explícitamente en la Figura 11a y la Figura 11b donde se capturan cortes verticales de las Figuras 10ab y 10cd en tablas que describen algunos diseños de interés de las constelaciones PAM-4. Los cortes de la SNR seleccionados representan los diseños que alcanzan las capacidades = {0.5, 0.75, 1.0, 1.25, 1.5} bits por símbolo (bps). Dado que la PAM-4 puede proporcionar como máximo $\log_2(4) = 2$ bps, estos puntos de diseño representan sistemas con tasas de código de información R = {1/4, 3/8, 1/2, 5/8, 3/4} respectivamente.

Las Figuras 12ab y 12cd presentan gráficos de lugar de puntos de constelaciones PAM-8 optimizadas para la capacidad de PD y la capacidad conjunta versus la capacidad alcanzable y la SNR. Las Figuras 13a y 13b proporcionan cortes de estos gráficos en las SNR correspondientes a capacidades alcanzables $\eta = \{0.5, 1.0, 1.5, 2.0, 2.5\}$ bps. Cada uno de estos cortes corresponde a sistemas con tasa de código $R = \eta bps/\log_2(8)$, lo que resulta en $R = \{1/6, 1/3, 1/2, 2/3, 5/6\}$. Como un ejemplo del rendimiento relativo de las constelaciones en estas tablas, se considera la Figura 13b que muestra una constelación PAM-8 optimizada para la capacidad de PD optimizada para la SNR = 9.00 dB, o 1.5 bps. A continuación examinamos el gráfico proporcionado en la Figura 8a y observamos que la brecha de la constelación optimizada a la capacidad gaussiana (CG) máxima es de aproximadamente 0.5 dB. A la misma eficiencia espectral, la brecha de la constelación PAM-8 tradicional es de aproximadamente 1.0 dB. La ventaja de la constelación optimizada es de 0.5 dB para la misma tasa (en este caso R=1/2). Esta ganancia puede obtenerse al cambiar sólo el asignador y el desasignador en el sistema de comunicación y dejar todos los otros bloques iguales.

Información similar se presenta en las Figuras 14abcd, y 15ab que proporcionan gráficos de lugar y tablas de diseño para constelaciones PAM-16 optimizadas para la capacidad de PD y la capacidad conjunta. Igualmente, las Figuras 16abcd, 17ab proporcionan gráficos de lugar y tablas de diseño para constelaciones PAM-32 optimizadas para la capacidad de PD y la capacidad conjunta.

Constelaciones PSK de capacidad optimizada

Las constelaciones de cambio de fase (PSK) tradicionales ya son bastante óptimas. Esto puede verse en el gráfico 180 que compara las brechas de la SNR de las constelaciones PSK tradicionales con las PSK de capacidad optimizada mostradas en la Figura 18 donde la brecha entre la capacidad de PD y la capacidad gaussiana se representa para las PSK-4,8,16,32 tradicionales y para las PSK-4,8,16,32 optimizadas para la capacidad de PD.

El gráfico de lugar de los puntos de la PSK-32 optimizada para la PD a través de la SNR se muestra en la Figura 19, que en realidad caracteriza a todas las PSK con una eficiencia espectral $\eta \leq 5$. Esto puede verse en la Figura 20. Debe señalarse que a una baja SNR (0.4 dB) el diseño óptimo de la PSK-32 es el mismo que la PSK-4 tradicional, a una SNR = 8.4 dB la PSK-32 óptima es la misma que la PSK-8 tradicional, a una SNR = 14.8 dB la PSK-32 óptima es la misma que la PSK-16 tradicional, y finalmente a una SNR mayor que 20.4 dB la PSK-32 optimizada es la misma que la PSK-32 tradicional. Existen SNR entre estos puntos discretos (por ejemplo, SNR = 2 y 15 dB) para las cuales la PSK-32 optimizada proporciona una capacidad de PD superior en comparación con las constelaciones PSK tradicionales.

Notamos ahora que el lugar de los puntos para la PSK-32 optimizada para PD en la Figura 19 junto con la brecha de la curva de capacidad gaussiana para la PSK-32 optimizada en la Figura 18 implica una metodología de diseño potencial. Específicamente, el diseñador podría lograr un rendimiento equivalente o mejor que el permitido por las PSK-4,8,16 tradicionales mediante el uso de sólo la PSK-32 optimizada junto con un único parámetro de sintonización que controle dónde deberían seleccionarse los puntos de la constelación en el lugar de la Figura 19. Dicho enfoque acoplaría un código de canal adaptativo de alta tasa que podría variar su tasa, por ejemplo, la tasa 4/5 para lograr una eficiencia espectral total (código más modulación PSK-32 optimizada) de 4 bits por símbolo, bajar hasta 1/5 para lograr una eficiencia espectral total de 1 bit por símbolo. Tal sistema de modulación y codificación adaptativo podría funcionar esencialmente en el flujo continuo óptimo representado por el contorno más a la derecha de la Figura 18.

Diseño de tasa adaptativa

En el ejemplo anterior, se describió el uso adaptativo espectral de la PSK-32. Técnicas similares a esta pueden aplicarse para otras constelaciones de capacidad optimizada a través del enlace entre un transmisor y un receptor. Por ejemplo, en el caso donde un sistema implementa la calidad de servicio, es posible indicar a un transmisor que aumente o disminuya la eficiencia espectral a demanda. En el contexto de la invención actual, una constelación de capacidad optimizada diseñada precisamente para la eficiencia espectral objetivo puede cargarse en el asignador de transmisión junto con una selección de tasa de código que cumple con el objetivo de tasa de usuario final. Cuando ocurre tal cambio de tasa de modulación/código, podría propagarse un mensaje al receptor de manera que el receptor, en anticipación al cambio, podría seleccionar una configuración de desasignador/descodificador para coincidir con la nueva configuración del lado de transmisión.

Por el contrario, el receptor podría implementar un mecanismo de control de par de tasa de código/constelación optimizada basado en la calidad del rendimiento. Tal enfoque incluiría cierta forma de medida de calidad del receptor. Esta podría ser la estimación del receptor de la SNR o la tasa de error de bits. Tomemos, por ejemplo, el caso donde la tasa de error de bits estaba por encima de cierto umbral aceptable. En este caso, a través de un canal de fondo, el receptor podría solicitar que el transmisor disminuya la eficiencia espectral del enlace al cambiar a un par de tasa de código/constelación optimizada de capacidad alternativa en los módulos del codificador y el asignador y luego indicar al receptor que intercambie en la pareja complementaria en los módulos del desasignador/descodificador.

Constelaciones QAM conformadas geométricamente

Las constelaciones de modulación de amplitud en cuadratura (QAM) pueden construirse mediante la ortogonalización de las constelaciones PAM en componentes en fase y cuadratura QAM. Las constelaciones construidas de esta manera pueden ser atractivas en muchas aplicaciones ya que tienen desasignadores de baja complejidad.

En la Figura 21 proporcionamos un ejemplo de una constelación de modulación de amplitud en cuadratura construida a partir de una constelación de modulación de amplitud de pulso. La modalidad ilustrada se construyó mediante el uso de una constelación PAM-8 optimizada para la capacidad de PD a la tasa de bits de usuario por dimensión de 1.5 bits (corresponde a una SNR de 9.0 dB) (ver la Figura 13b). Los pares de etiqueta y punto en esta constelación PAM-8 son {(000, -1.72), (001, -0.81), (010, 1.72), (011, -0.62), (100, 0.62), (101, 0.02), (110, 0.81), (111, -0.02)}. La examinación de la Figura 21 muestra que la construcción de la constelación QAM se logra al replicar un conjunto completo de puntos PAM-8 en la dimensión en cuadratura para cada uno de los 8 puntos PAM-8 en la dimensión en fase. El etiquetado se logra al asignar las etiquetas PAM-8 al intervalo LSB en la dimensión en fase y al intervalo MSB en la dimensión en cuadratura. El producto exterior 8x8 resultante forma una QAM-64 altamente estructurada para la cual pueden construirse desasignadores de muy baja complejidad. Debido a la ortogonalidad de los componentes en fase y en cuadratura, las características de capacidad de la constelación QAM-64 resultante son idénticas a las de la constelación PAM-8 en una base por dimensión.

Optimización de la constelación N-dimensional

En lugar de diseñar constelaciones en 1-D (PAM, por ejemplo) y luego extender a 2-D (QAM), es posible tomar ventaja directa en la etapa de optimización del grado de libertad adicional presentado por una dimensión espacial extra. En general, es posible diseñar constelaciones N-dimensionales y las etiquetas asociadas. La complejidad de la etapa de optimización crece exponencialmente en el número de dimensiones, al igual que la complejidad del desasignador del receptor resultante. Tales construcciones constituyen modalidades de la invención y simplemente requieren más “tiempo de ejecución” para producir.

Capacidad. Constelaciones optimizadas para los canales de desvanecimiento

Procesos similares a los descritos anteriormente pueden usarse para diseñar constelaciones de capacidad optimizada para los canales de desvanecimiento de acuerdo con las modalidades de la invención. Los procesos son esencialmente los mismos, con la excepción de que la manera en que se calcula la capacidad se modifica para tener en cuenta el canal de desvanecimiento. Un canal de desvanecimiento puede describirse mediante el uso de la siguiente ecuación:

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

donde X es la señal transmitida, N es una señal de ruido gaussiano blanco aditivo y a(t) es la distribución de desvanecimiento, que es una función del tiempo.

En el caso de un canal de desvanecimiento, la SNR instantánea en el receptor cambia de acuerdo con una distribución de desvanecimiento. La distribución de desvanecimiento es Rayleigh y tiene la propiedad de que la SNR promedio del sistema permanece igual que en el caso del canal AWGN, E[X2]/E[N2]. Por lo tanto, la capacidad del canal de desvanecimiento puede calcularse al tomar la expectativa de la capacidad AWGN, en una SNR promedio dada, sobre la distribución de desvanecimiento de Rayleigh de a que impulsa la distribución de la SNR instantánea.

Muchos canales de desvanecimiento siguen una distribución de Rayleigh. Las Figuras 22a - 24b son gráficos de lugar de constelaciones PAM-4, 8 y 16 que se han optimizado para la capacidad de PD en un canal de desvanecimiento de Rayleigh. Se proporcionan gráficos de lugar versus la tasa de bits de usuario por dimensión y versus la SNR. Procesos similares pueden usarse para obtener constelaciones de capacidad optimizada que se optimizan mediante el uso de otras medidas de capacidad, tal como la capacidad conjunta, y/o mediante el uso de diferentes esquemas de modulación.

Reivindicaciones

1. Un sistema de comunicación digital (10), que comprende:
un transmisor (14) configurado para transmitir señales a través de un canal de comunicación (15);
en donde el transmisor (14), comprende:
un codificador (20) configurado para recibir bits de usuario y bits codificados de salida a una tasa de bits codificada de salida expandida;
un asignador (22) configurado para asignar los bits codificados a símbolos en una constelación de símbolos modulada; y
un modulador (24) configurado para generar una señal modulada para la transmisión a través del canal de comunicación mediante el uso de símbolos generados por el asignador.
en donde el transmisor se caracteriza porque la constelación de símbolos es una constelación de símbolos separada desigualmente obtenida mediante la optimización de las ubicaciones y el etiquetado de los puntos de constelación para maximizar la capacidad de descodificación paralela en una relación de señal a ruido dada, de manera que la constelación de símbolos proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza la distancia mínima, d_{min}, entre los puntos de constelación de la constelación de símbolos.

2. El sistema de comunicación (10) de acuerdo con la reivindicación 1 que comprende además:
un receptor (16) configurado para recibir señales a través de un canal de comunicación (15),
en donde el receptor comprende:
un demodulador (30) configurado para demodular una señal recibida a través del canal de comunicación;
un desasignador (32) configurado para estimar las probabilidades a partir de la señal demodulada mediante el uso de la constelación de símbolos; y
un descodificador (34) que se configura para estimar los bits descodificados a partir de las probabilidades generadas por el desasignador (32).

3. Un sistema de comunicación digital (10), que comprende:
un receptor (16) configurado para recibir señales a través de un canal de comunicación (15); el receptor que comprende:
un demodulador (30) configurado para demodular una señal recibida a través del canal de comunicación;
un desasignador (32) configurado para estimar las probabilidades a partir de la señal demodulada mediante el uso de una constelación de símbolos;
un descodificador (34) que se configura para estimar los bits descodificados a partir de las probabilidades generadas por el desasignador (32); y
en donde el receptor se caracteriza porque la constelación de símbolos es una constelación de símbolos separada desigualmente obtenida mediante la optimización de las ubicaciones y el etiquetado de los puntos de constelación para maximizar la capacidad de descodificación paralela en una relación de señal a ruido dada, de manera que la constelación de símbolos proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza la distancia mínima, dmin, entre los puntos de constelación de la constelación de símbolos.

4. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde la constelación de símbolos separada desigualmente se optimiza sujeta a restricciones de optimización que incluyen cualquiera de:
límites inferior y superior en la ubicación de los puntos;
energía pico a promedio de la constelación resultante; y
media en la constelación resultante.

5. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde el canal es un canal de ruido gaussiano blanco aditivo, AWGN.

6. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde el canal es un canal de desvanecimiento.

7. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde los puntos de constelación de la constelación de símbolos separada desigualmente son variables en dependencia de la tasa de usuario.

8. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde los puntos de constelación se obtuvieron al optimizar las ubicaciones y el etiquetado de los puntos de constelación para maximizar la medida de capacidad de descodificación paralela en una relación de señal a ruido dada, de manera que los puntos de constelación obtenidos tienen una capacidad mayor que la capacidad de una constelación que maximiza la distancia mínima, d_{min}, entre los puntos de constelación de la constelación de símbolos.

9. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde los puntos de constelación proporcionan una brecha SNR a la capacidad gaussiana en una capacidad dada que es un margen

predeterminado menor que la brecha de la relación de señal a ruido a la capacidad gaussiana de una constelación de señales que maximiza la distancia mínima, d_{min}, entre los puntos de constelación de la constelación de símbolos.

10. El sistema de comunicación (10) de acuerdo con la reivindicación 9, en donde el margen predeterminado es de 0.5 dB.

11. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde la constelación de símbolos comprende un anillo de puntos de constelación que se separan desigualmente alrededor del anillo.

12. El sistema de comunicación (10) de acuerdo con la reivindicación 11, en donde la constelación de símbolos es una constelación PSK no uniforme.

13. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, configurado para utilizar un código LDPC.

14. El sistema de comunicación (10) de acuerdo con cualquiera de las reivindicaciones 2 o 3, en donde no hay iteraciones de propagación de creencias entre el descodificador y el desasignador de constelaciones.

15. El sistema de comunicación (10) de acuerdo con cualquier reivindicación anterior, en donde la constelación se selecciona junto con una tasa de código para cumplir con un requisito de energía de transmisión máxima.

16. Un método realizado en el sistema de comunicación digital (10), el método que comprende:
un transmisor (14) que transmite señales a través de un canal de comunicación (15) mediante;
un codificador (20) que recibe los bits de usuario y envía los bits codificados a una tasa de bits codificada de salida expandida;
un asignador (22) que asigna los bits codificados a símbolos en una constelación de símbolos modulada; y
un modulador (24) que genera una señal modulada para la transmisión a través del canal de comunicación mediante el uso de los símbolos generados por el asignador,
en donde el método se caracteriza porque la constelación de símbolos es una constelación de símbolos separada desigualmente obtenida mediante la optimización de las ubicaciones y el etiquetado de los puntos de constelación para maximizar la capacidad de descodificación paralela en una relación de señal a ruido dada, de manera que la constelación de símbolos proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza la distancia mínima, d_{min}, entre los puntos de constelación de la constelación de símbolos.

17. Un método realizado en un sistema de comunicación digital (10), el método que comprende:
un receptor (16) que recibe señales a través de un canal de comunicación (15) mediante:
un demodulador (30) que demodula una señal recibida a través del canal de comunicación;
un desasignador (32) que estima las probabilidades a partir de la señal demodulada mediante el uso de una constelación de símbolos; y
un descodificador (34) que estima los bits descodificados a partir de las probabilidades generadas por el desasignador (32),
en donde el método se caracteriza porque la constelación de símbolos es una constelación de símbolos separada desigualmente obtenida mediante la optimización de las ubicaciones y el etiquetado de los puntos de constelación para maximizar la capacidad de descodificación paralela en una relación de señal a ruido dada, de manera que la constelación de símbolos proporciona una capacidad dada en una relación de señal a ruido reducida en comparación con una constelación de señales que maximiza la distancia mínima, d_{min}, entre los puntos de constelación de la constelación de símbolos.

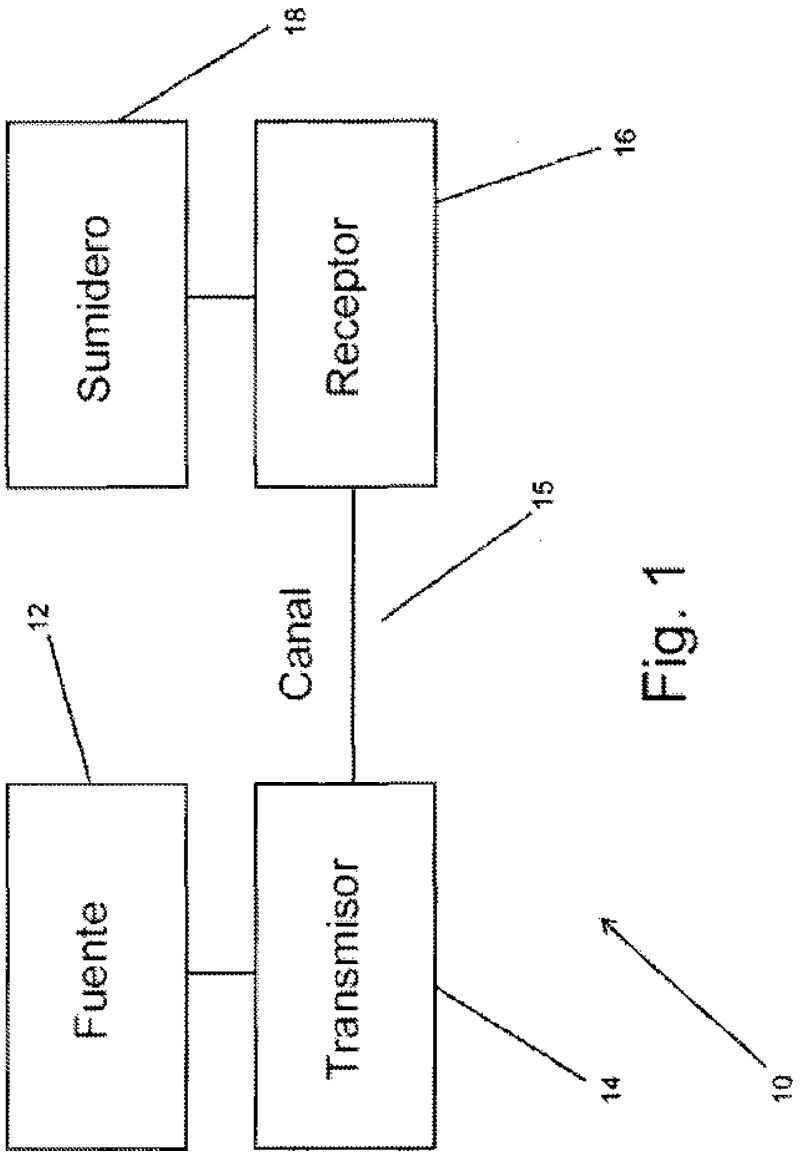


Fig. 1

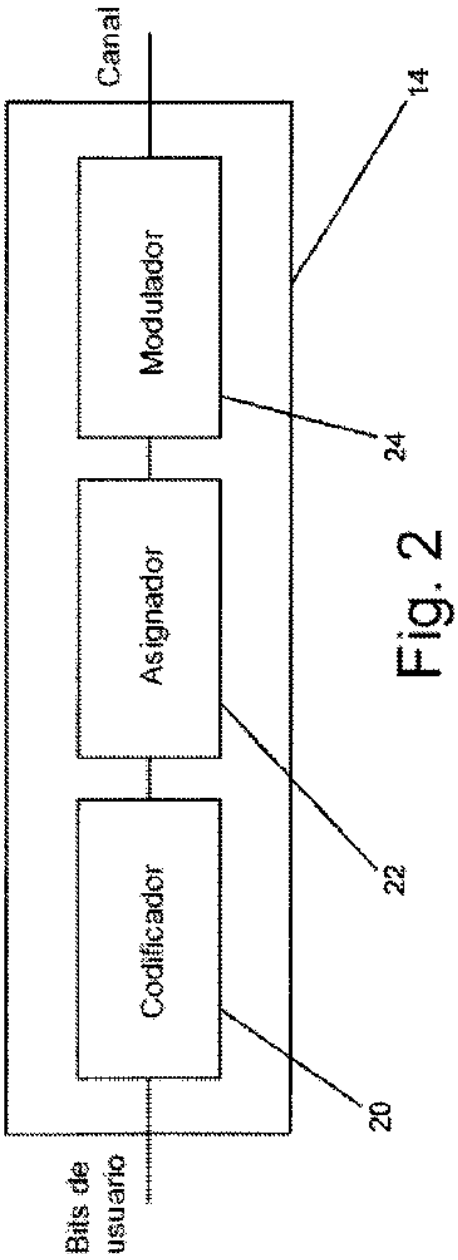

Bits de usuario
Codificador
Asignador
Modulador
Canal
20
22
24
14

Fig. 2

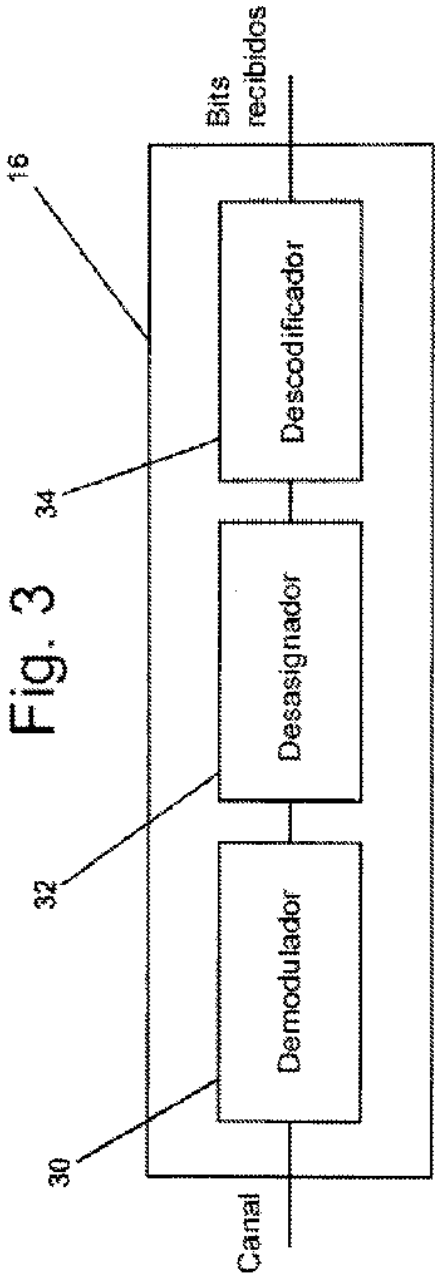
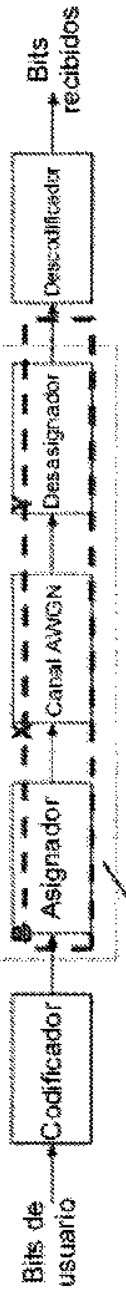

Canal
Demodulador
Desasignador
Descodificador
Bits recibidos
30
32
34
16

Fig. 3



40

Fig. 4a



42

Fig. 4b

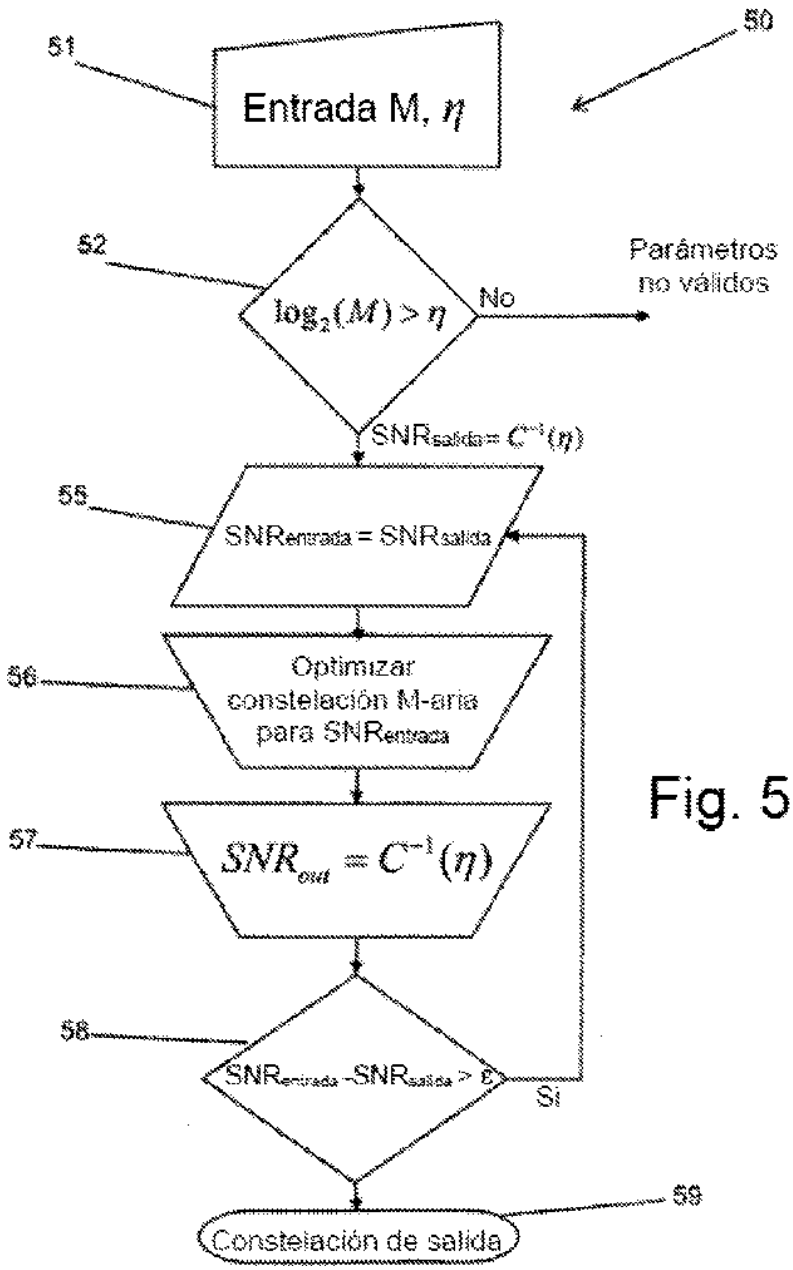

50
51
Entrada M, η
52
log2(M) > η
No
Parámetros no válidos
SNRsalida = C⁻¹(η)
55
SNRentrada = SNRsalida
56
Optimizar constelación M-aria para SNRentrada
57
SNRout = C⁻¹(η)
58
SNRentrada -SNRsalida > ε
Si
59
Constelación de salida

Fig. 5

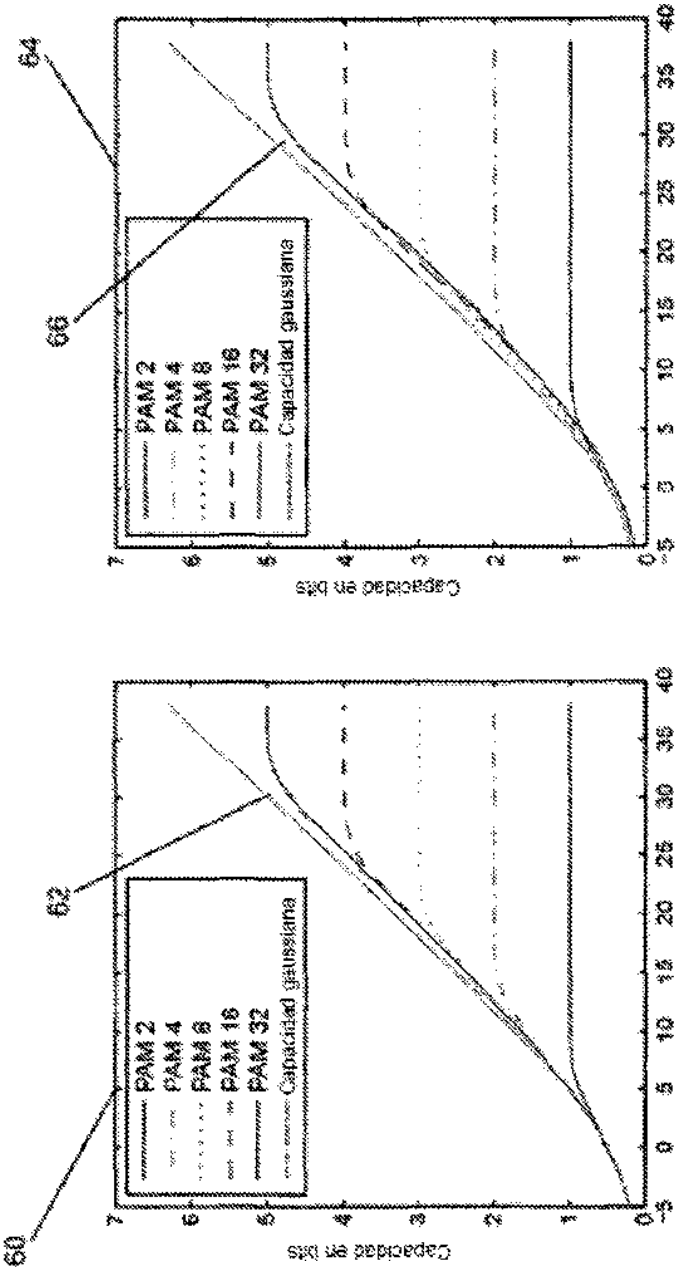
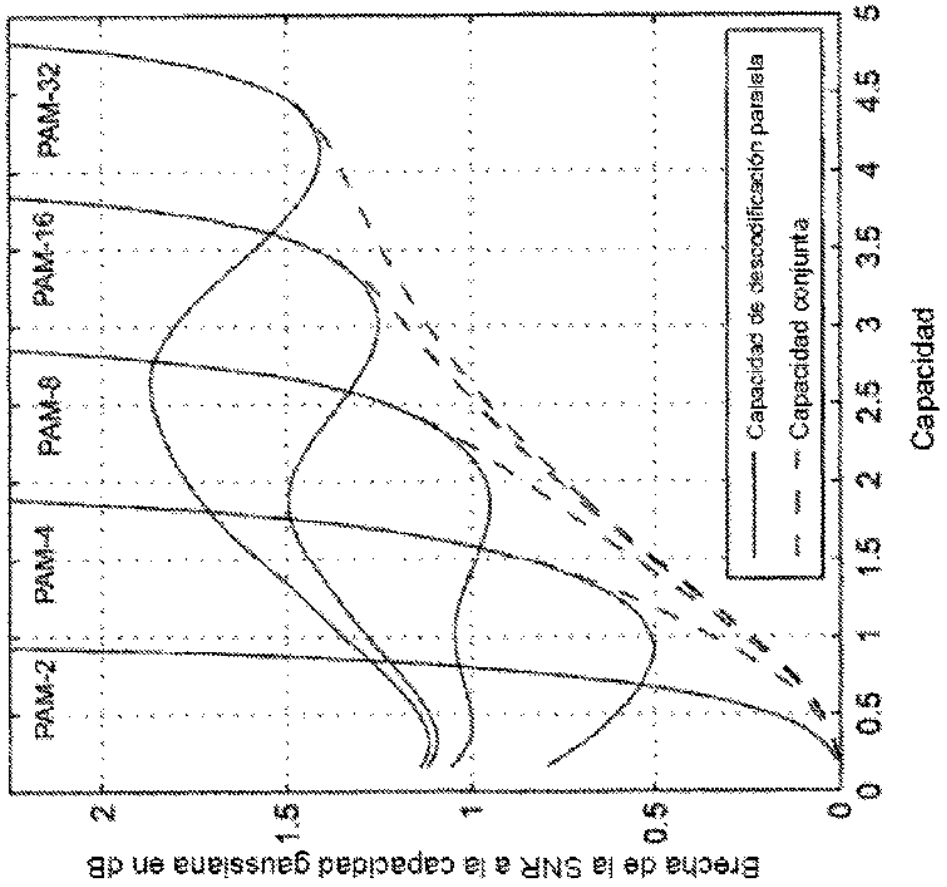

60
62
64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Capacidad gaussiana
Capacidad en bits
SNR en dB
(a)
(b)

Fig. 6a

Fig. 6b

Fig. 7


PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
Capacidad de descodificación paralela
Capacidad conjunta
Brecha de la SNR a la capacidad gaussiana en dB
Capacidad
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

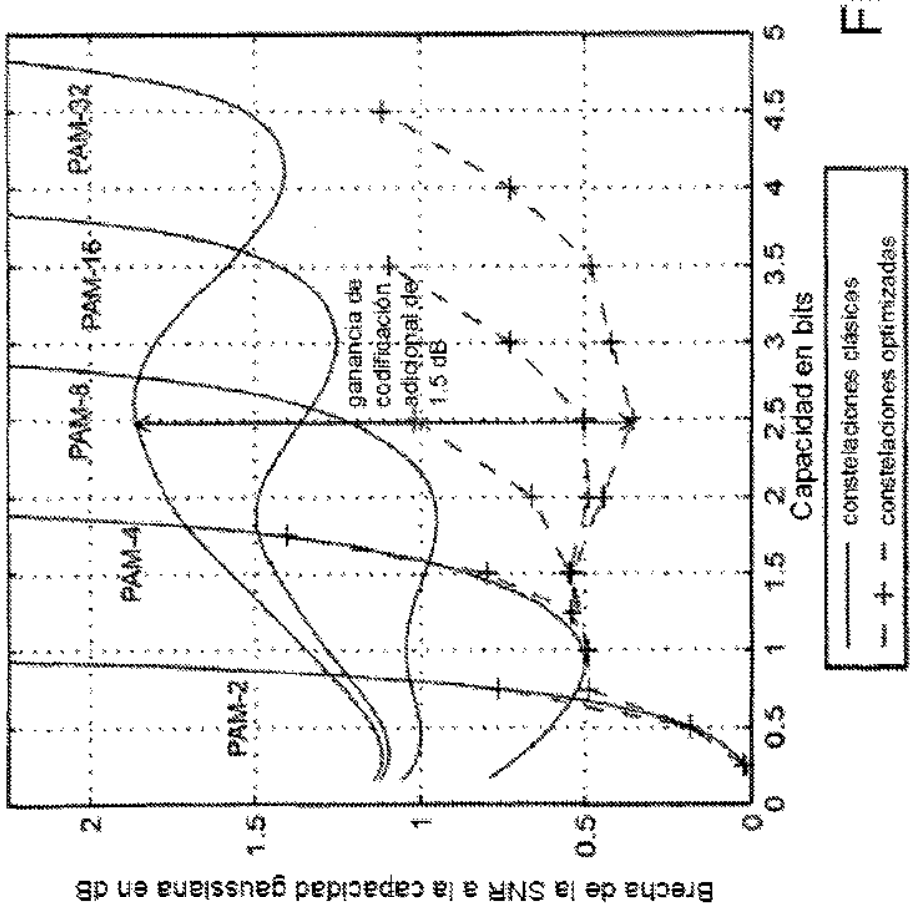

PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
ganancia de codificación adicional de 1.5 dB
Capacidad en bits
Brecha de la SNR a la capacidad gaussiana en dB
constelaciones clásicas
constelaciones optimizadas

Fig. 8a

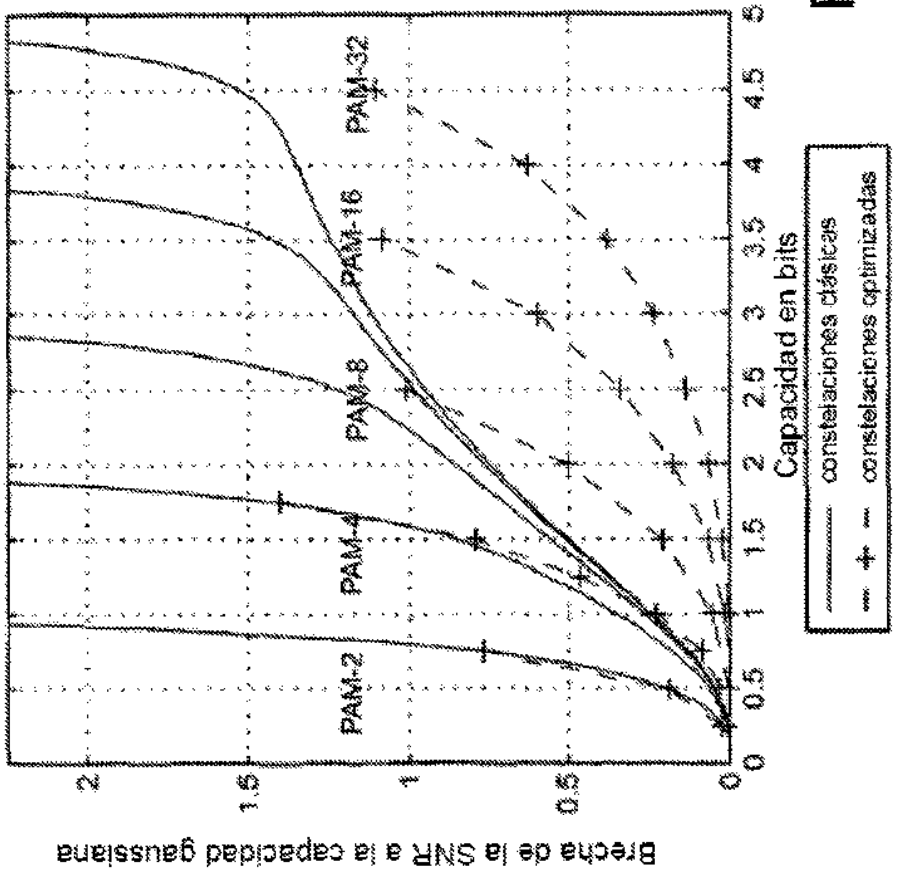

Brecha de la SNR a la capacidad gaussiana
Capacidad en bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
constelaciones clásicas
constelaciones optimizadas

Fig. 8b

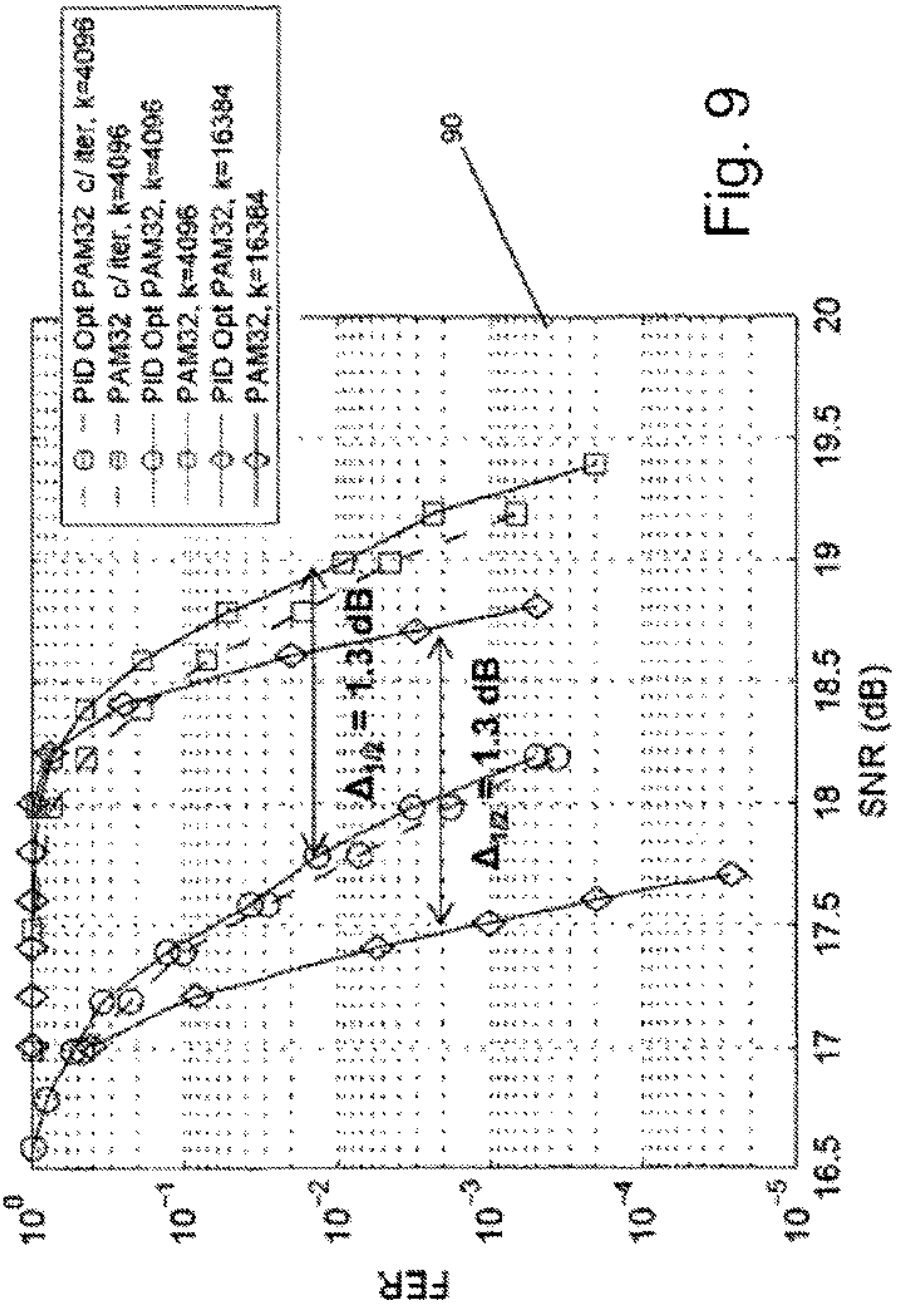

PID Opt PAM32 c/ iter, k=4096
PAM32 c/ iter, k=4096
PID Opt PAM32, k=4096
PAM32, k=4096
PID Opt PAM32, k=16384
PAM32, k=16384
90
$\Delta_{1/2}$ = 1.3 dB
$\Delta_{1/2}$ = 1.3 dB
FER
SNR (dB)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20

Fig. 9

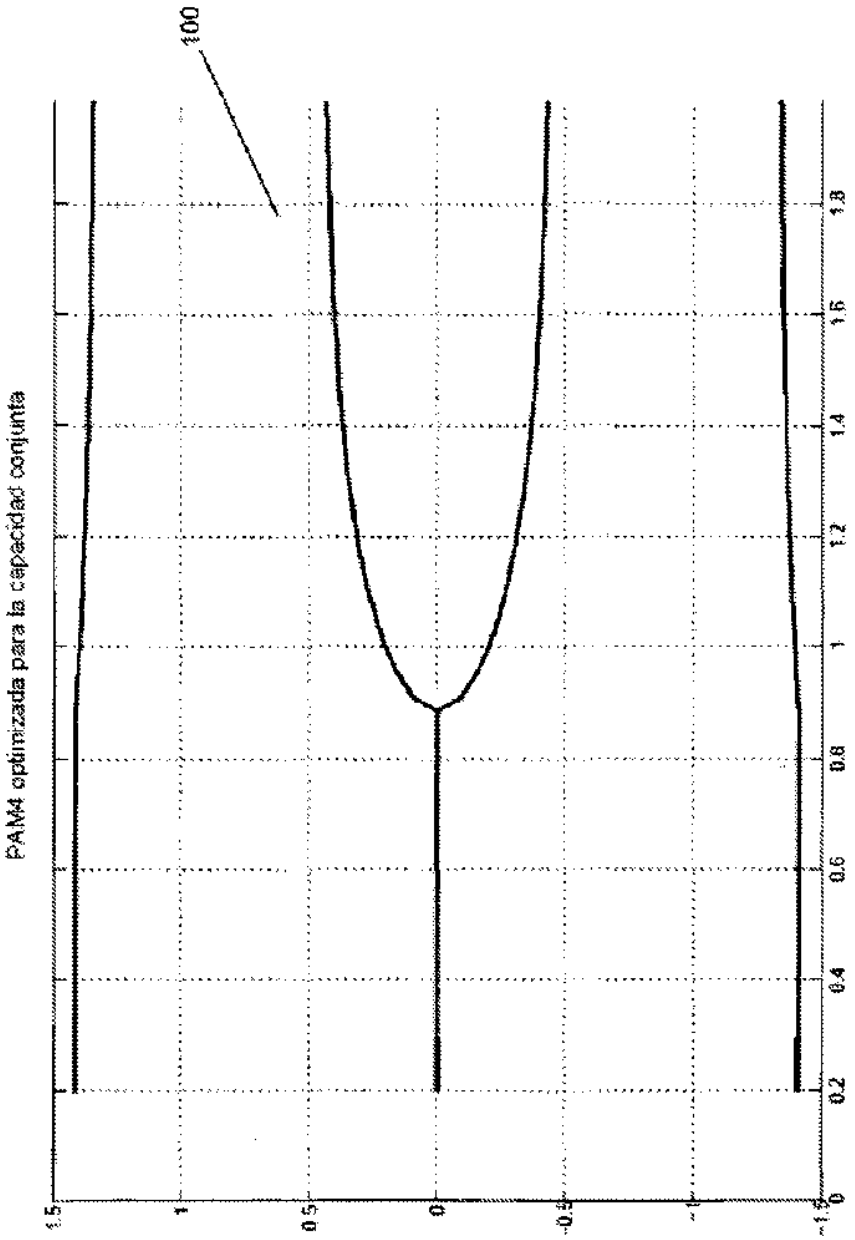

100
PAM4 optimizada para la capacidad conjunta
Capacidad en bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

Fig. 10a

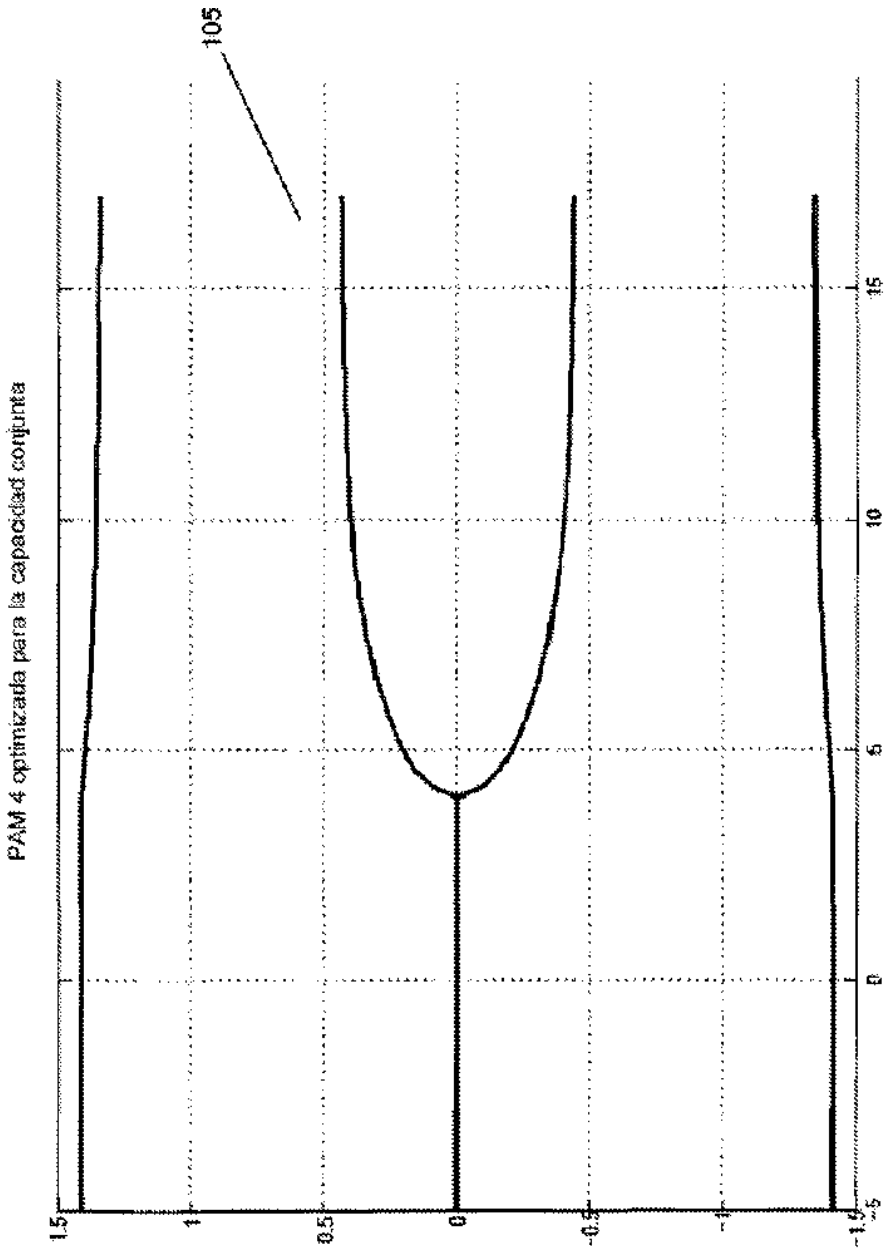

105
PAM 4 optimizada para la capacidad conjunta
SNR en dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

Fig. 10b

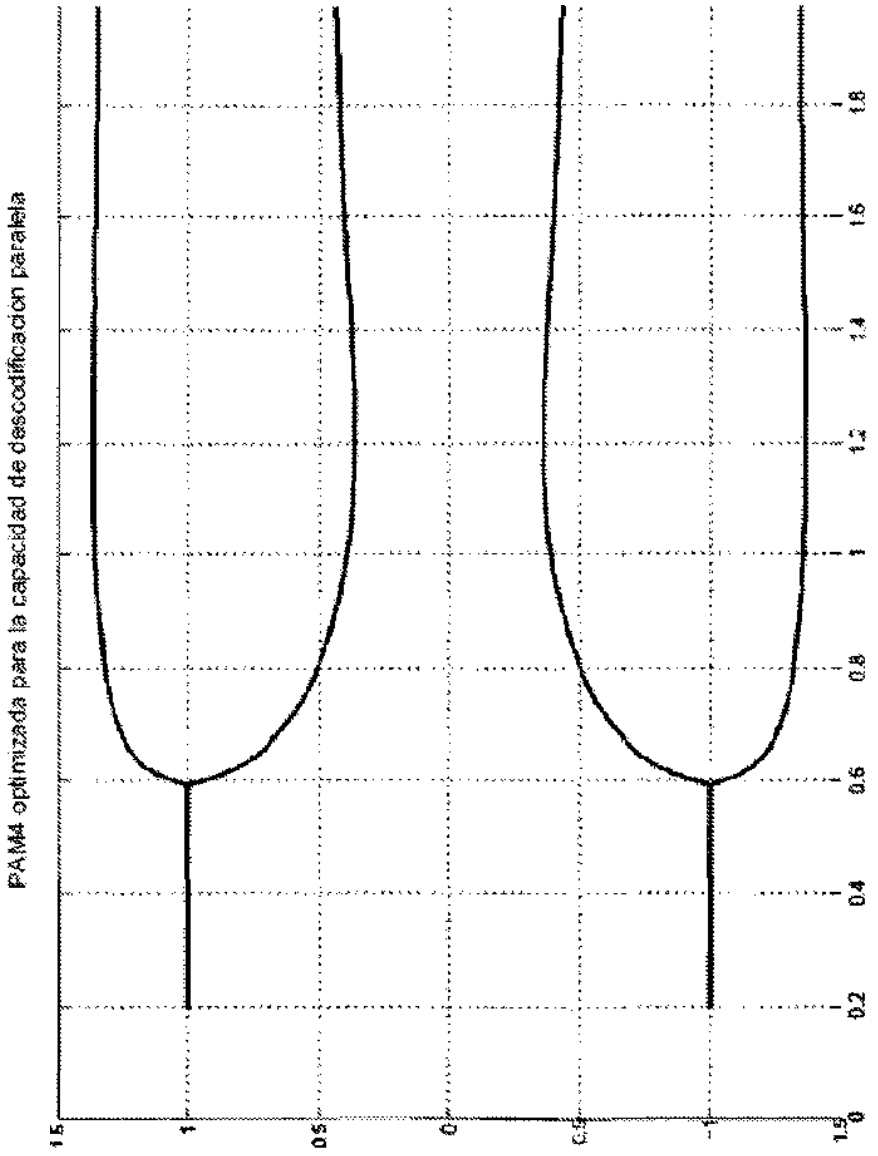

PAM4 optimizada para la capacidad de descodificación paralela
Capacidad en bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

Fig. 10c

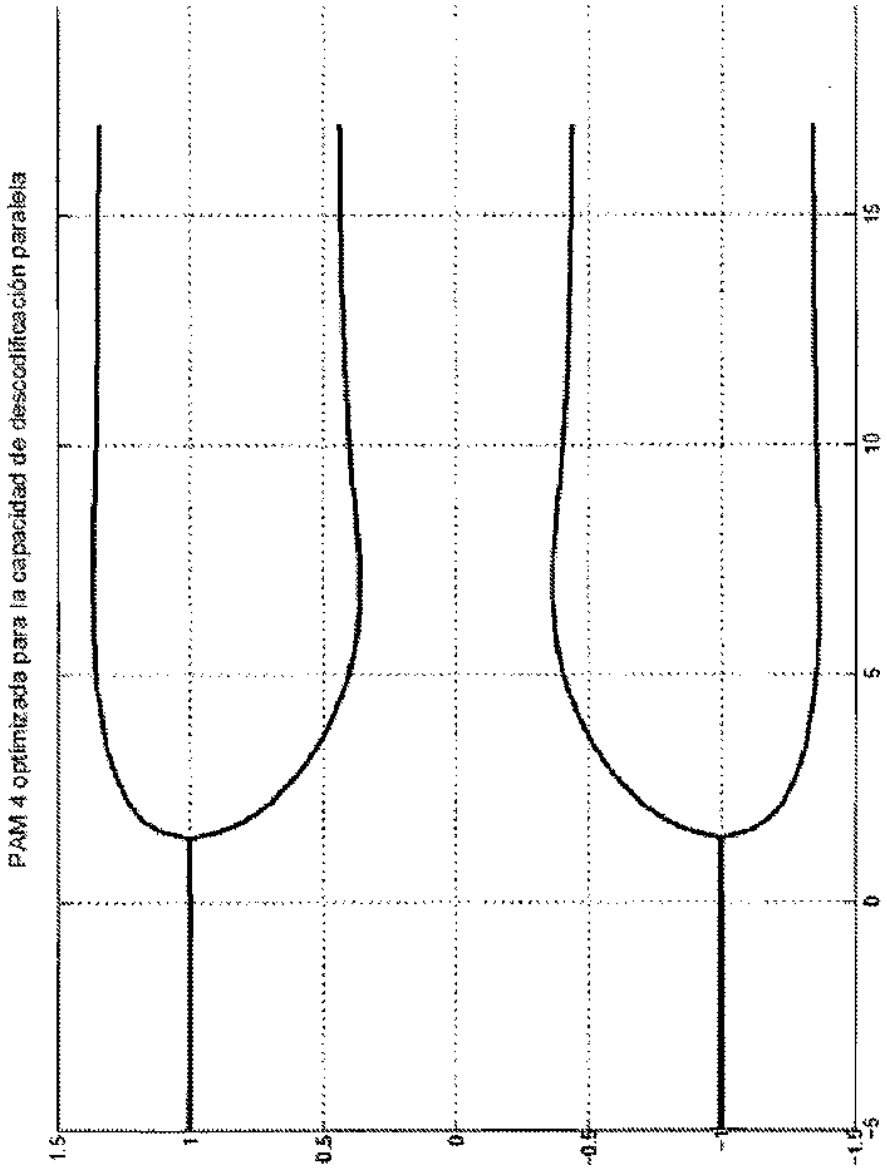

PAM 4 optimizada para la capacidad de descodificación paralela
SNR en dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

Fig. 10d

Constelaciones PAM-4 optimizadas para la capacidad conjunta a diferentes tasas

(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

Constelaciones PAM-4 optimizadas para la capacidad de descodificación paralela a diferentes tasas

(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

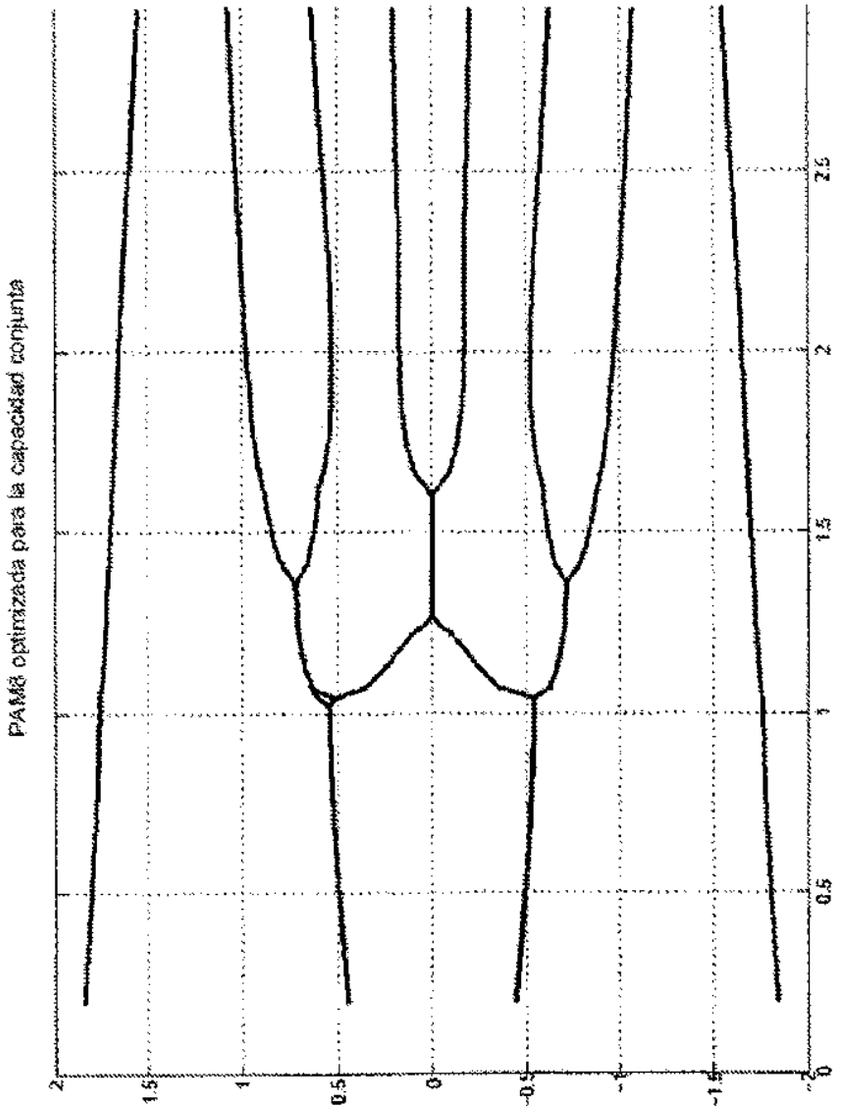

PAM8 optimizada para la capacidad conjunta
Capacidad en bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

Fig. 12a

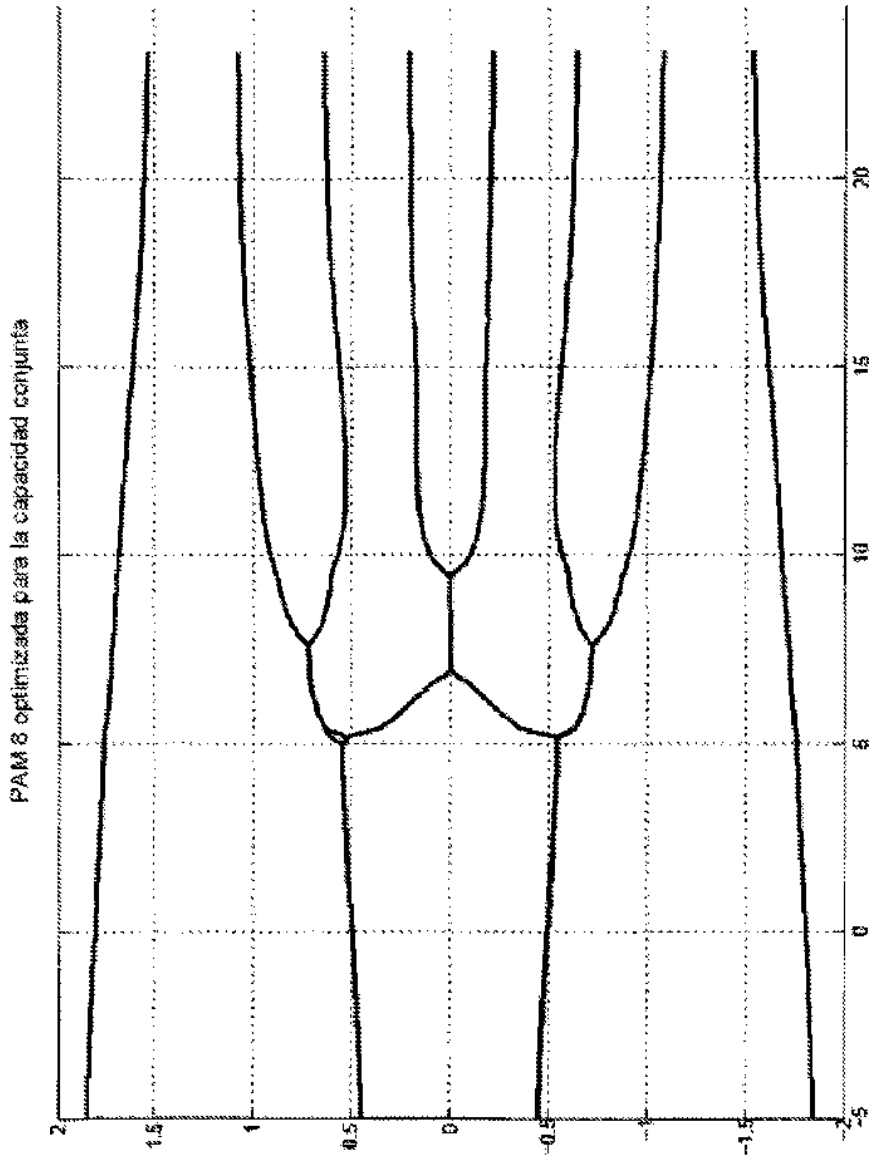

PAM 8 optimizada para la capacidad conjunta
SNR en dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

Fig. 12b

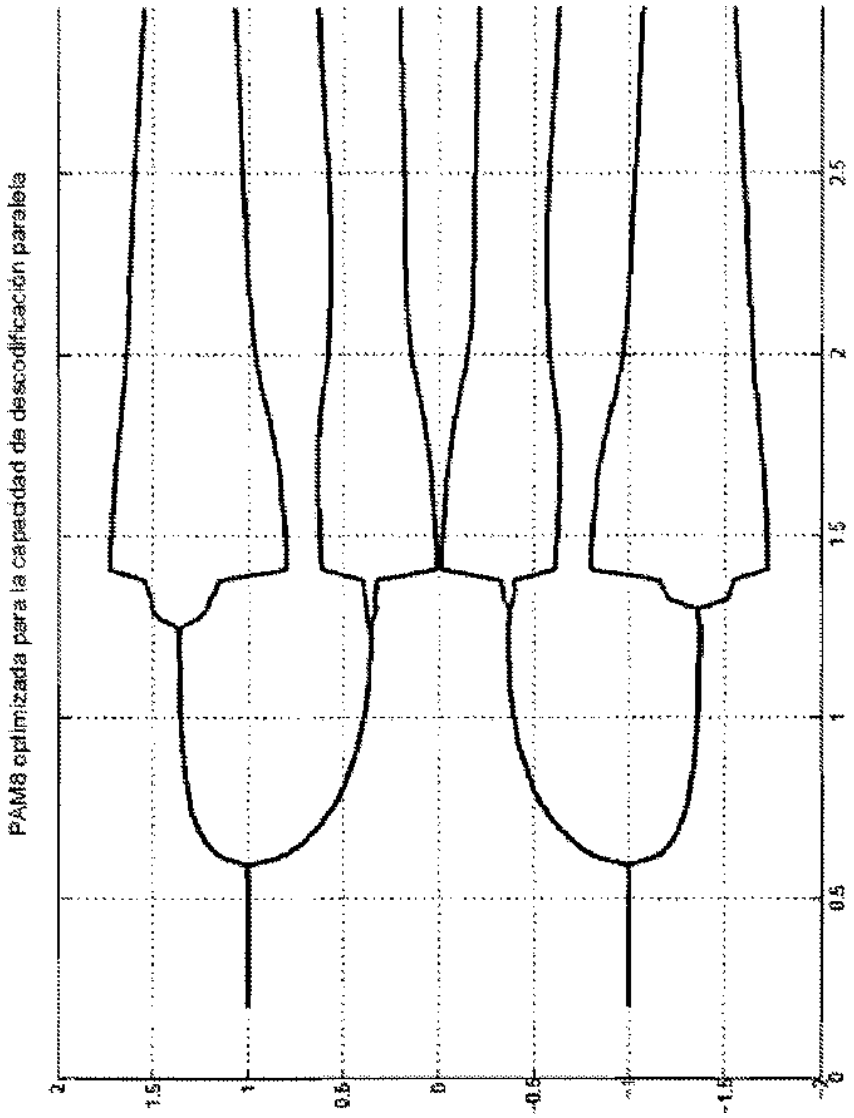

PAM8 optimizada para la capacidad de descodificación paralela
Capacidad en bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

Fig. 12c

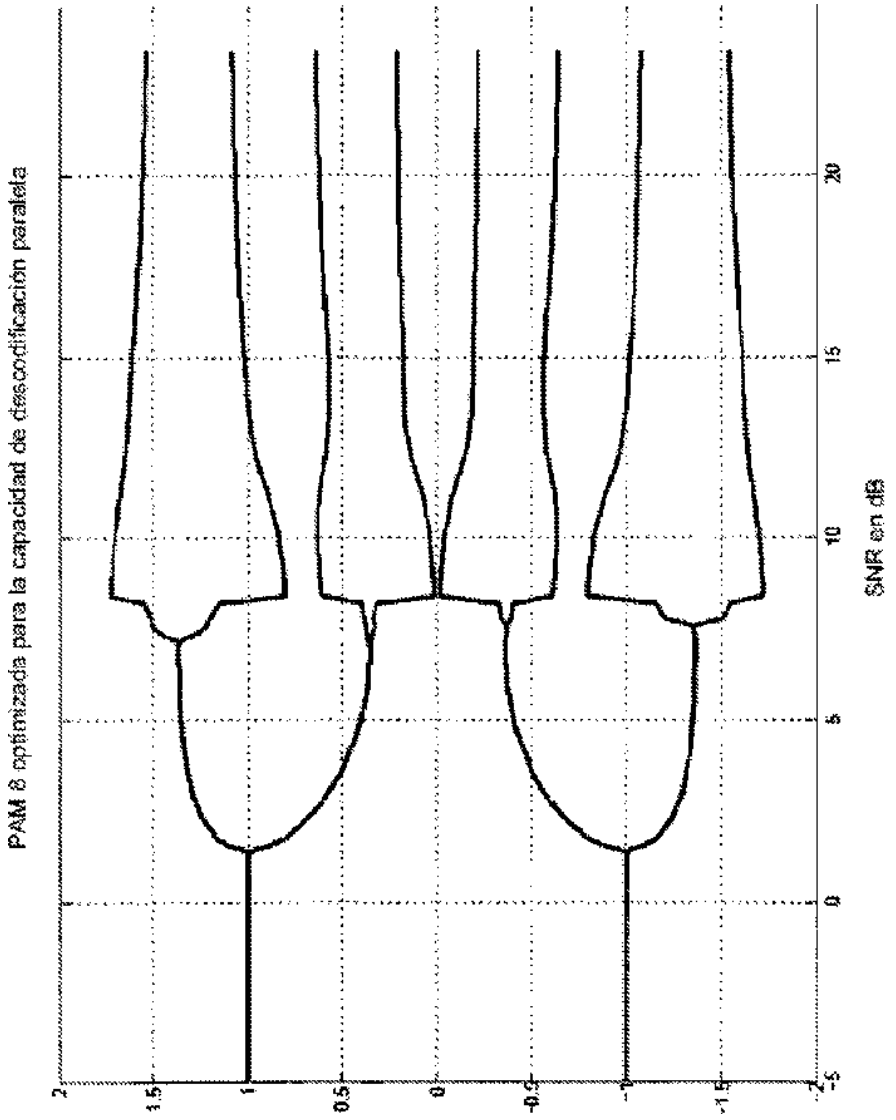

PAM 8 optimizada para la capacidad de descodificación paralela
SNR en dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

Fig. 12d

Constelaciones PAM-8 optimizadas para la capacidad conjunta a diferentes tasas

(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

Constelaciones PAM-8 optimizadas para la capacidad de descodificación paralela a diferentes tasas

(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

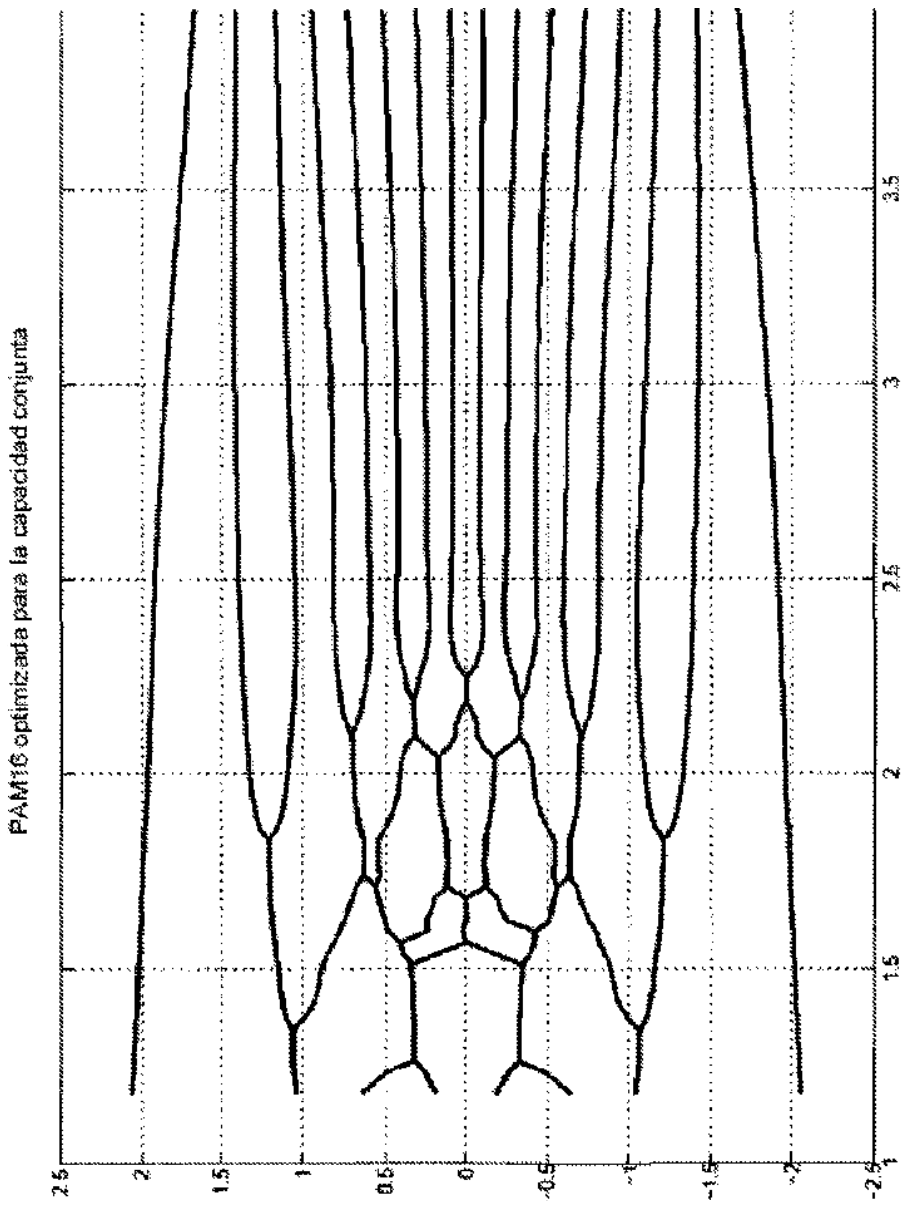

PAM16 optimizada para la capacidad conjunta
Capacidad en bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

Fig. 14a

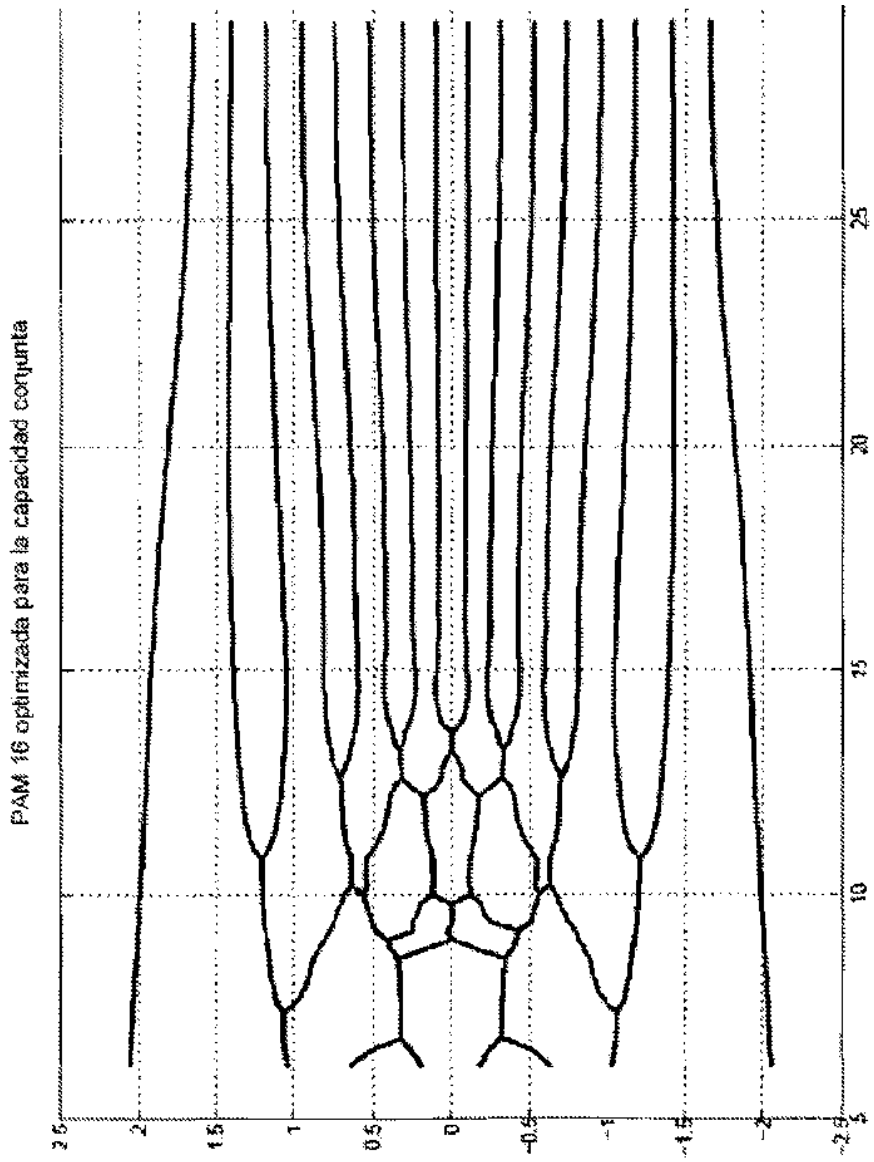

PAM 16 optimizada para la capacidad conjunta
SNR en dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

Fig. 14b

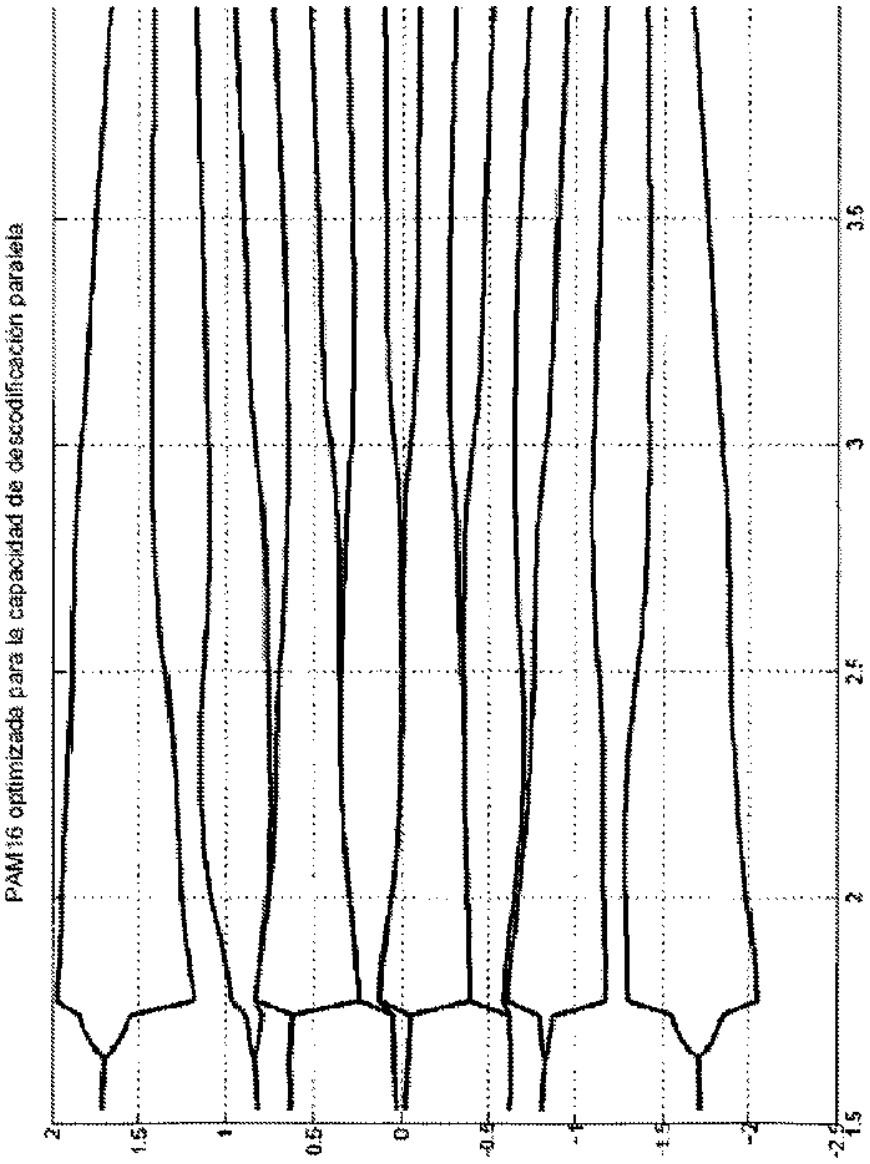

PAM16 optimizada para la capacidad de descodificación paralela
Capacidad en bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5

Fig. 14c

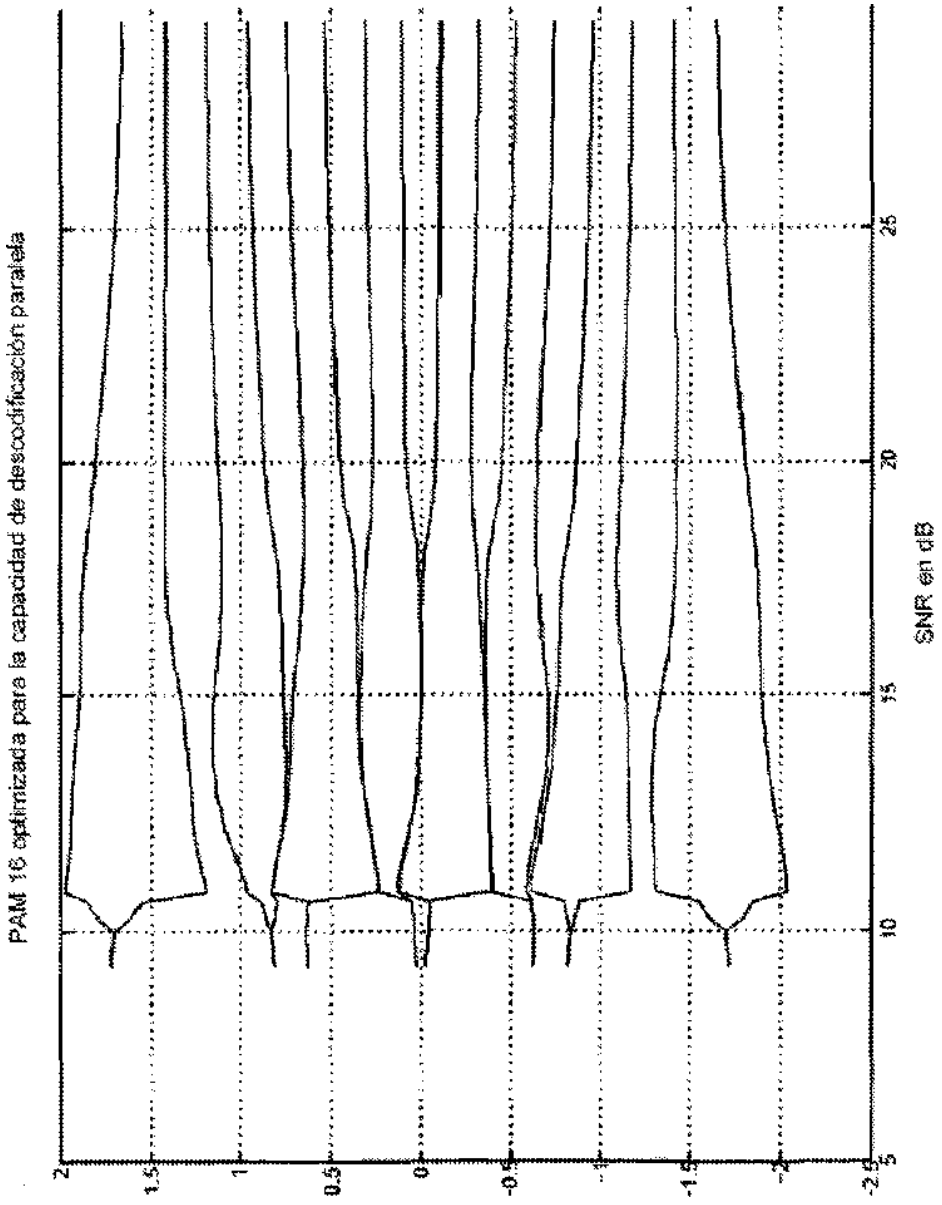

PAM 16 optimizada para la capacidad de descodificación paralela
SNR en dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25

Fig. 14d

Constelaciones PAM-16 optimizadas para la capacidad conjunta a diferentes tasas

(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

Constelaciones PAM-16 optimizadas para la capacidad de descodificación paralela a diferentes tasas

(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

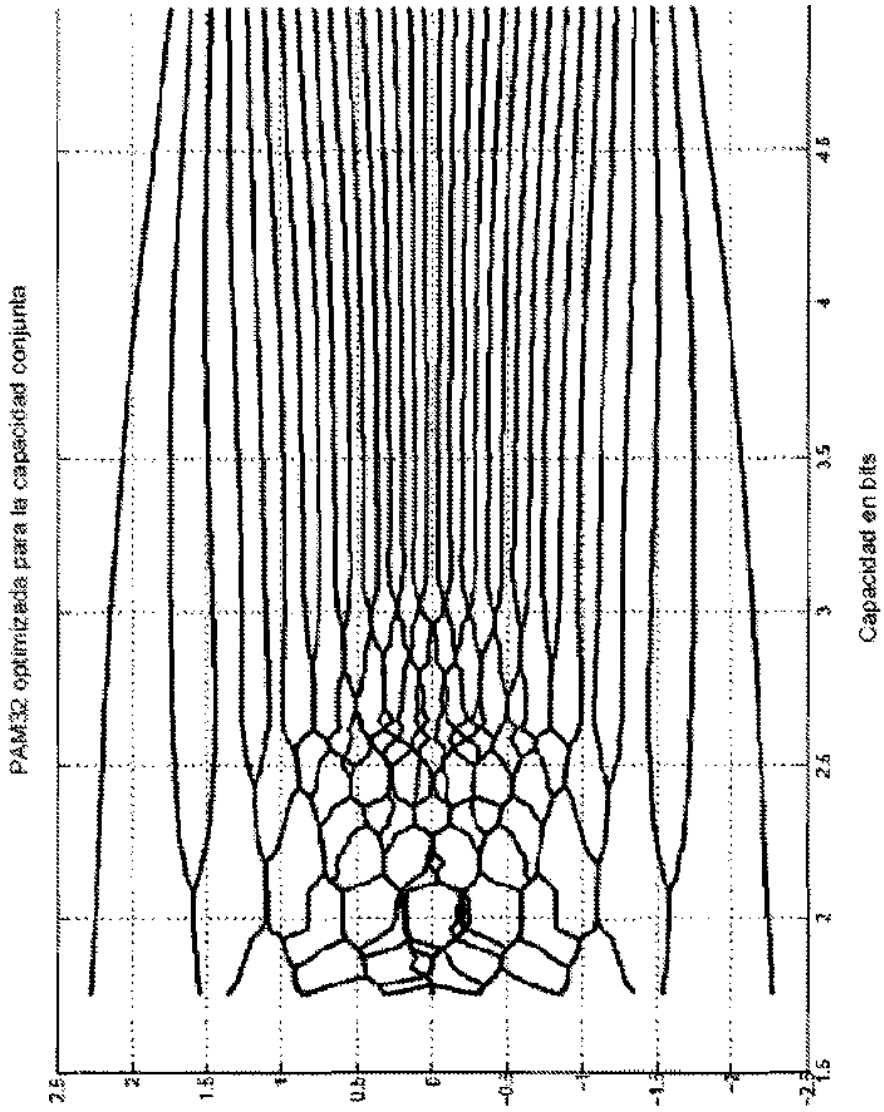

PAM32 optimizada para la capacidad conjunta
Capacidad en bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

Fig. 16a

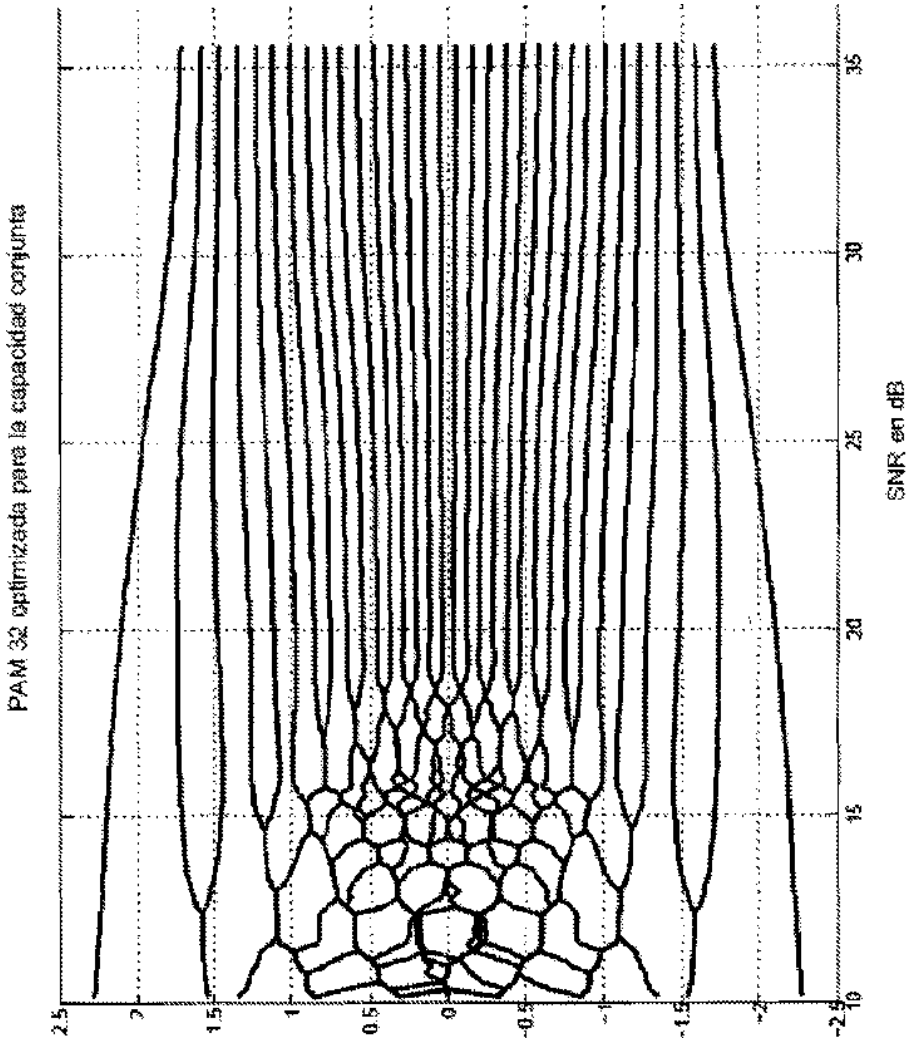

PAM 32 optimizada para la capacidad conjunta
SNR en dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

Fig. 16b

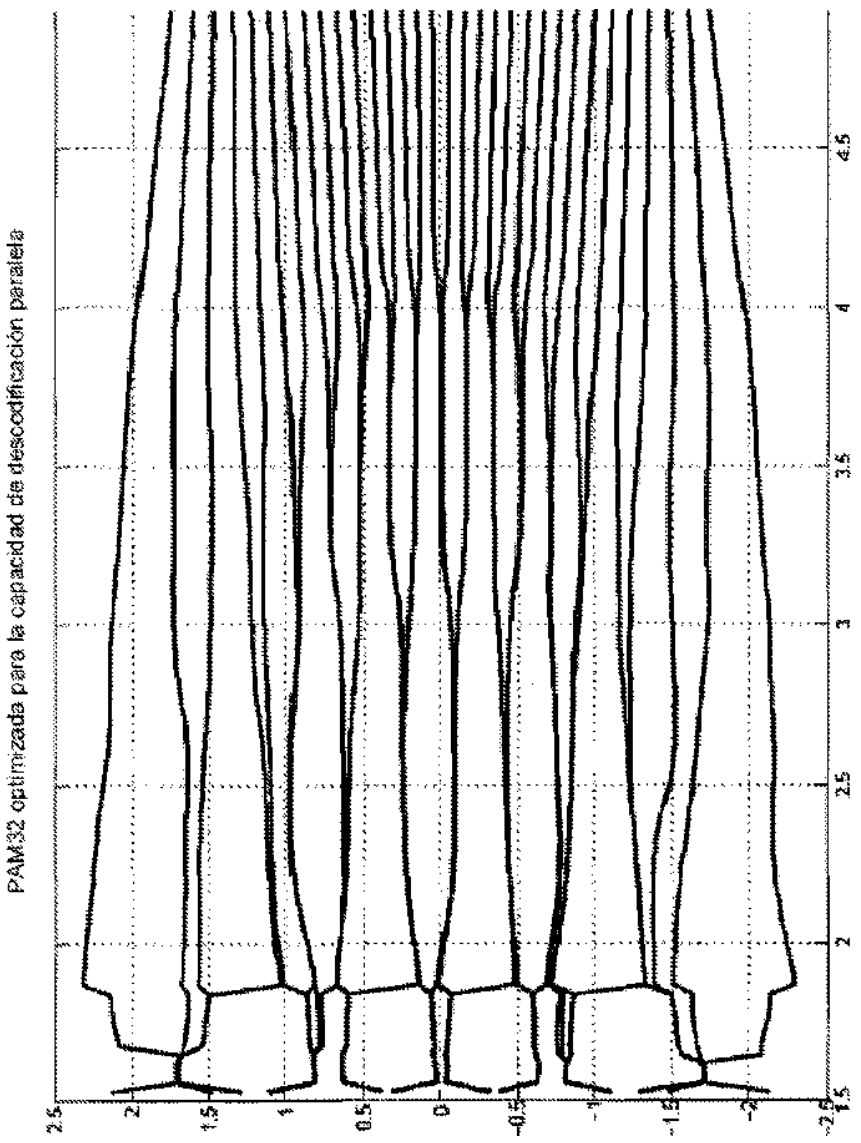

PAM32 optimizada para la capacidad de descodificación paralela
Capacidad en bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

Fig. 16c

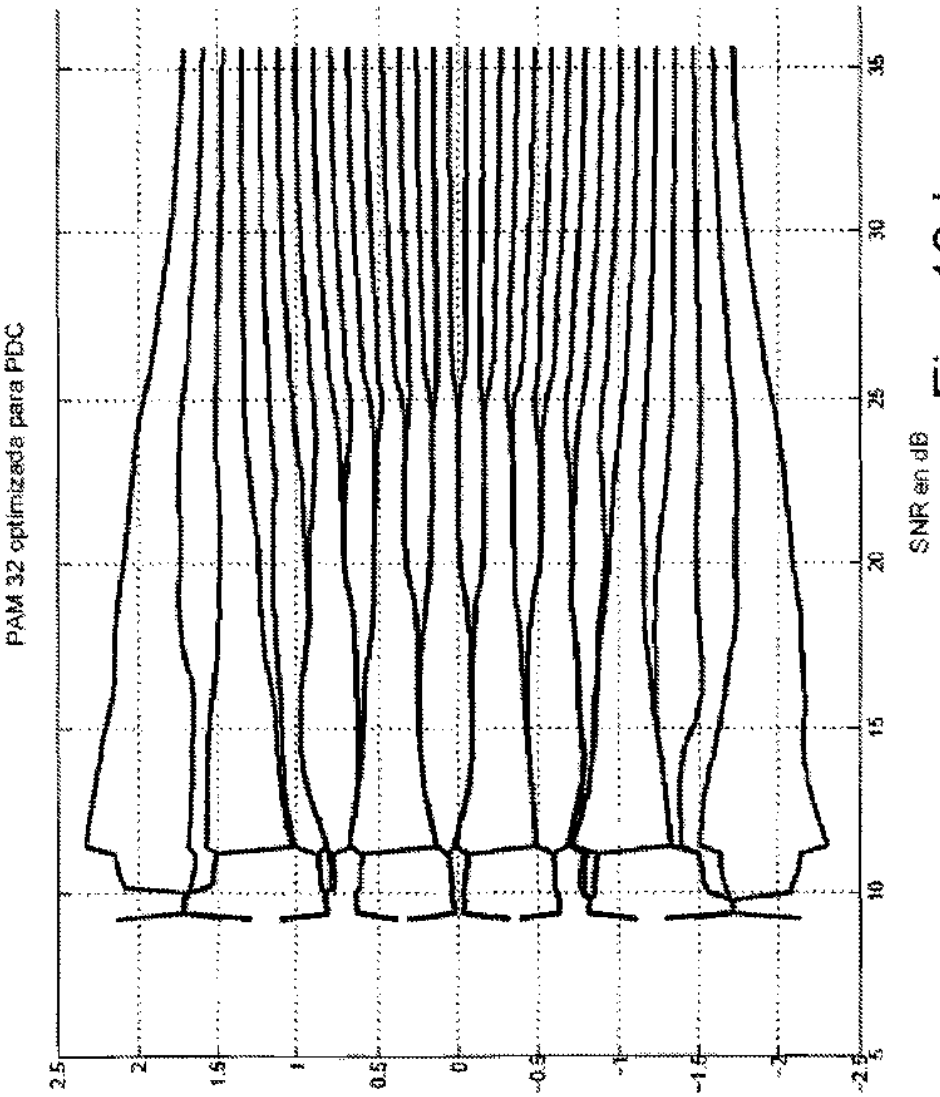

PAM 32 optimizada para PDC
SNR en dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

Fig. 16d

Constelaciones PAM-32 optimizadas para la capacidad conjunta a diferentes tasas

(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

Constelaciones PAM-32 optimizadas para la capacidad de descodificación paralela a diferentes tasas

(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

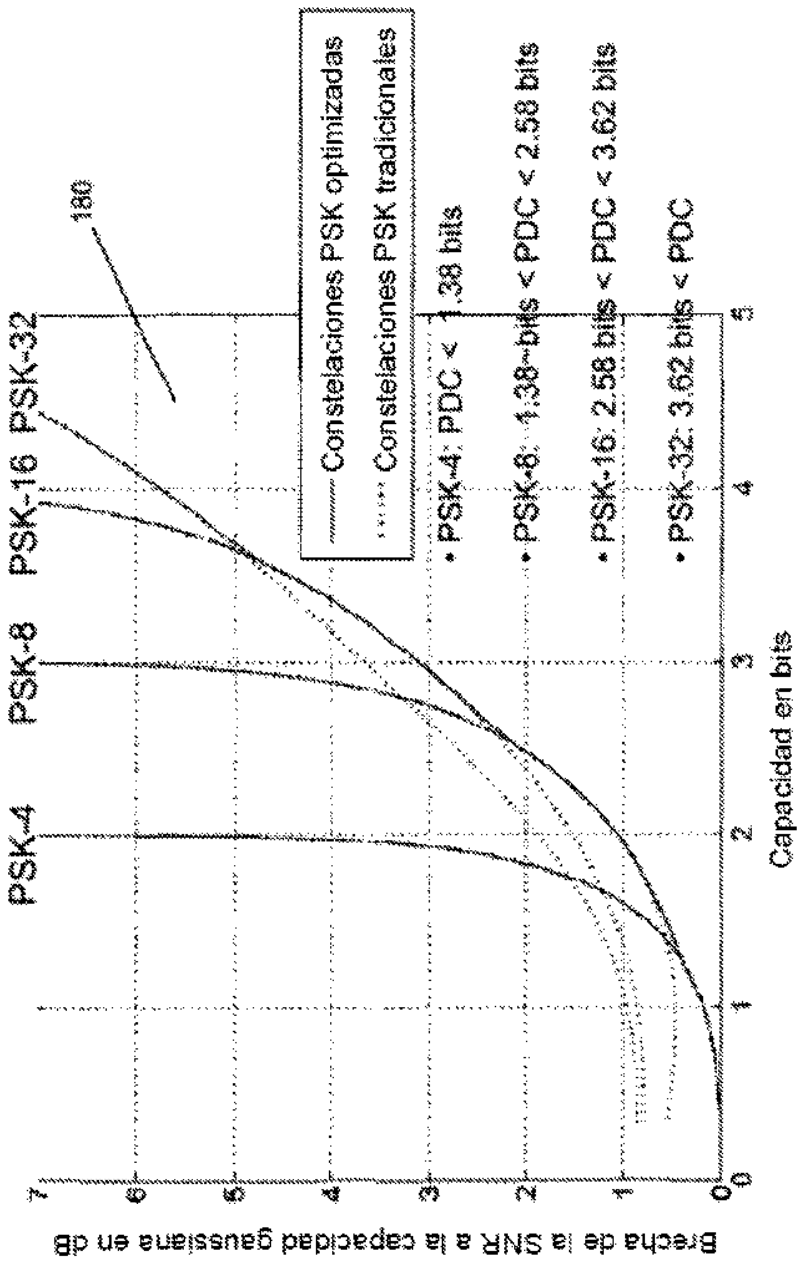

180
PSK-4
PSK-8
PSK-16
PSK-32
Constelaciones PSK optimizadas
Constelaciones PSK tradicionales
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38 bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
Capacidad en bits
Brecha de la SNR a la capacidad gaussiana en dB
0
1
2
3
4
5
6
7

Fig. 18

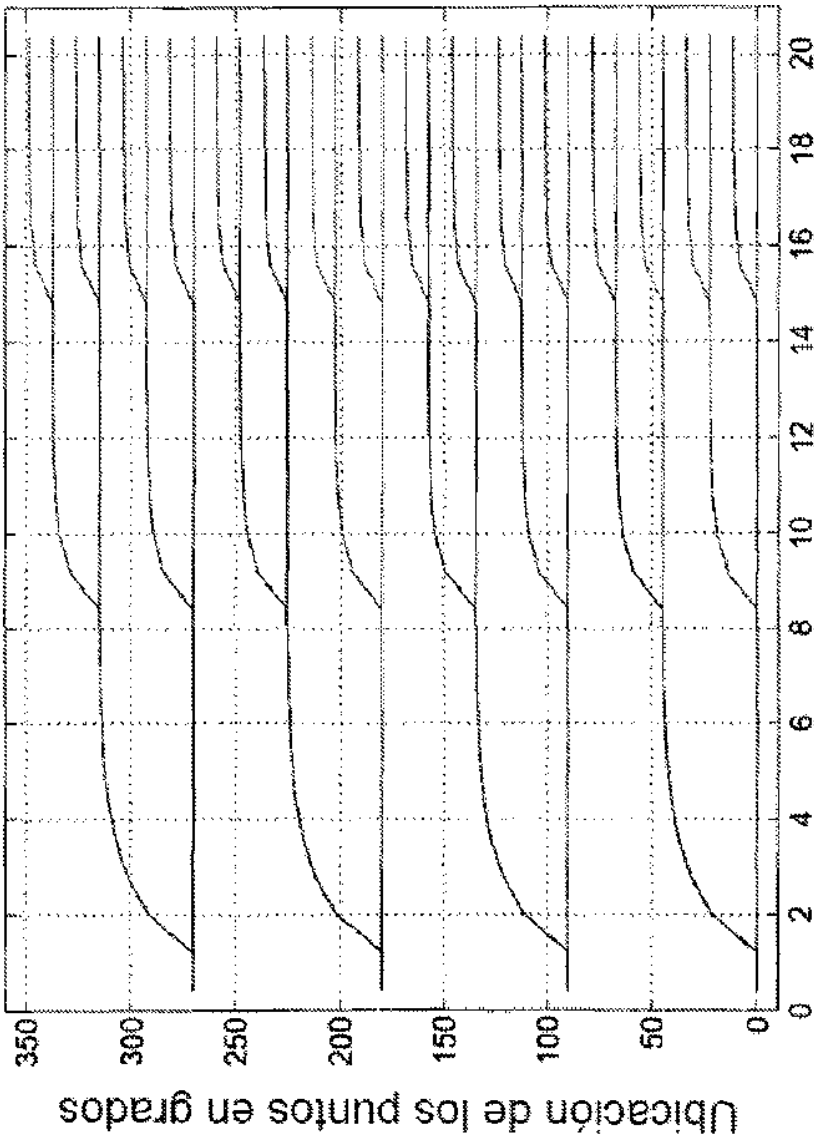

Ubicación de los puntos en grados
SNR en dB
0
50
100
150
200
250
300
350
0
2
4
6
8
10
12
14
16
18
20

Fig. 19

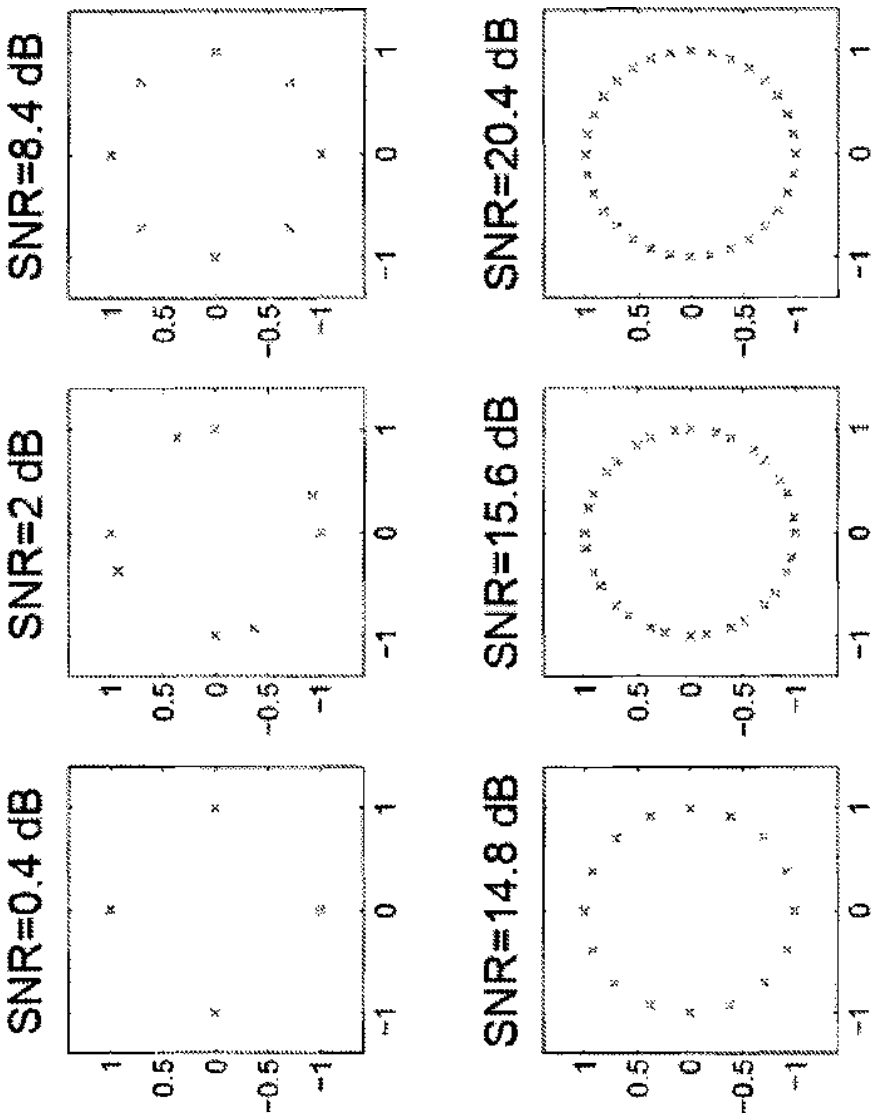


Fig. 20

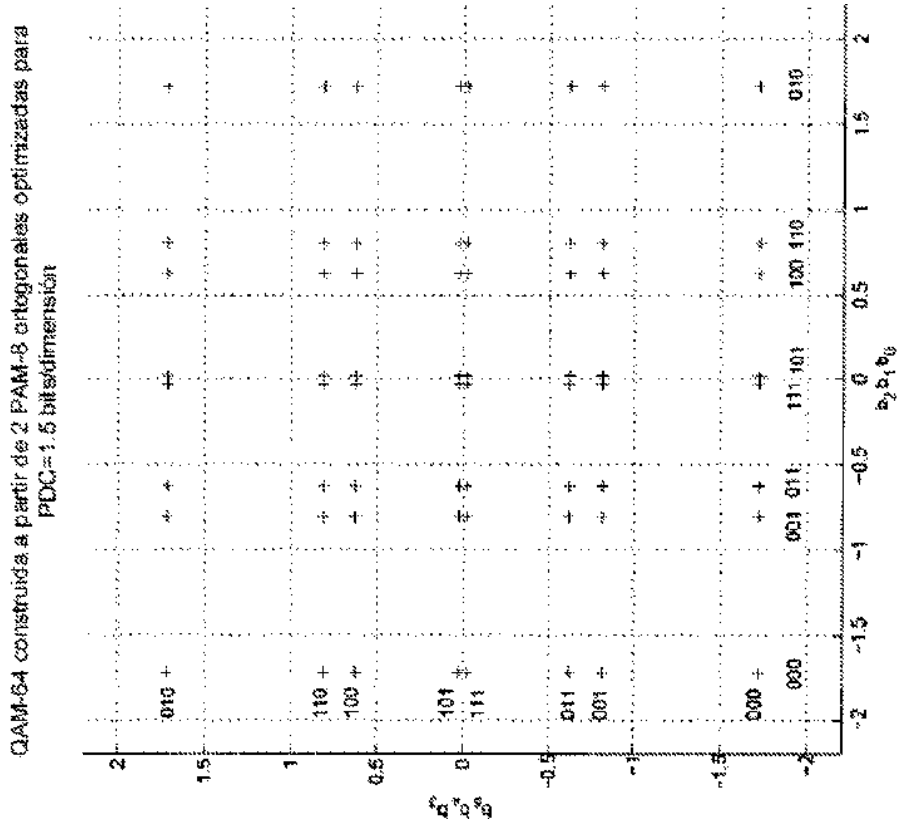

QAM-64 construida a partir de 2 PAM-8 ortogonales optimizadas para
PEDC=1.5 bits/dimensión
010
110
100
101
111
011
001
000
010
110 100
101 111
011 001
000
$s_Q s_Q s_Q$
$s_I s_I s_I$

Fig. 21

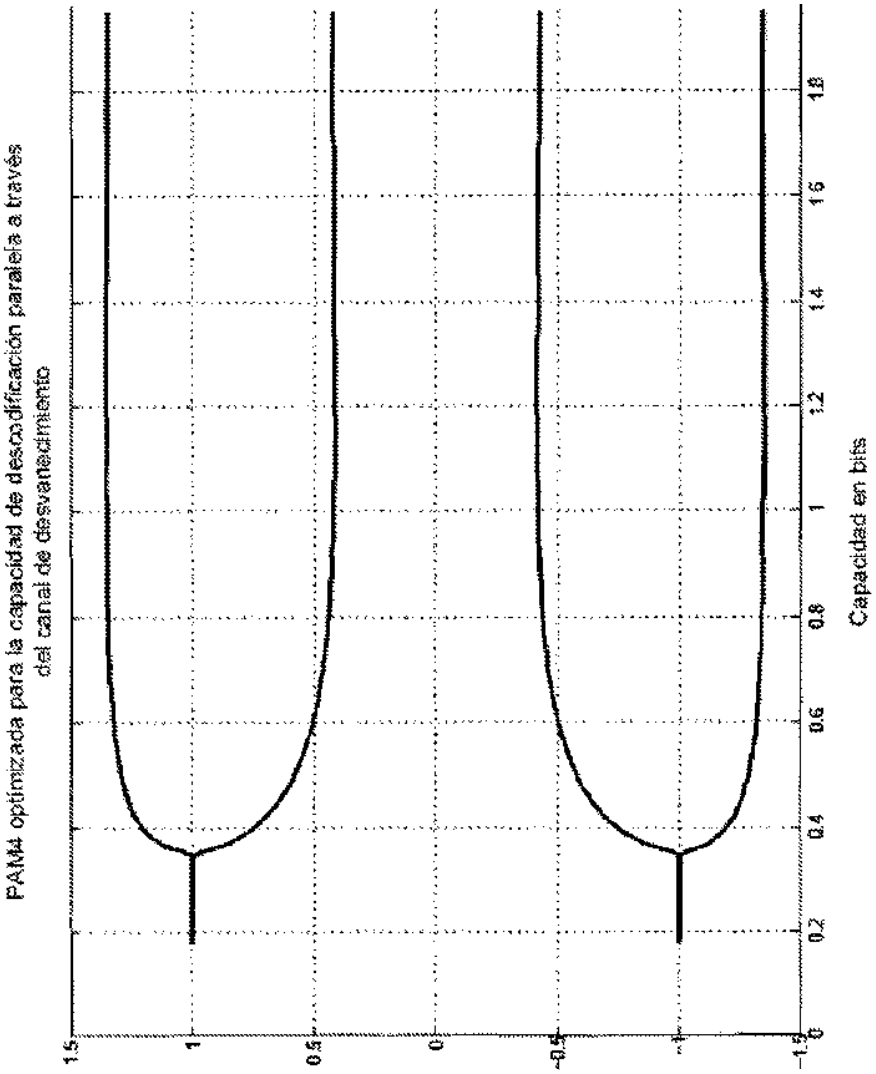

PAM4 optimizada para la capacidad de descodificación paralela a través del canal de desvanecimiento
Capacidad en bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

Fig. 22a

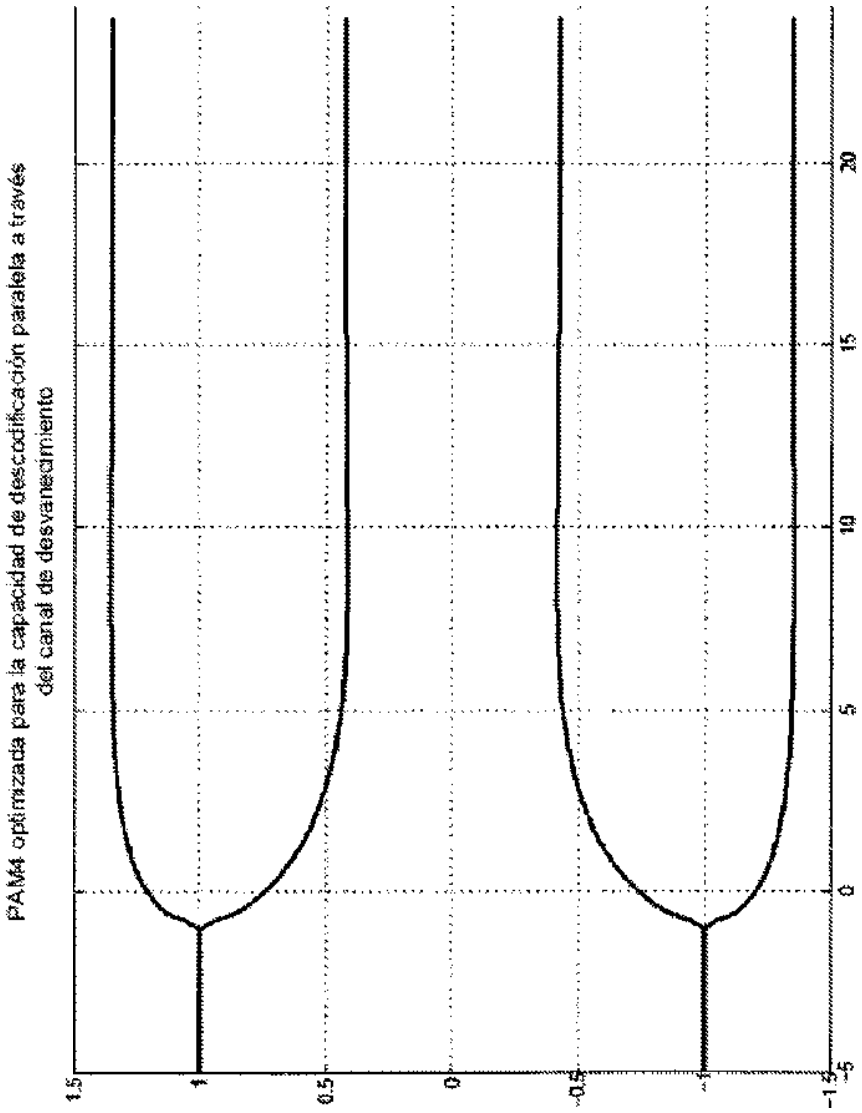

PAM4 optimizada para la capacidad de descodificación paralela a través del canal de desvanecimiento
SNR en dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20

Fig. 22b

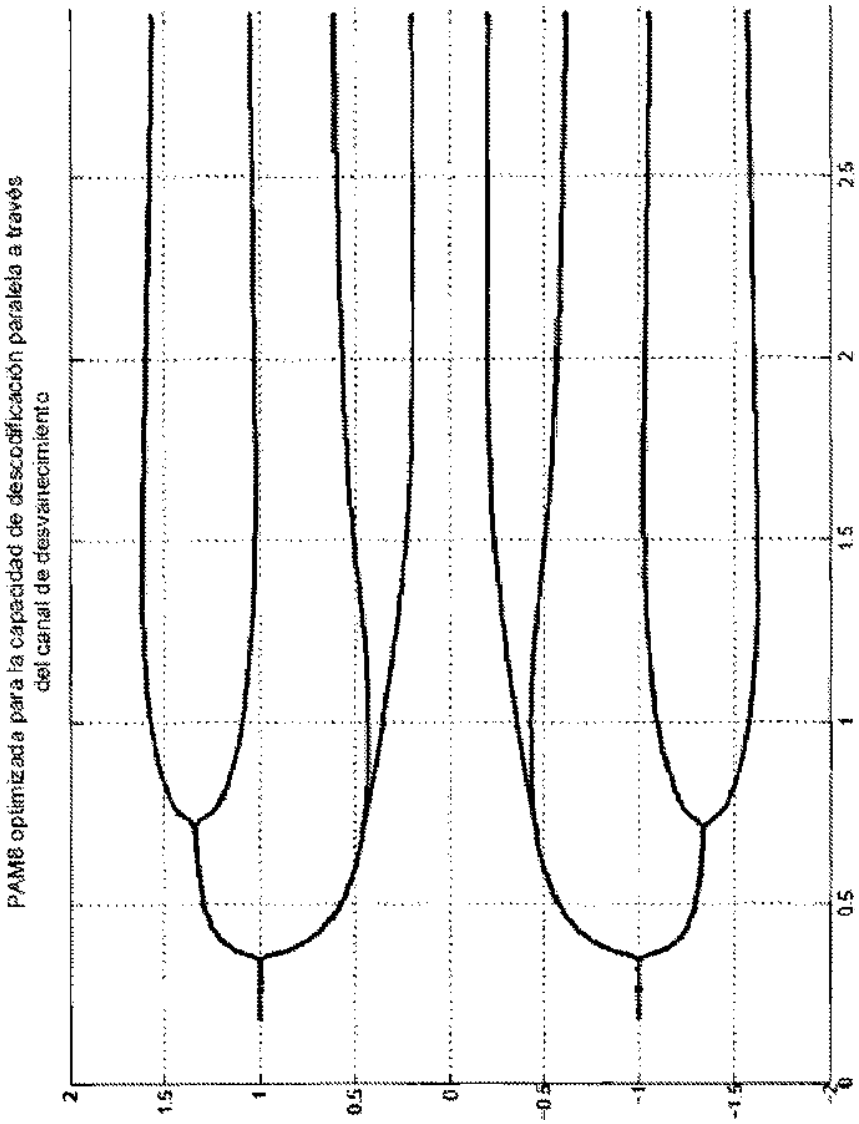

PAM8 optimizada para la capacidad de descodificación paralela a través del canal de desvanecimiento
Capacidad en bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

Fig. 23a

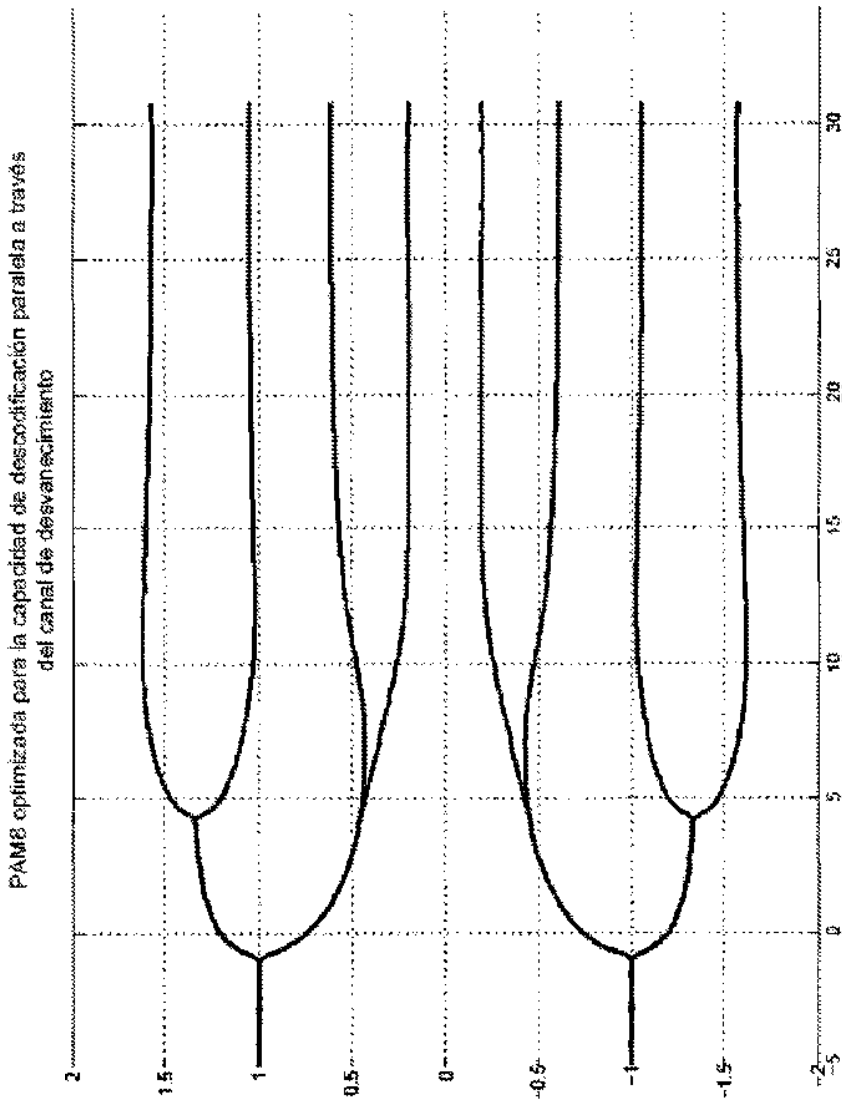

PAM8 optimizada para la capacidad de descodificación paralela a través del canal de desvanecimiento
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30
SNR en dB

Fig. 23b

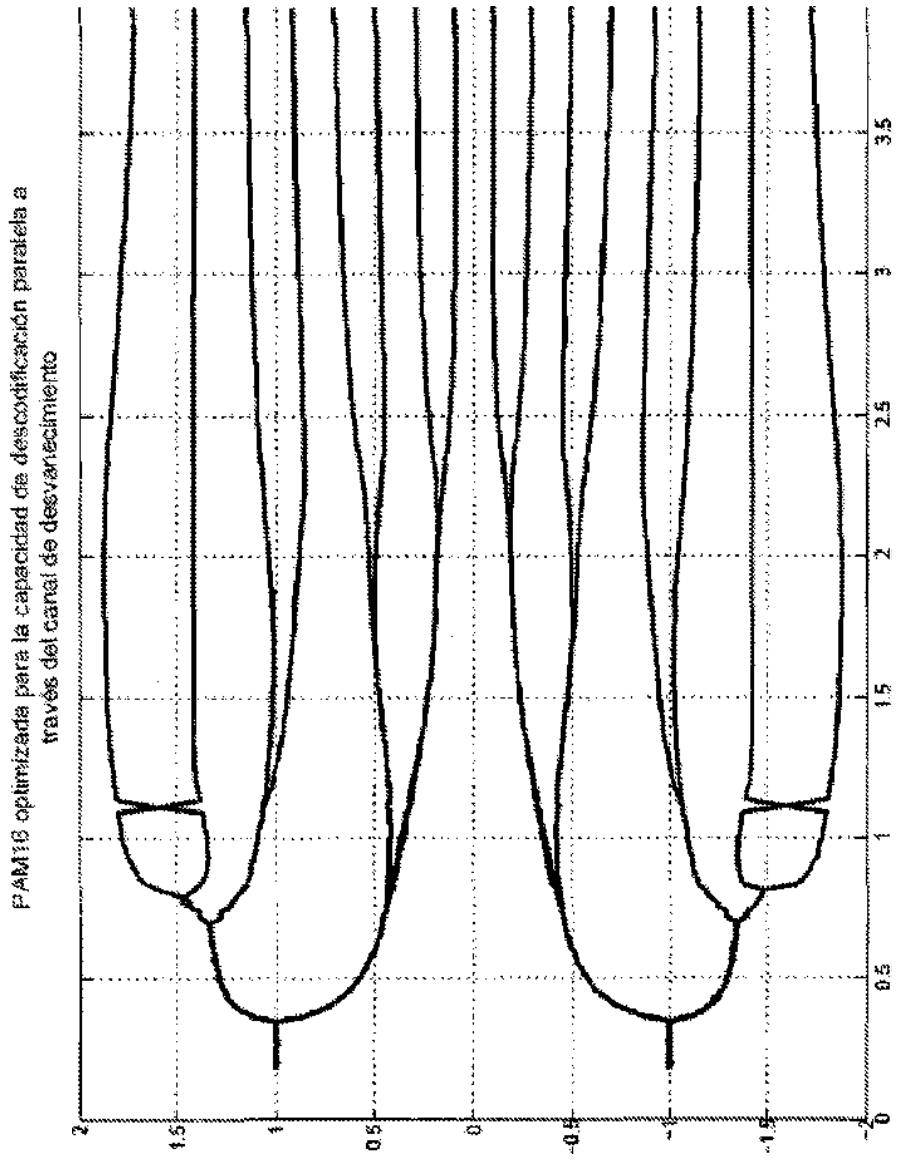

PAM16 optimizada para la capacidad de descodificación paralela a través del canal de desvanecimiento
Capacidad en bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

Fig. 24a

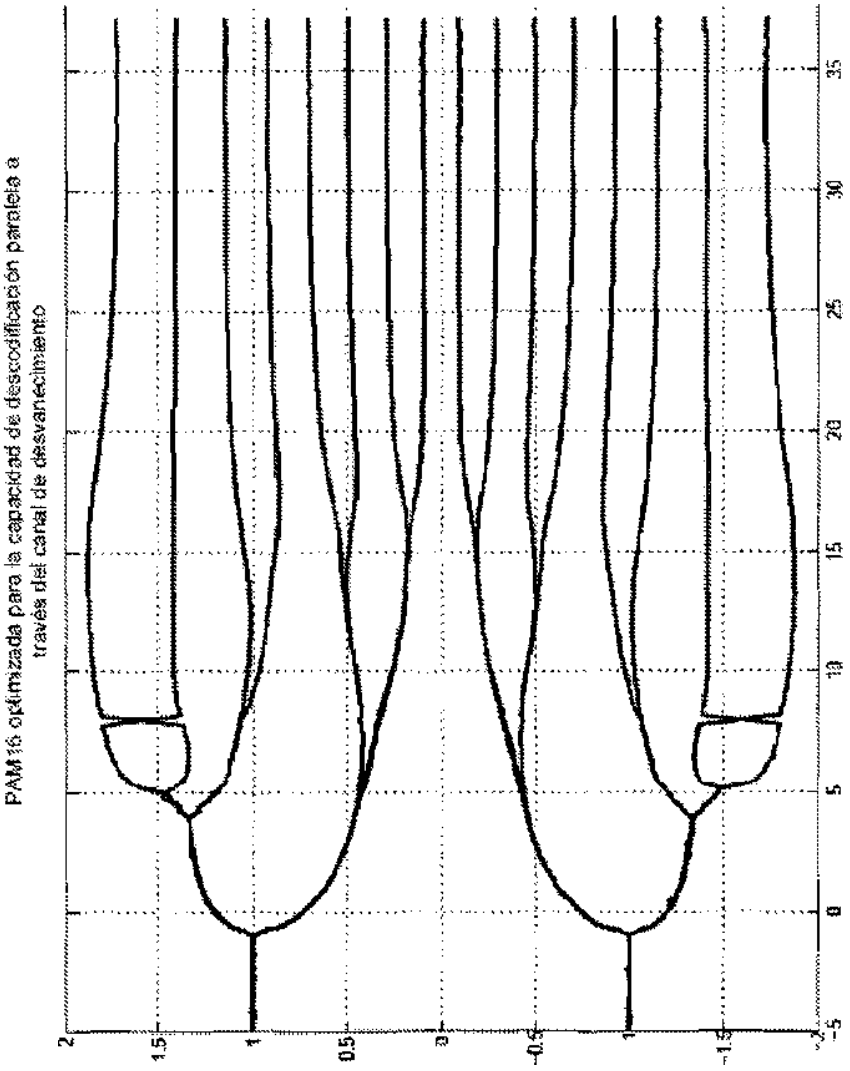

PAM16 optimazada para la capacidad de descodificación paralela a través del canal de desvanecimiento
SNR en dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
35
30
25
20
15
10
5
0
-5

Fig. 24b

[51] H04L 25/02 [11] 40011604
H04L 25/03 CN110086733 A
H04B 7/06
[13] A
[25] ZH
[21] 42020002139.2 [22] 03.02.2020
[30] 11.05.2012 US,13/469843
[54] METHODS AND ARRANGEMENTS FOR CSI REPORTING
用於 CSI 報告的方法和裝置
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[51] H04L 25/03 [11] 40011604
H04L 25/02 CN110086733 A
[13] A

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[51] H04L 27/00 [11] 40011550
H04L 5/00 CN110098913 A
[13] A

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[51] H04L 27/26 [11] 40011550
H04L 5/00 CN110098913 A
[13] A

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[51] H04L 27/26 [11] 40011785
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[13] A

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[51] H04L 27/34 [11] 40011480
H04L 1/00 EP3518485 A1
[13] A
[25] EN
[21] 42020001313.4 [22] 13.01.2020

[30] 05.06.2007 US,60/933,319



[54] METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS
用於具優化容量星座的信號的方法及配置

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[51] H04L 29/06
H04L 29/08

[11] 40011346
CN110062016 A
[13] A

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[51] H04L 29/06
H04L 12/26

[11] 40011365
CN110061887 A
[13] A

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[51] H04L 29/06
H04L 29/08

[11] 40011406
CN110099074 A
[13] A

[25] ZH
[21] 42020001021.3 [22] 06.01.2020

[54] ANOMALY DETECTION METHOD AND SYSTEM FOR INTERNET OF THINGS DEVICE AND ELECTRONIC DEVICE
一種物聯網設備的異常檢測方法、系統及電子設備

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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}.

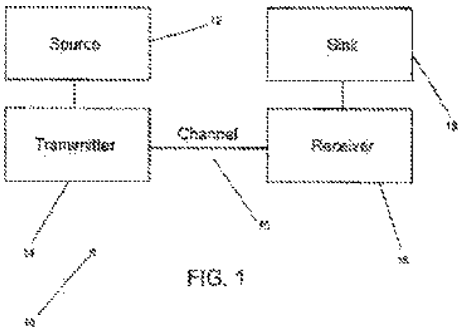


FIG. 1

EP 3 518 485 A1

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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 ρ (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.

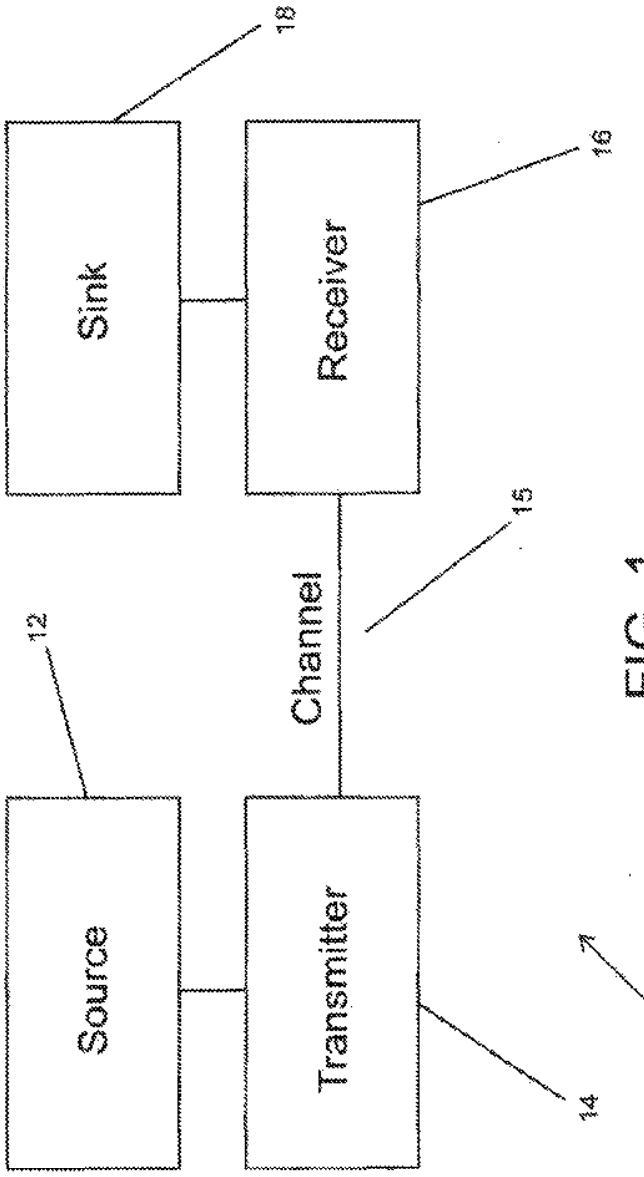


FIG. 1

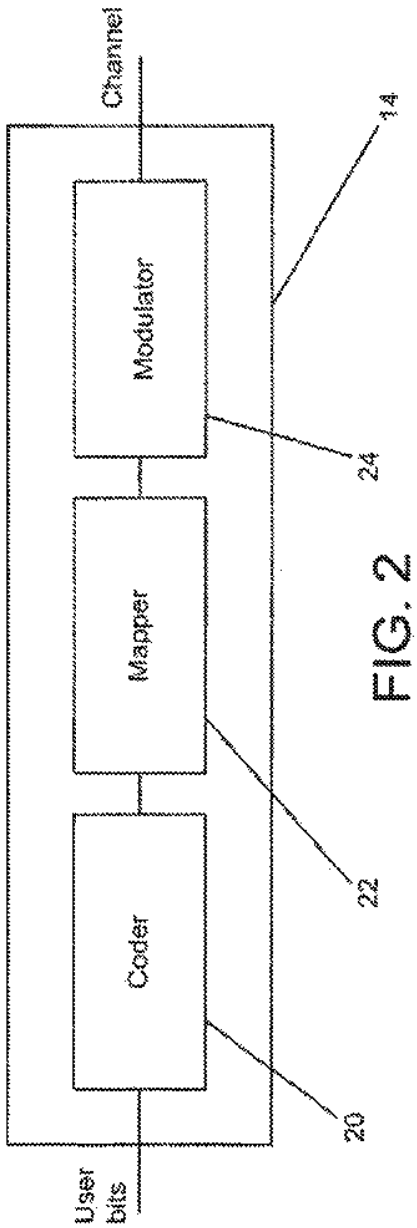

Channel
Modulator
Mapper
Coder
User bits
14
24
22
20

FIG. 2

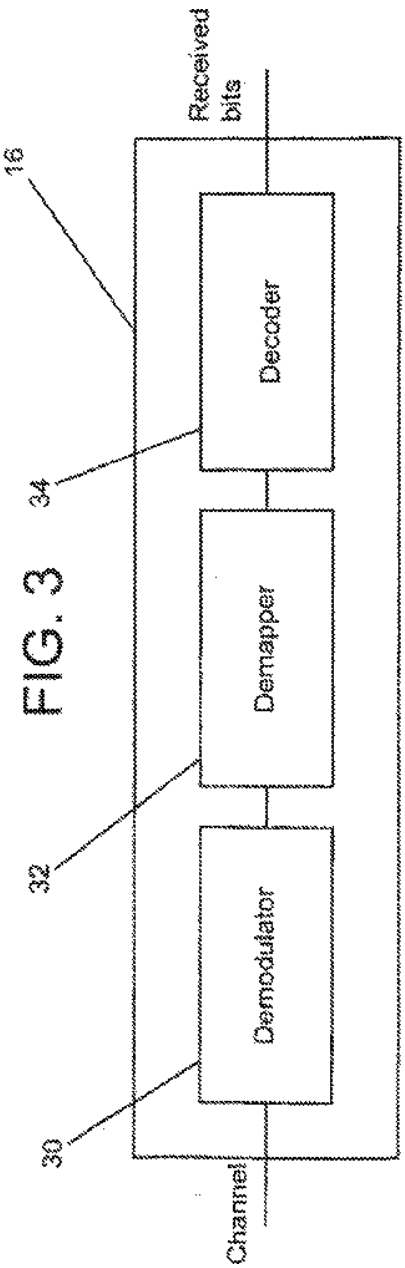

Received bits
Decoder
Demapper
Demodulator
Channel
16
34
32
30

FIG. 3

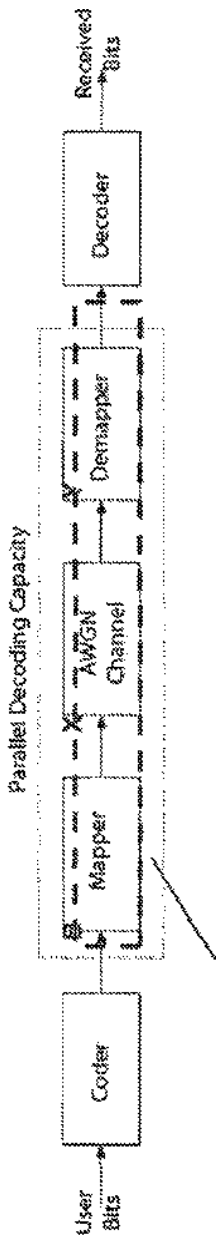

Parallel Decoding Capacity
User Bits
Coder
B
Mapper
X
AWGN Channel
Demapper
Decoder
Received Bits
40

FIG. 4a

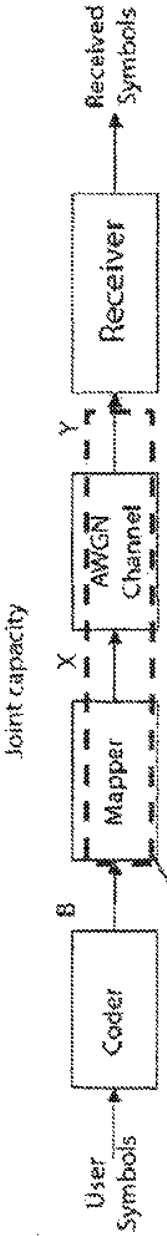

Joint capacity
User Symbols
Coder
B
Mapper
X
AWGN Channel
Y
Receiver
Received Symbols
42

FIG. 4b

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

FIG. 5

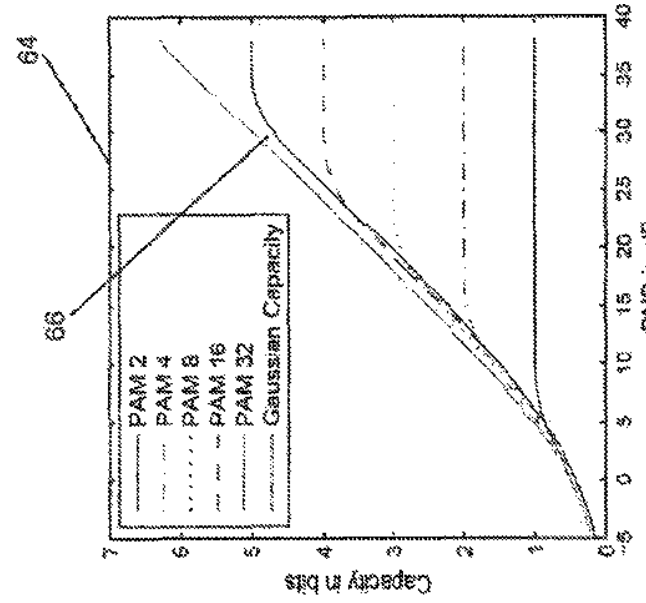

64
66
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(b)

FIG. 6b

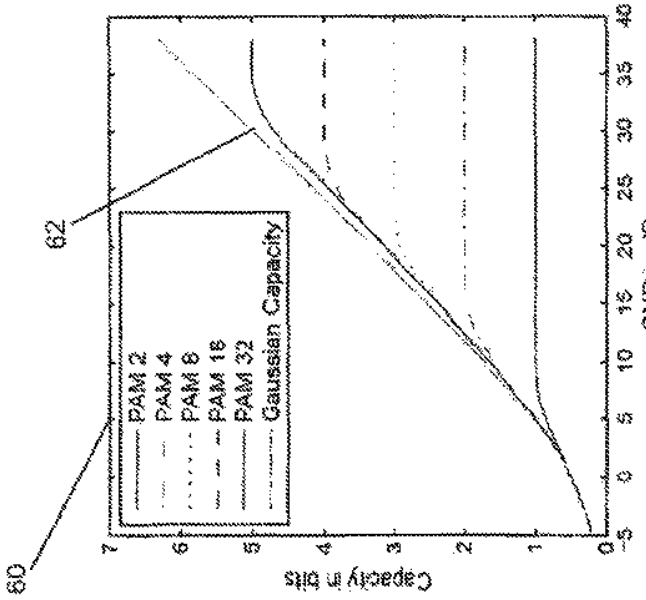

60
62
PAM 2
PAM 4
PAM 8
PAM 16
PAM 32
Gaussian Capacity
Capacity in bits
SNR in dB
(a)

FIG. 6a

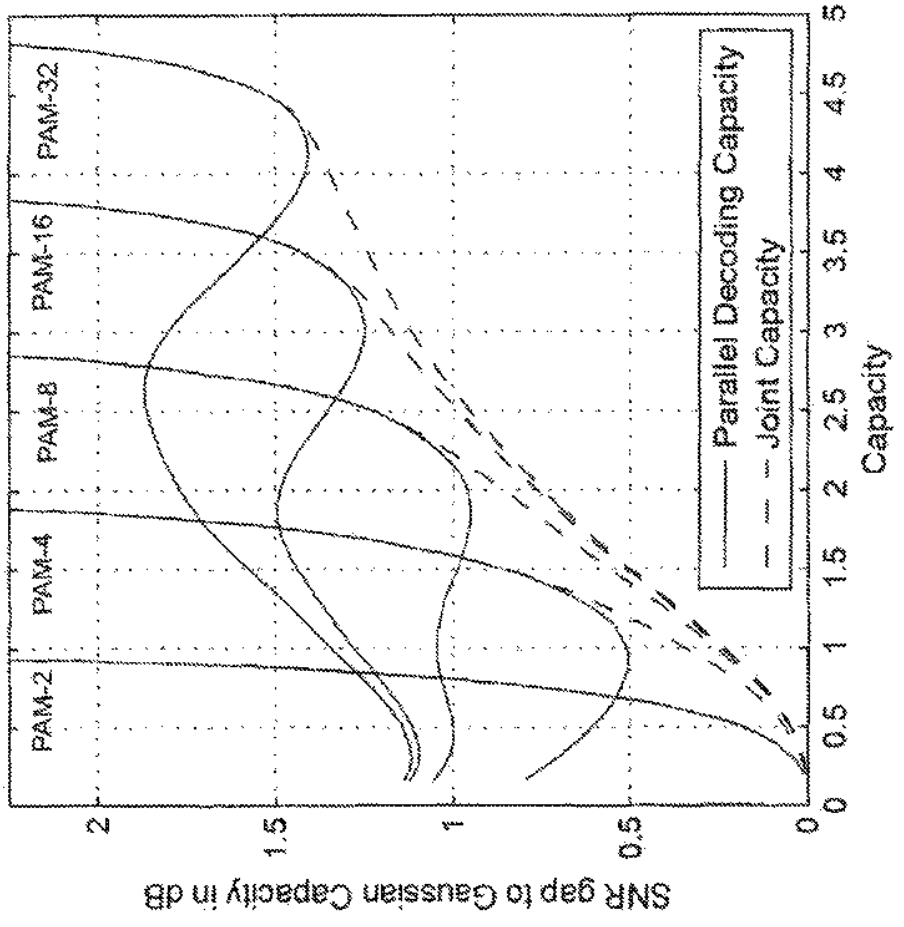

PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
Parallel Decoding Capacity
Joint Capacity
SNR gap to Gaussian Capacity in dB
Capacity
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 7

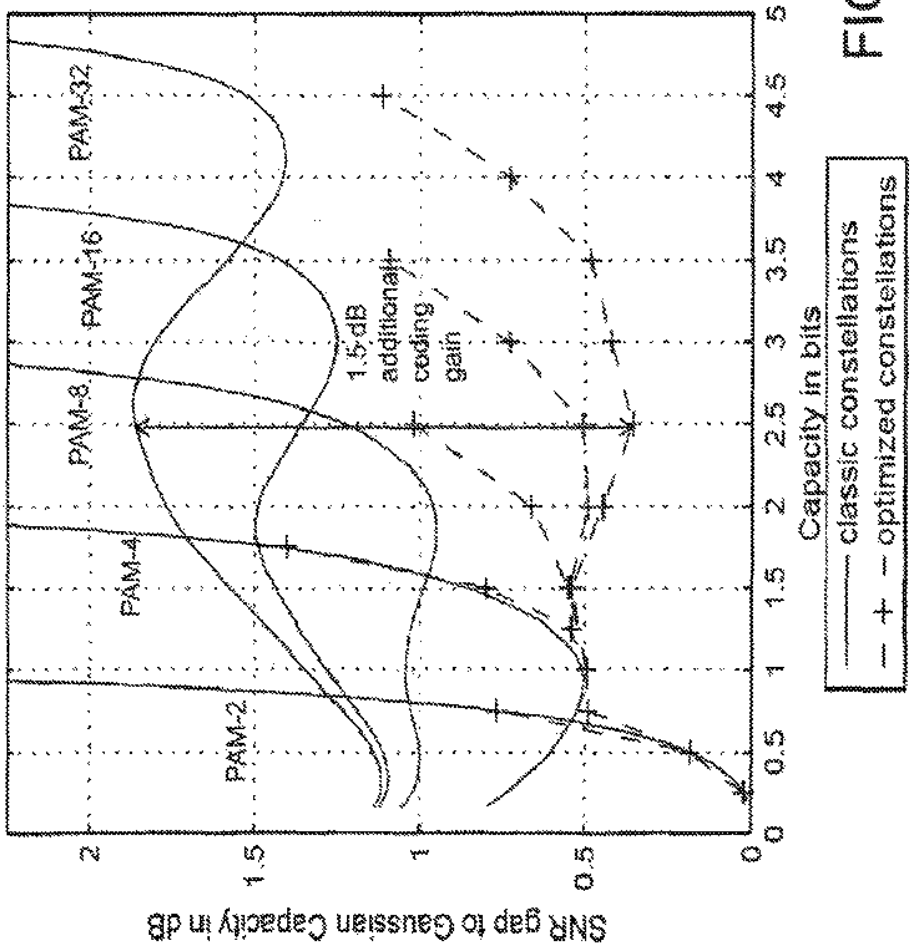

SNR gap to Gaussian Capacity in dB
Capacity in bits
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
1.5 dB additional coding gain
classic constellations
optimized constellations

FIG. 8a

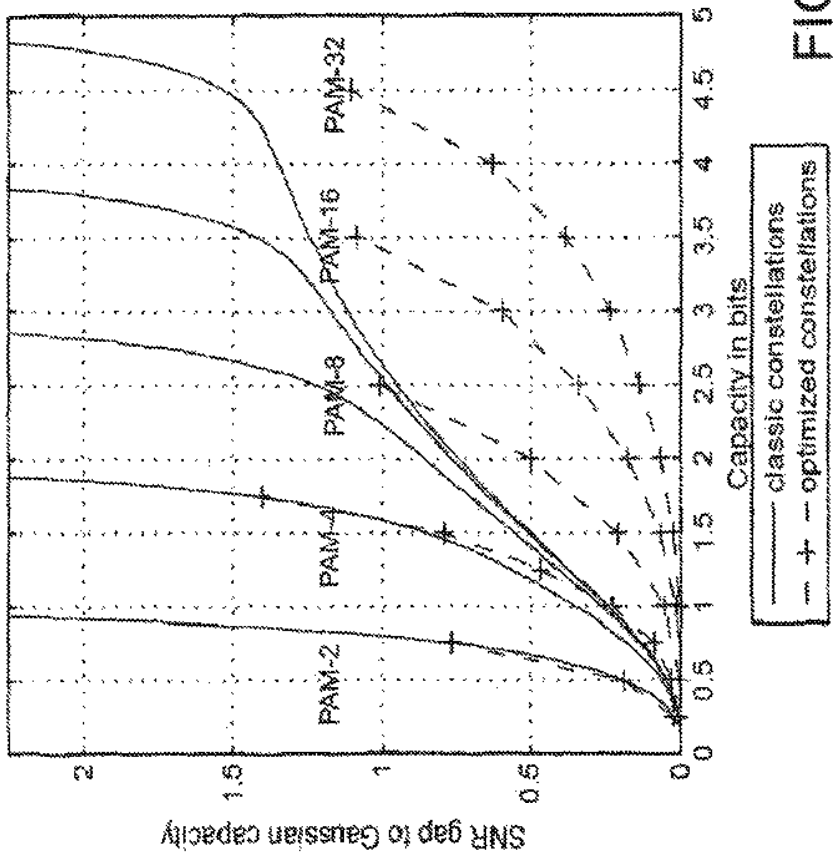

SNR gap to Gaussian capacity
Capacity in bits
classic constellations
optimized constellations
PAM-2
PAM-4
PAM-8
PAM-16
PAM-32
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5

FIG. 8b

FER
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
16.5
17
17.5
18
18.5
19
19.5
20
SNR (dB)
$\Delta_{1/2}$ = 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

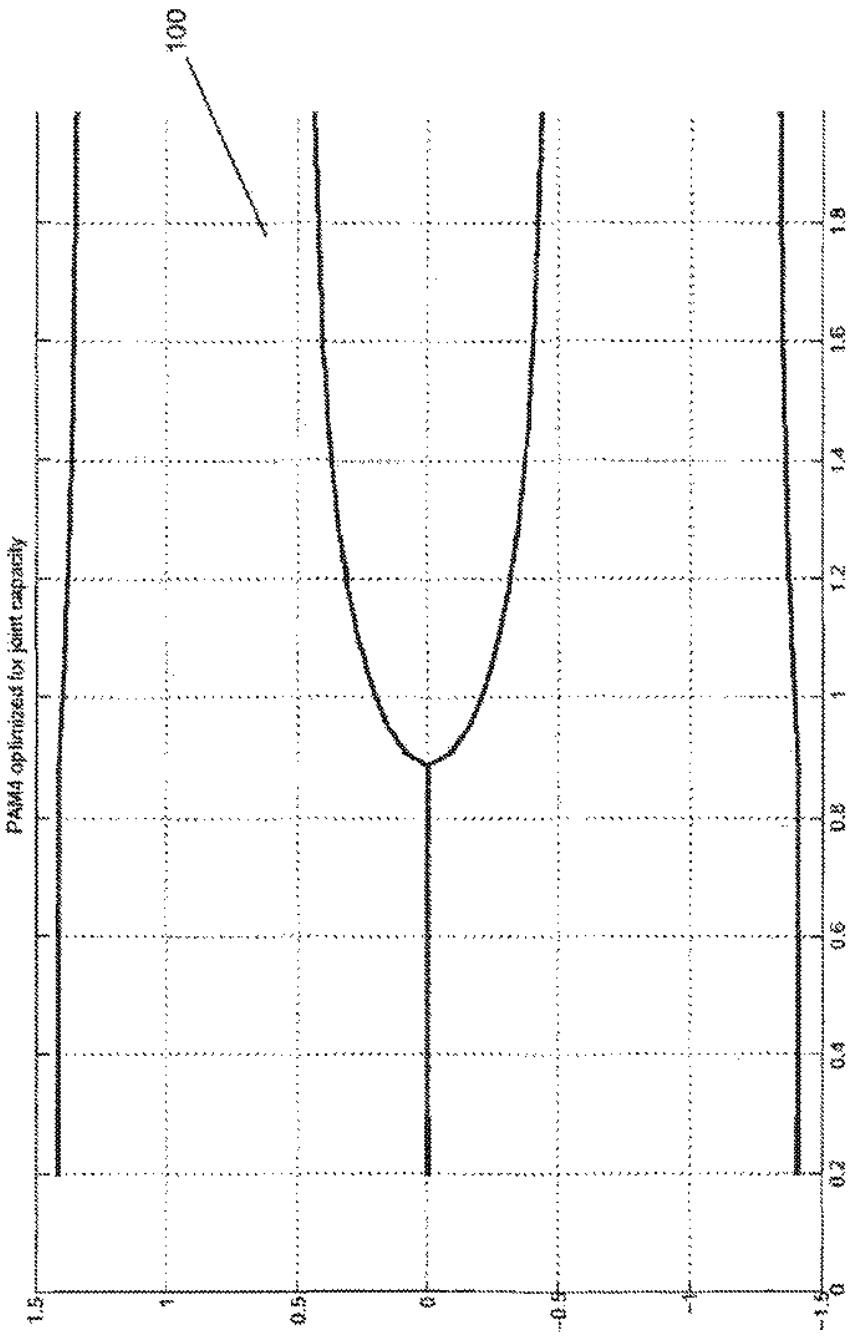

100
PAM4 optimized for joint capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10a

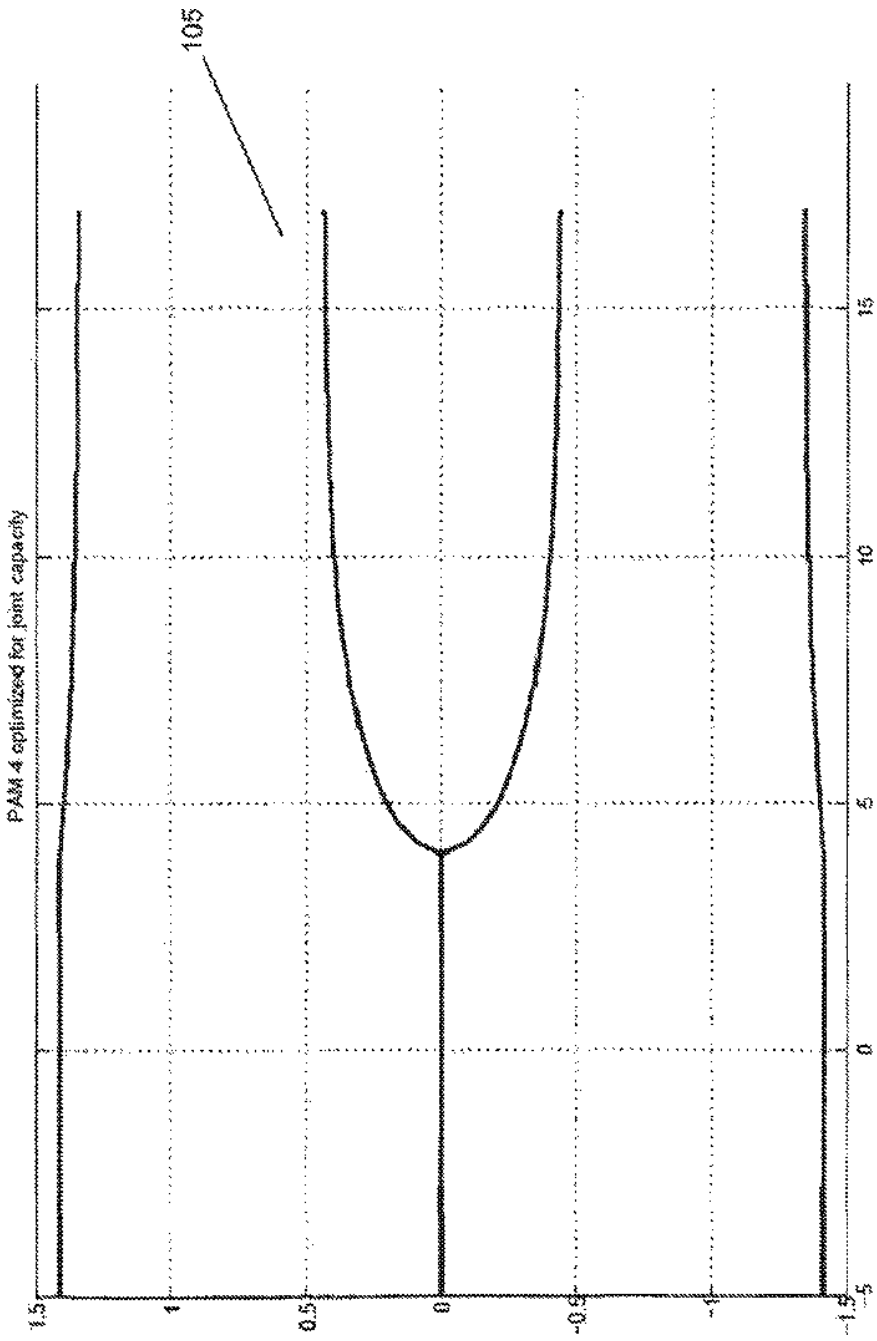

105
PAM 4 optimized for joint capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

FIG. 10b

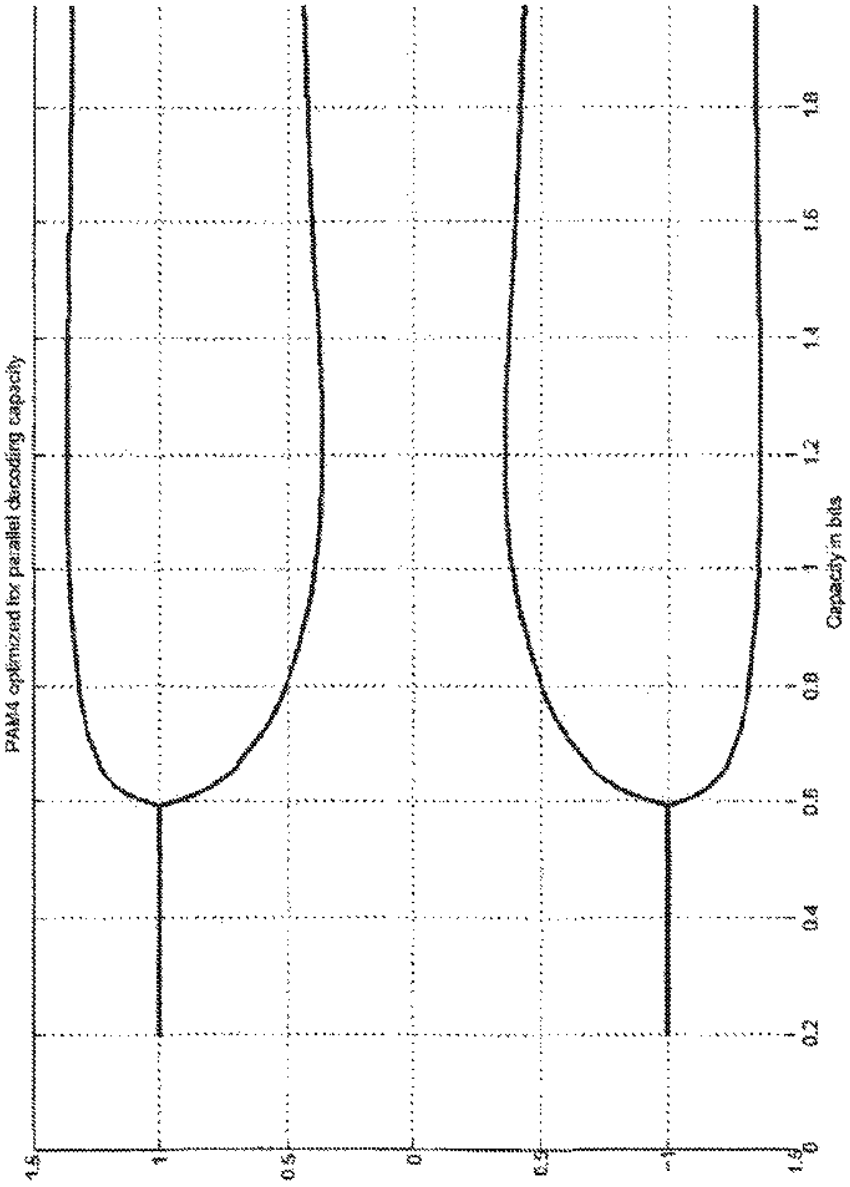

PAM4 optimized for parallel decoding capacity
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 10c

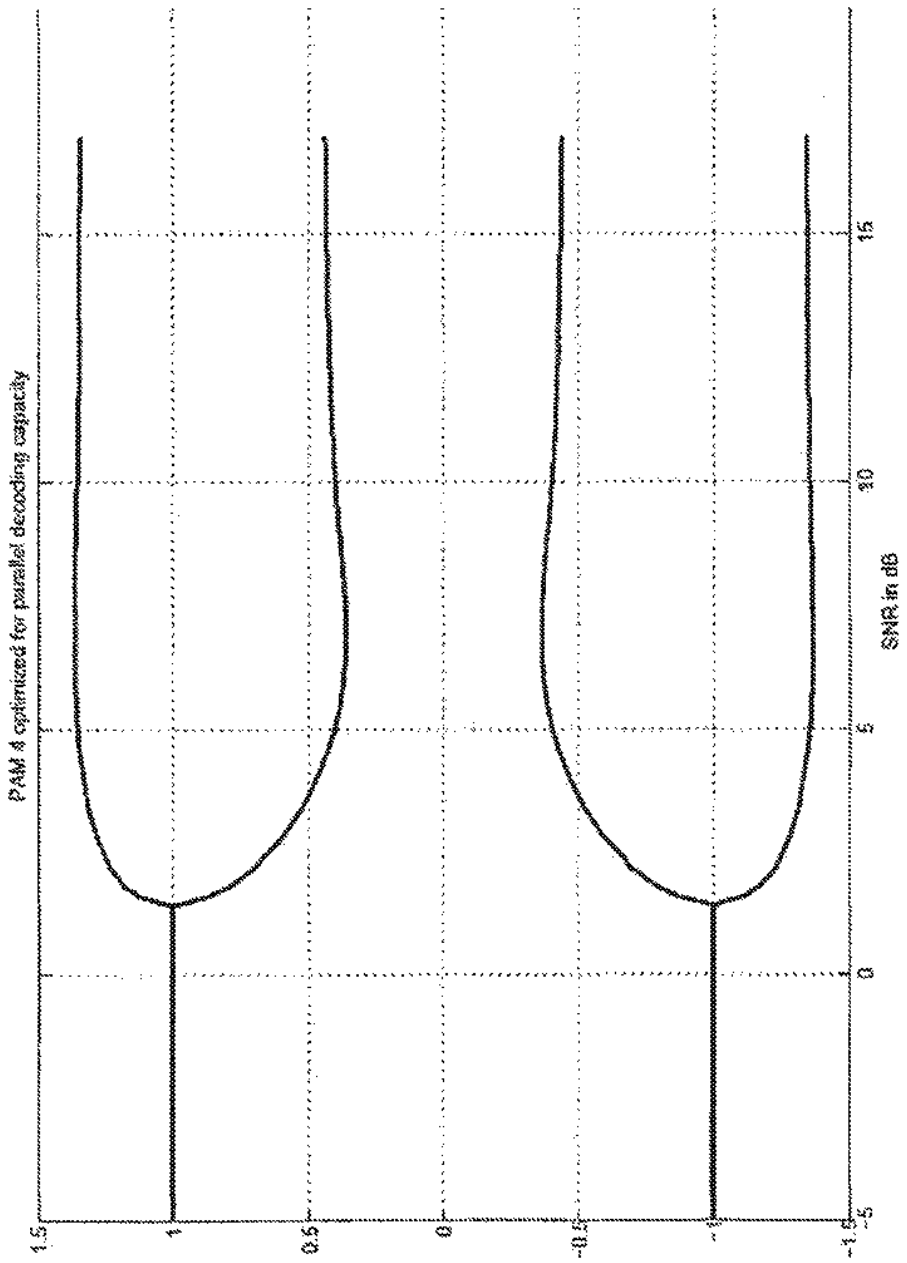

PAM 4 optimized for parallel decoding capacity
SNR in dB
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15

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

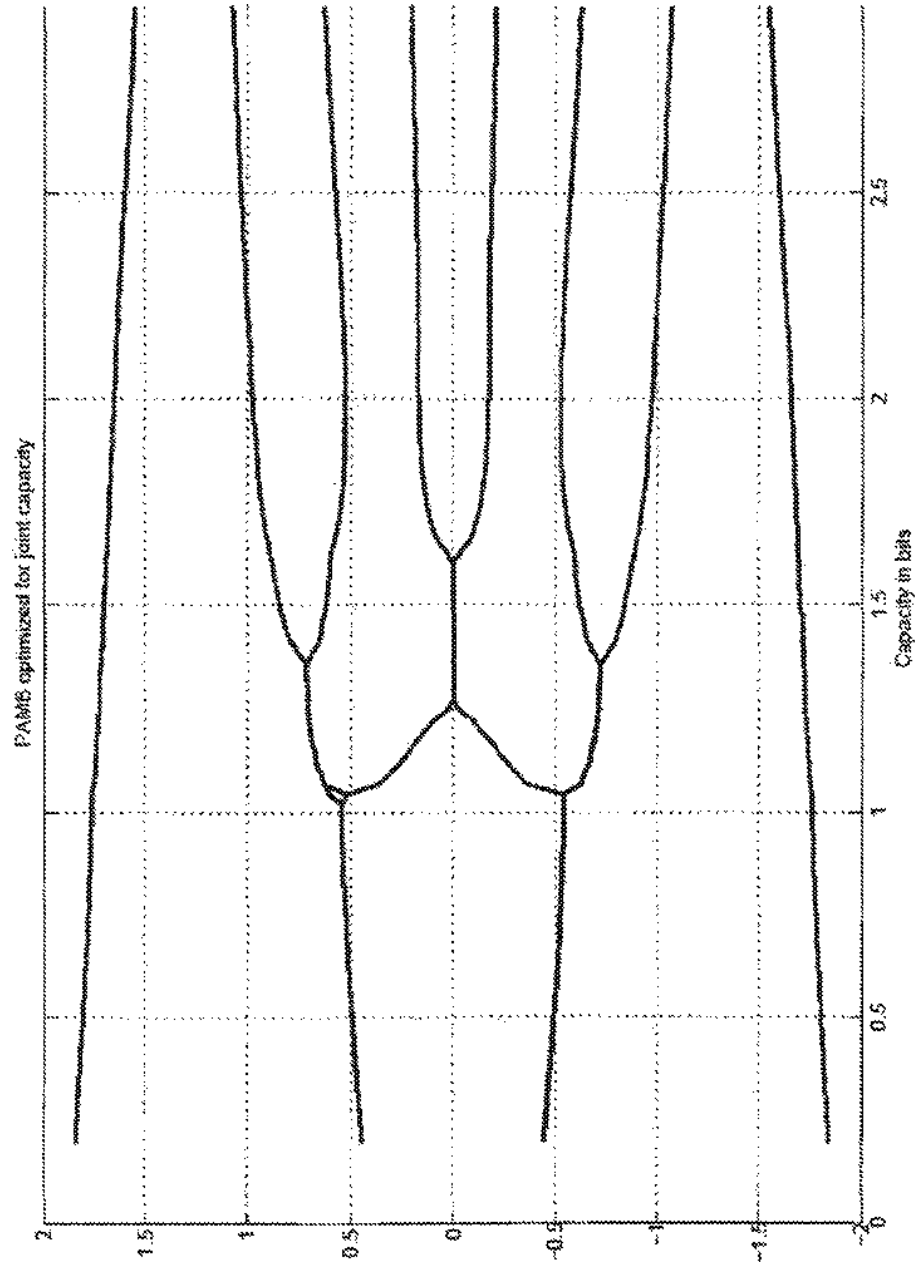

PAM6 optimized for joint capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12a

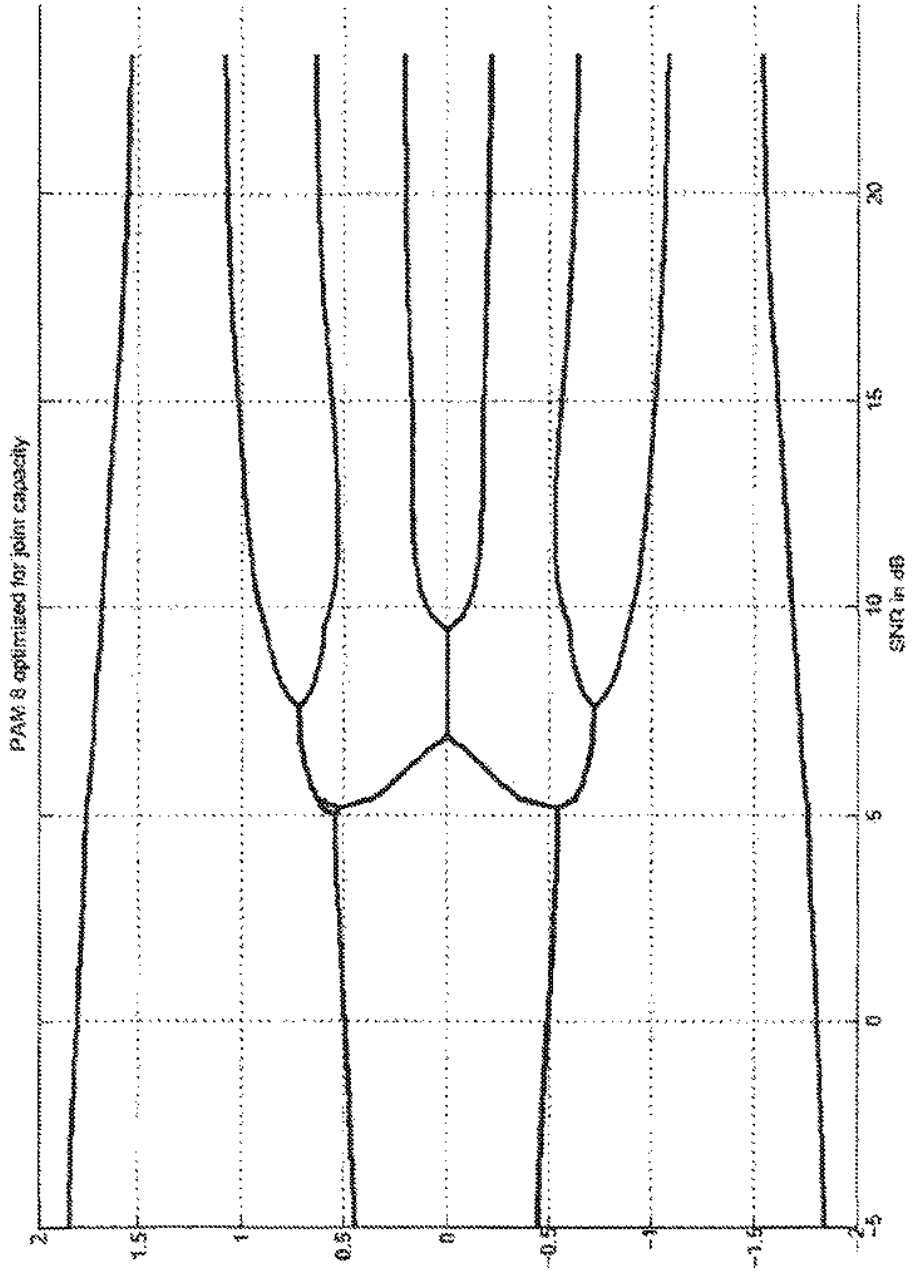

PAM-8 optimized for joint capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

FIG. 12b

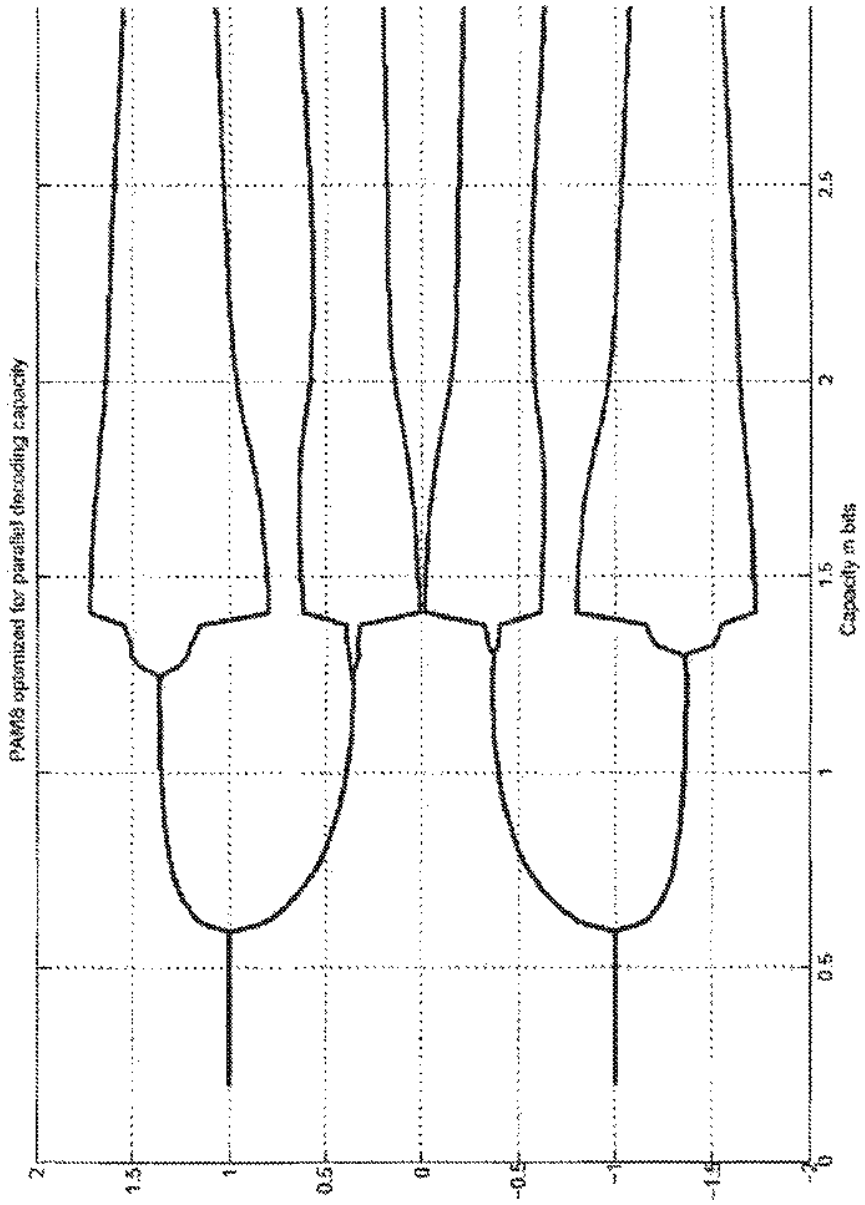

PAM8 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 12c

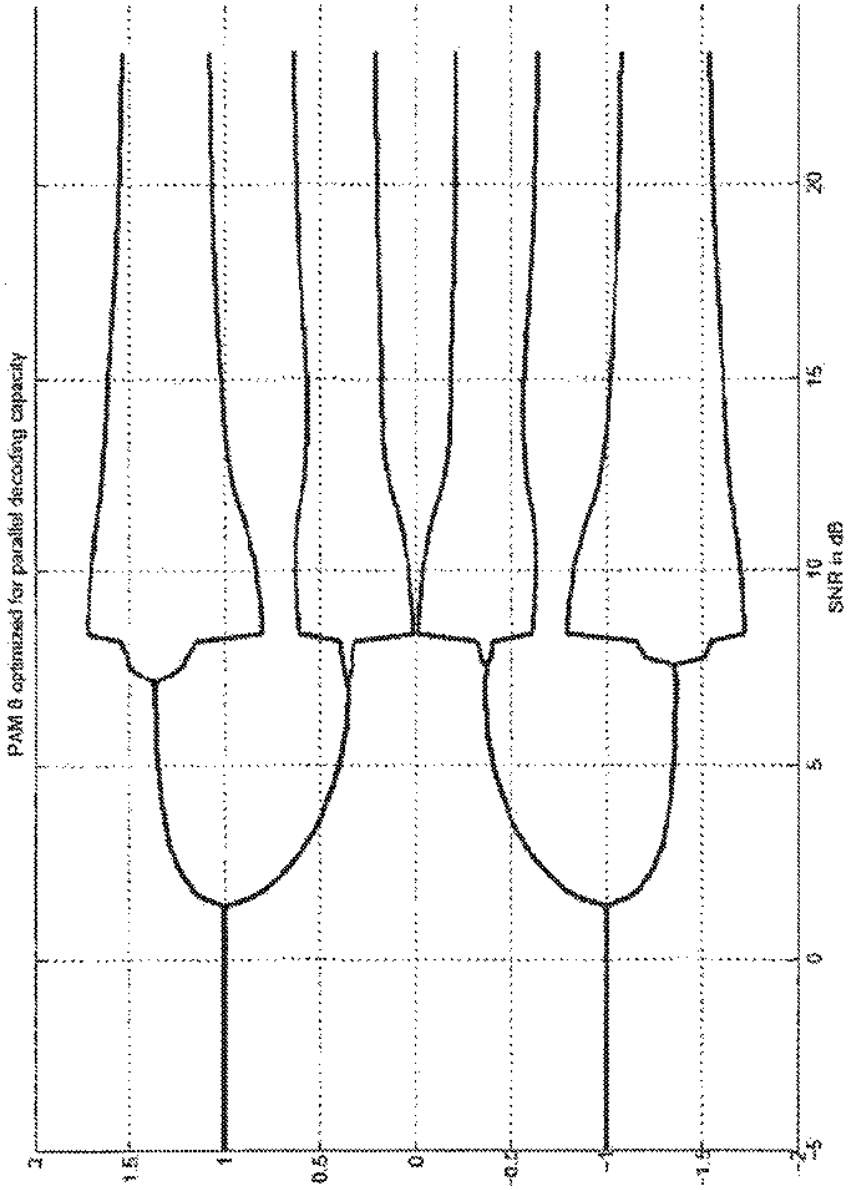

PAM 8 optimized for parallel decoding capacity
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20

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

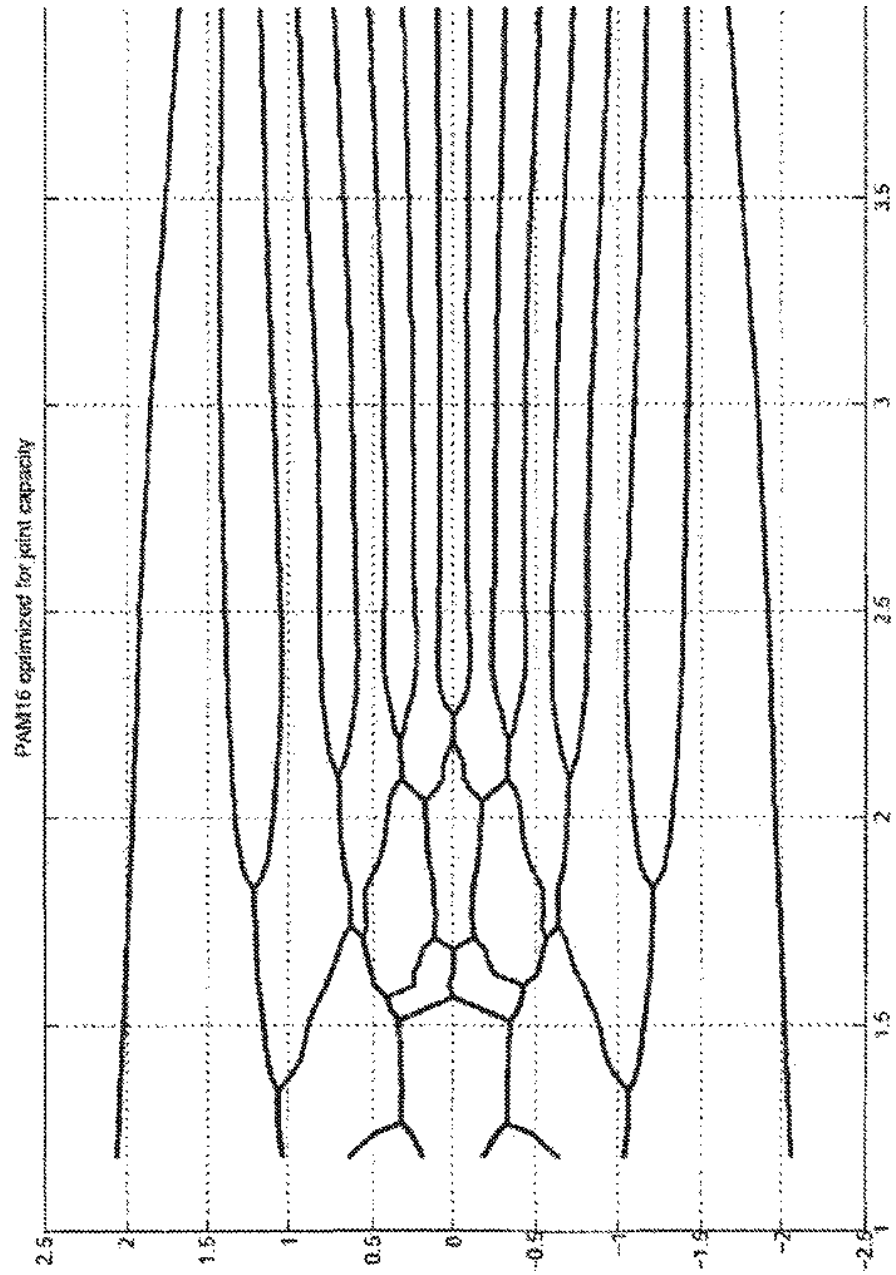

PAM16 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1
1.5
2
2.5
3
3.5

FIG. 14a

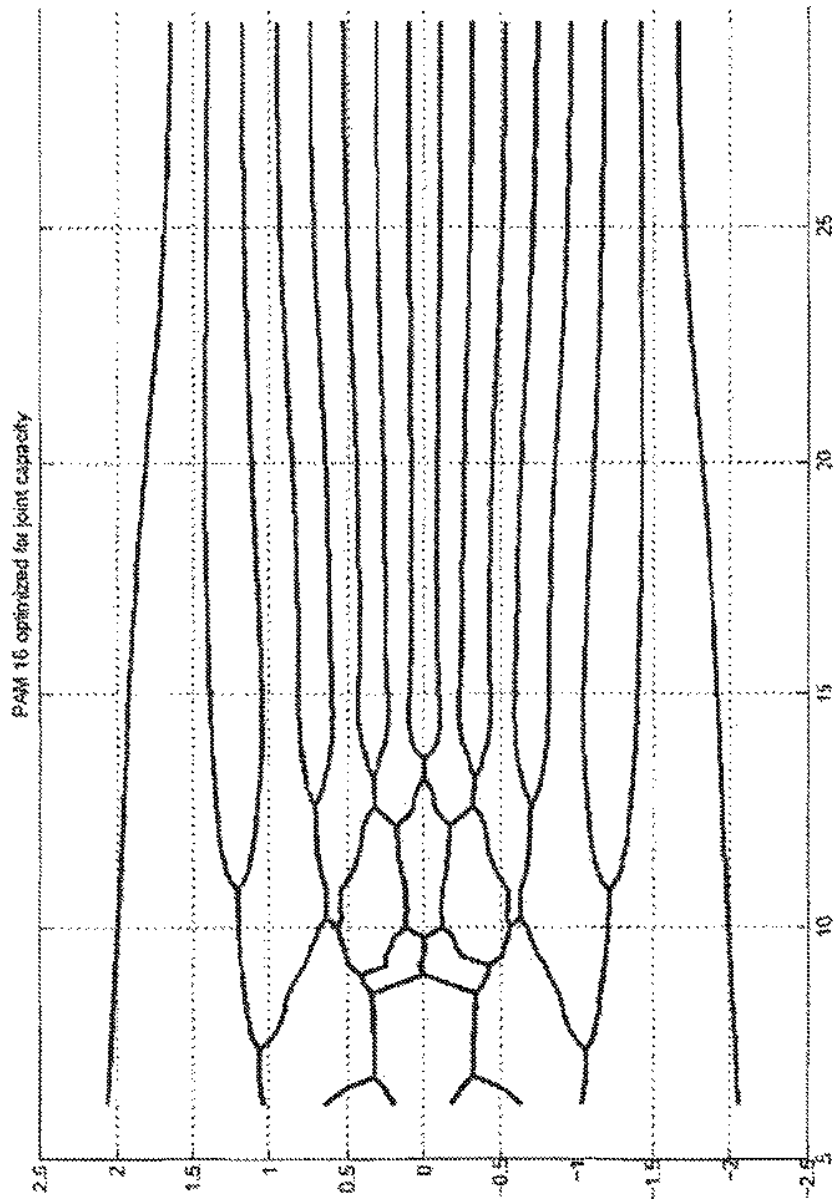

PAM 16 optimized for joint capacity
SNR in dB
25
20
15
10
5
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5

FIG. 14b

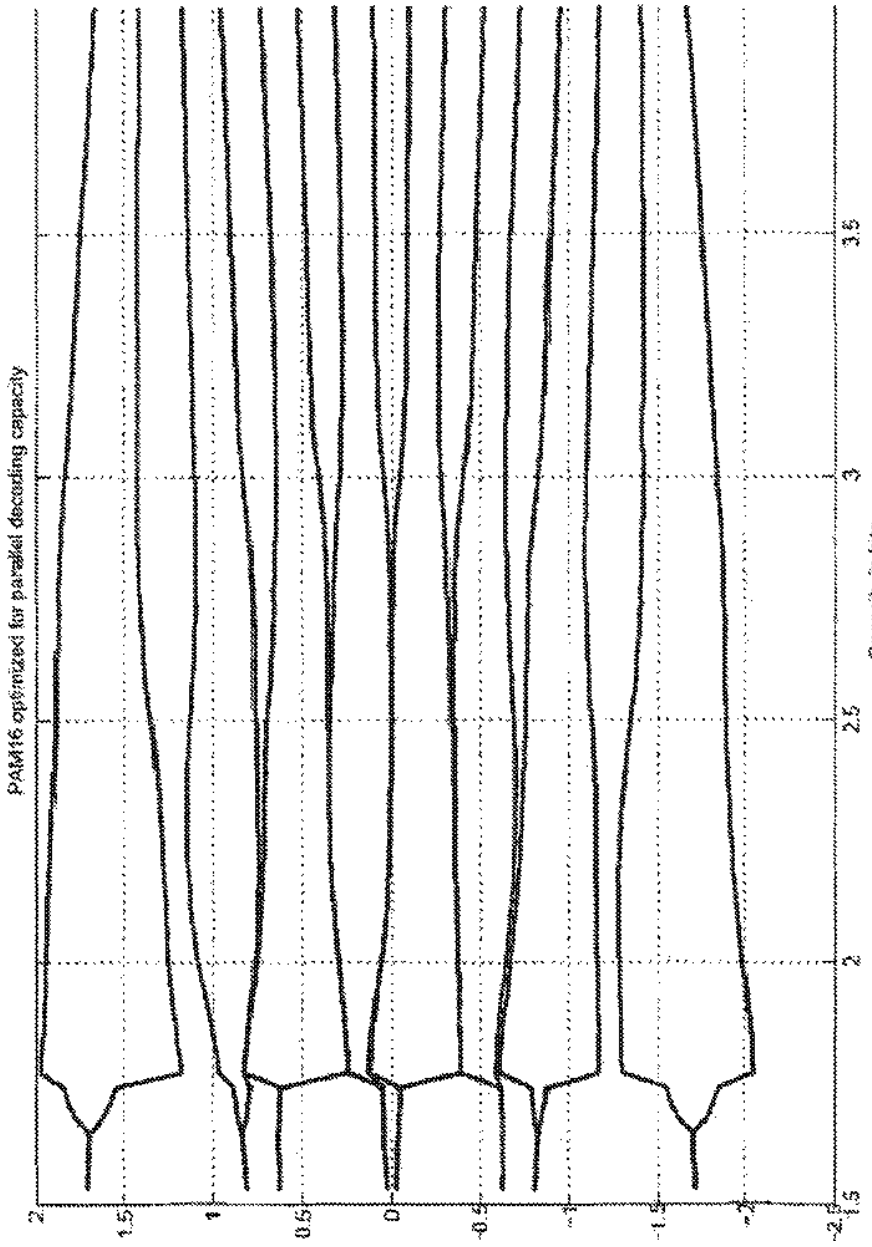

PAM16 optimized for parallel decoding capacity
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5

FIG. 14c

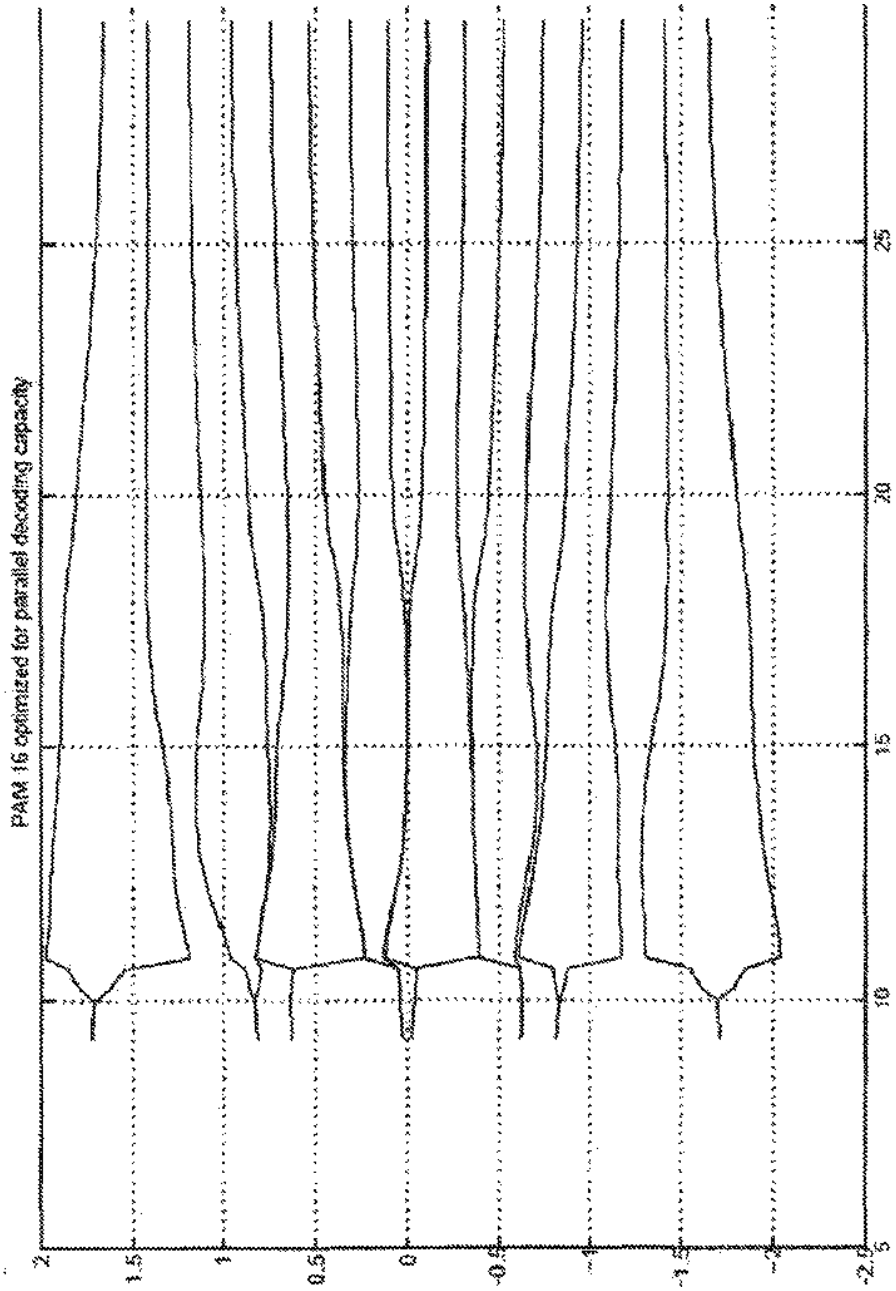

PAM 16 optimized for parallel decoding capacity
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
5
10
15
20
25
SNR in dB

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

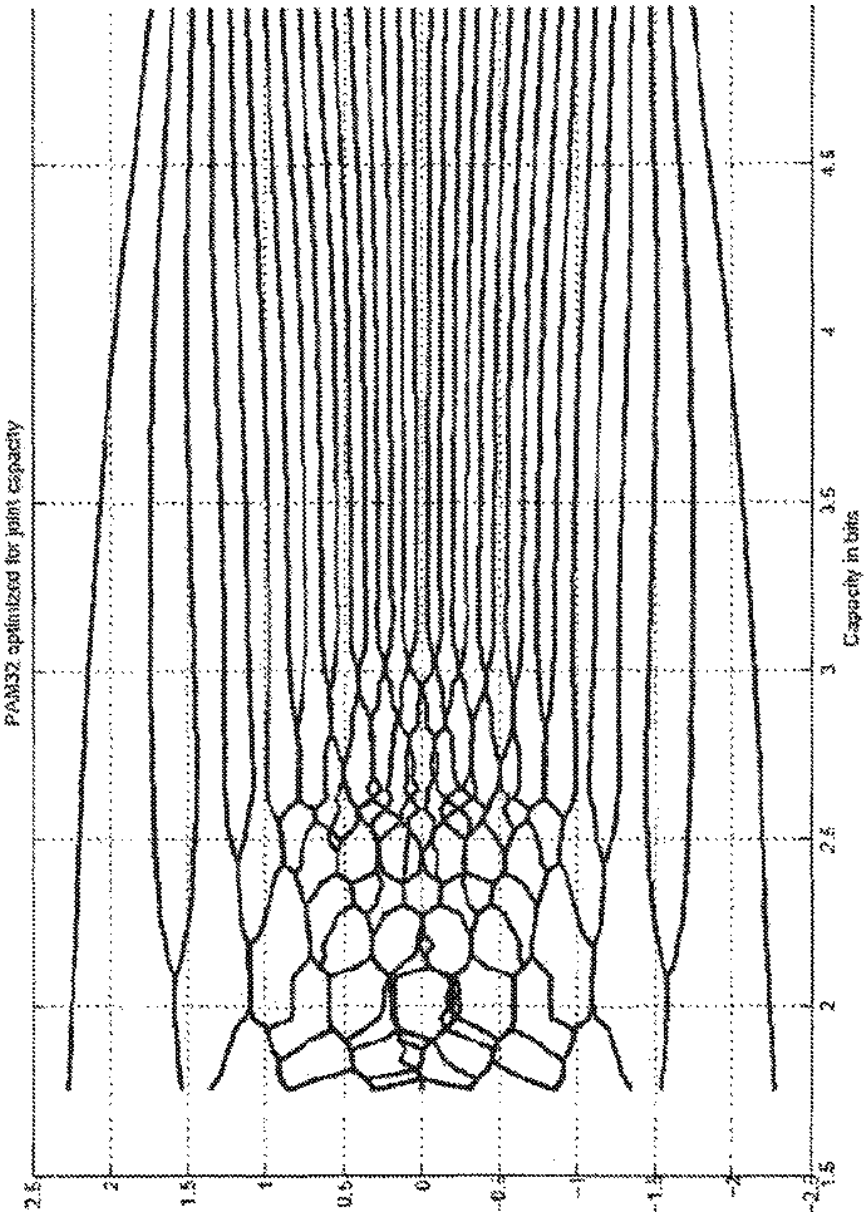

PAM32 optimized for joint capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16a

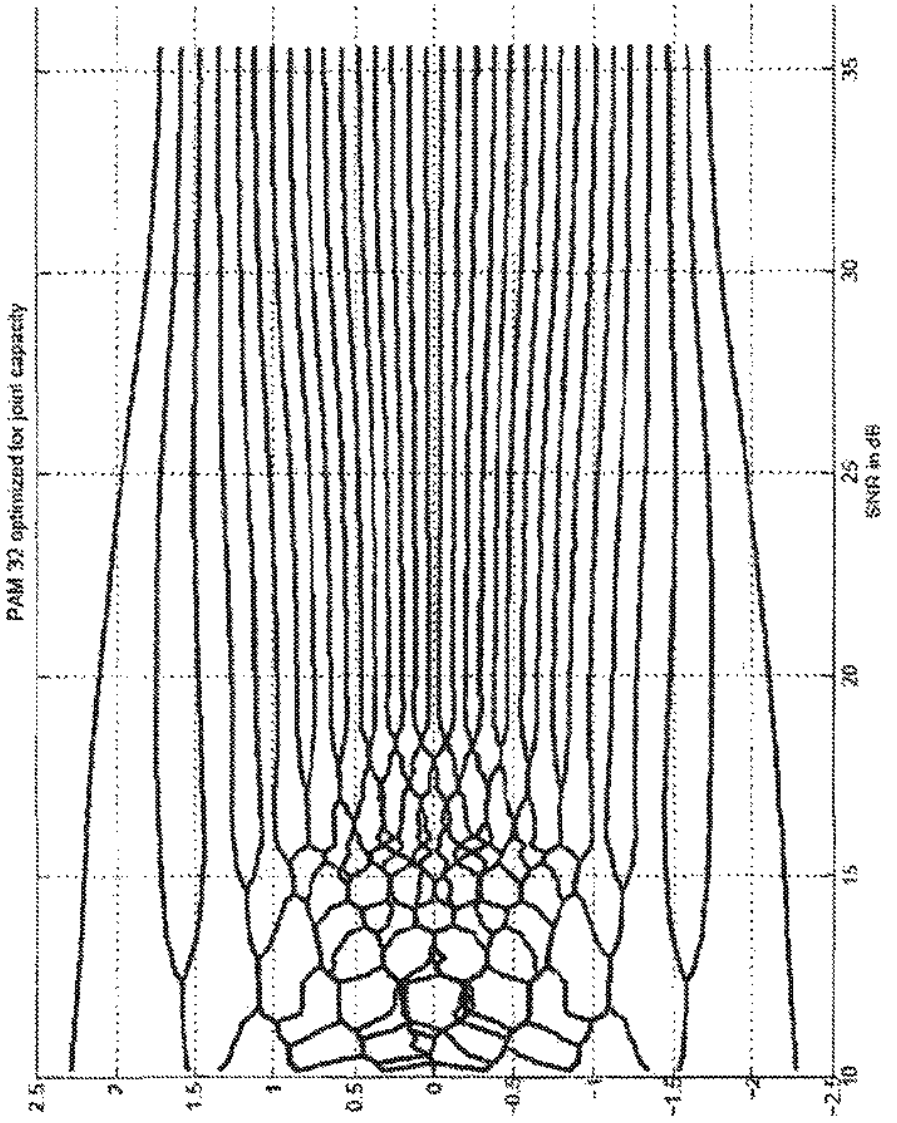

PAM 32 optimized for joint capacity
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
10
15
20
25
30
35

FIG. 16b

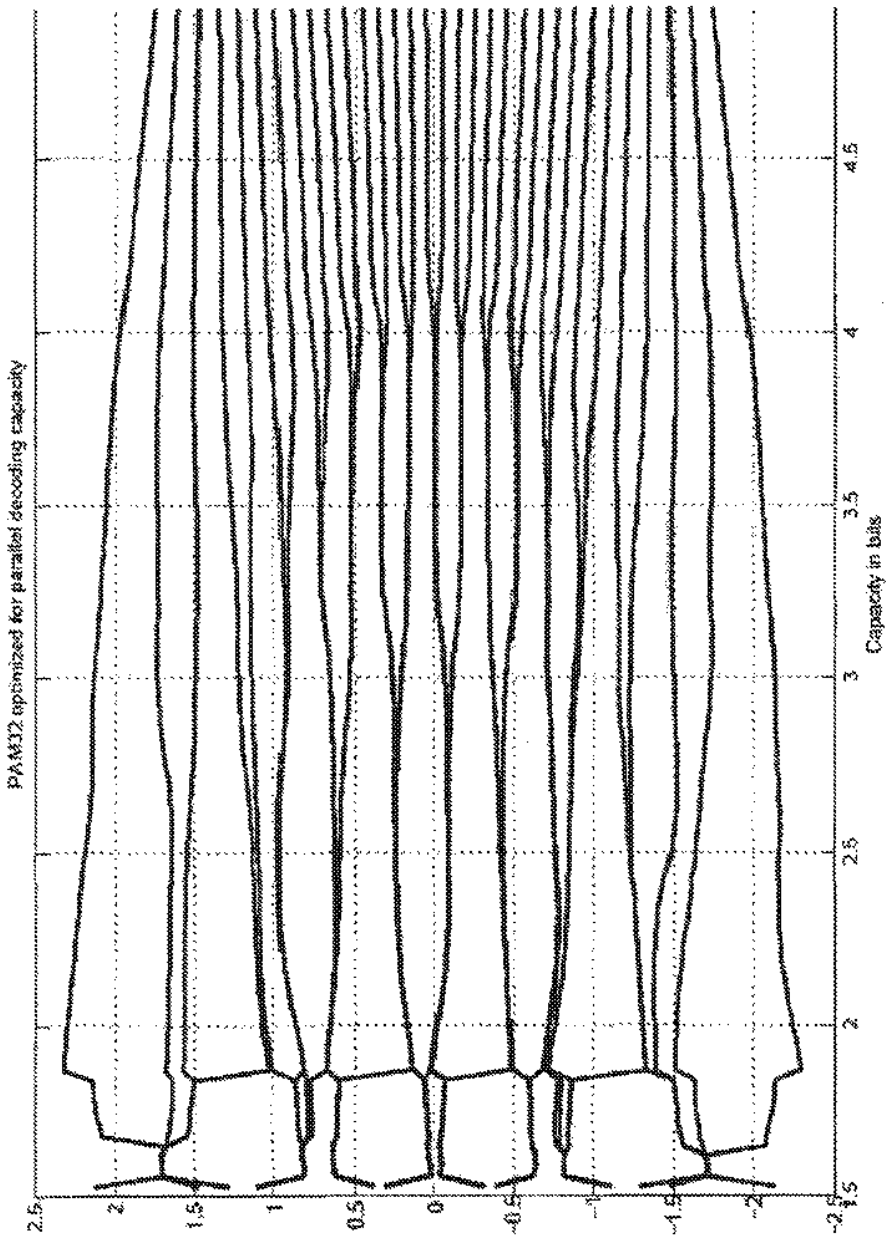

PAM32 optimized for parallel decoding capacity
Capacity in bits
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
1.5
2
2.5
3
3.5
4
4.5

FIG. 16c

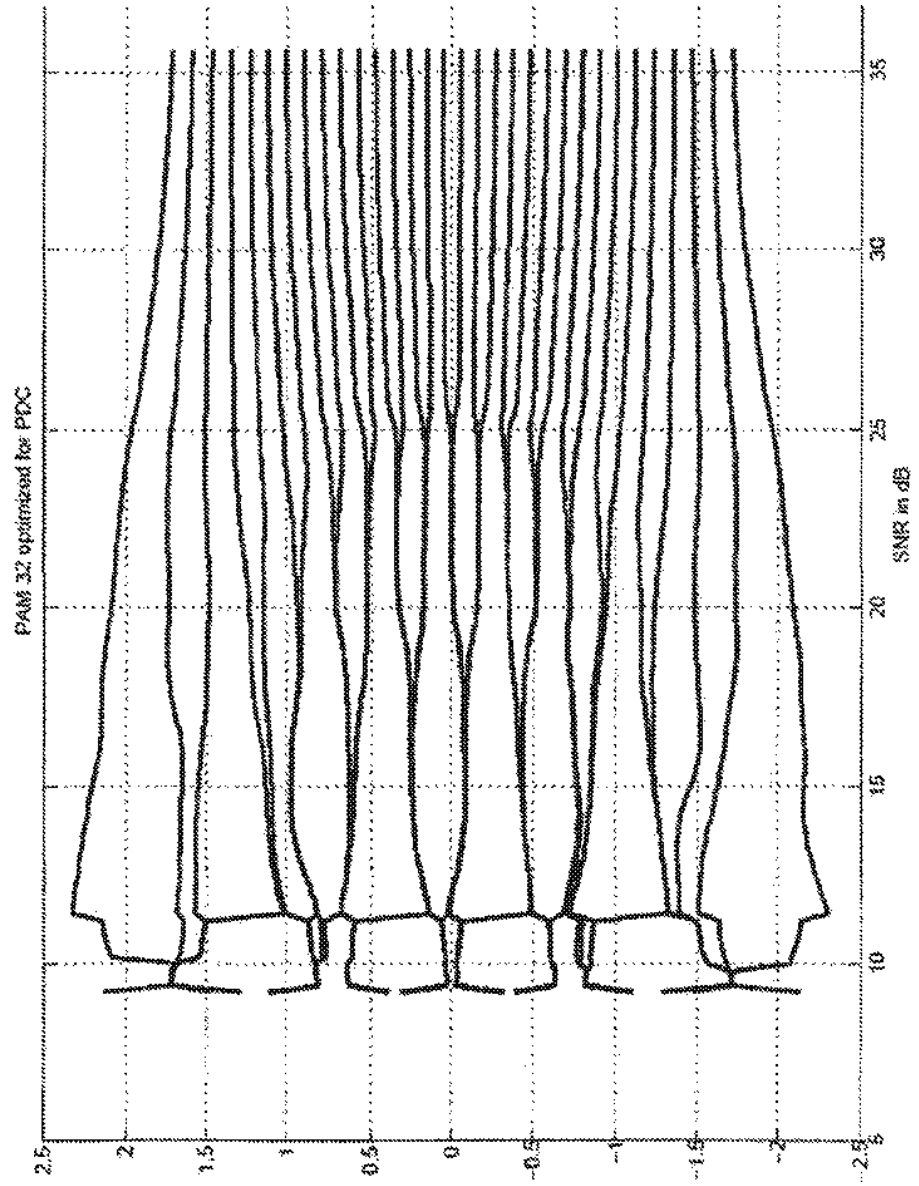

PAM 32 optimized for PDC
SNR in dB
2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
5
10
15
20
25
30
35

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

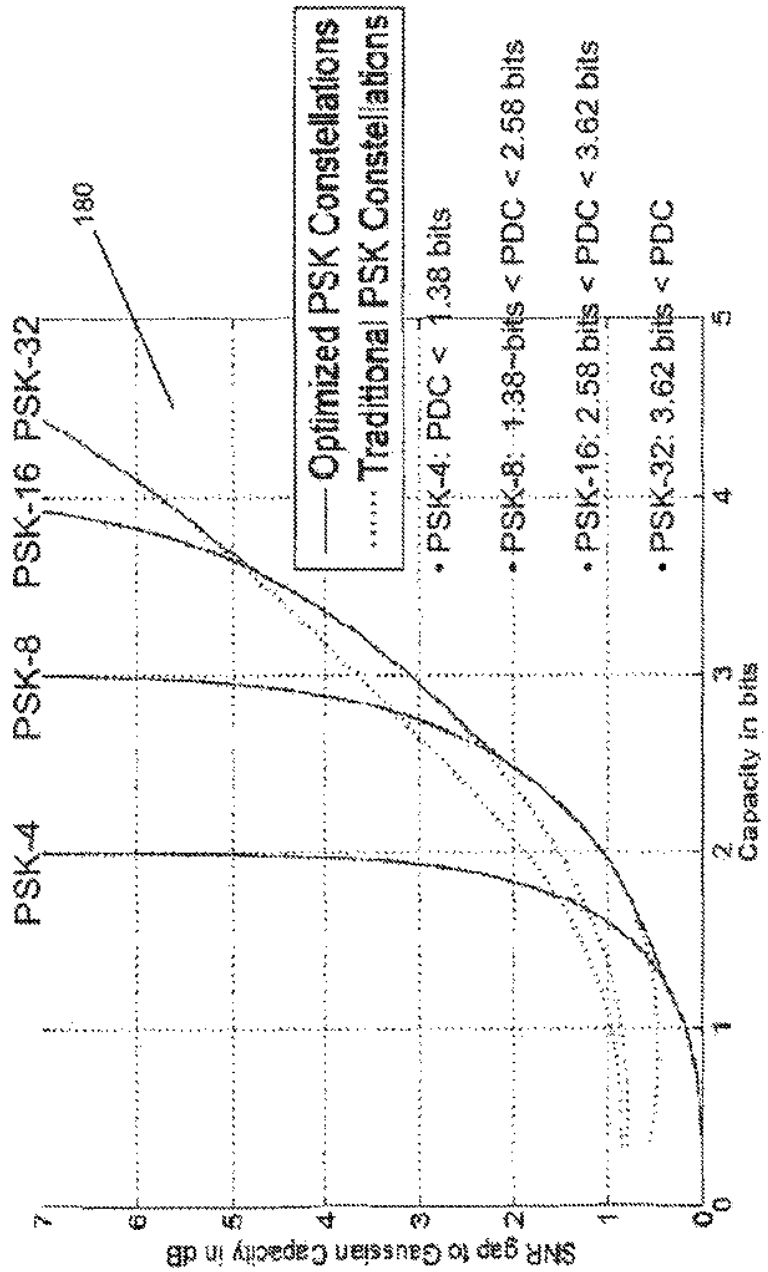

180
PSK-4
PSK-8
PSK-16
PSK-32
Optimized PSK Constellations
Traditional PSK Constellations
• PSK-4: PDC < 1.38 bits
• PSK-8: 1.38~bits < PDC < 2.58 bits
• PSK-16: 2.58 bits < PDC < 3.62 bits
• PSK-32: 3.62 bits < PDC
SNR gap to Gaussian Capacity in dB
Capacity in bits
0
1
2
3
4
5
6
7

FIG. 18

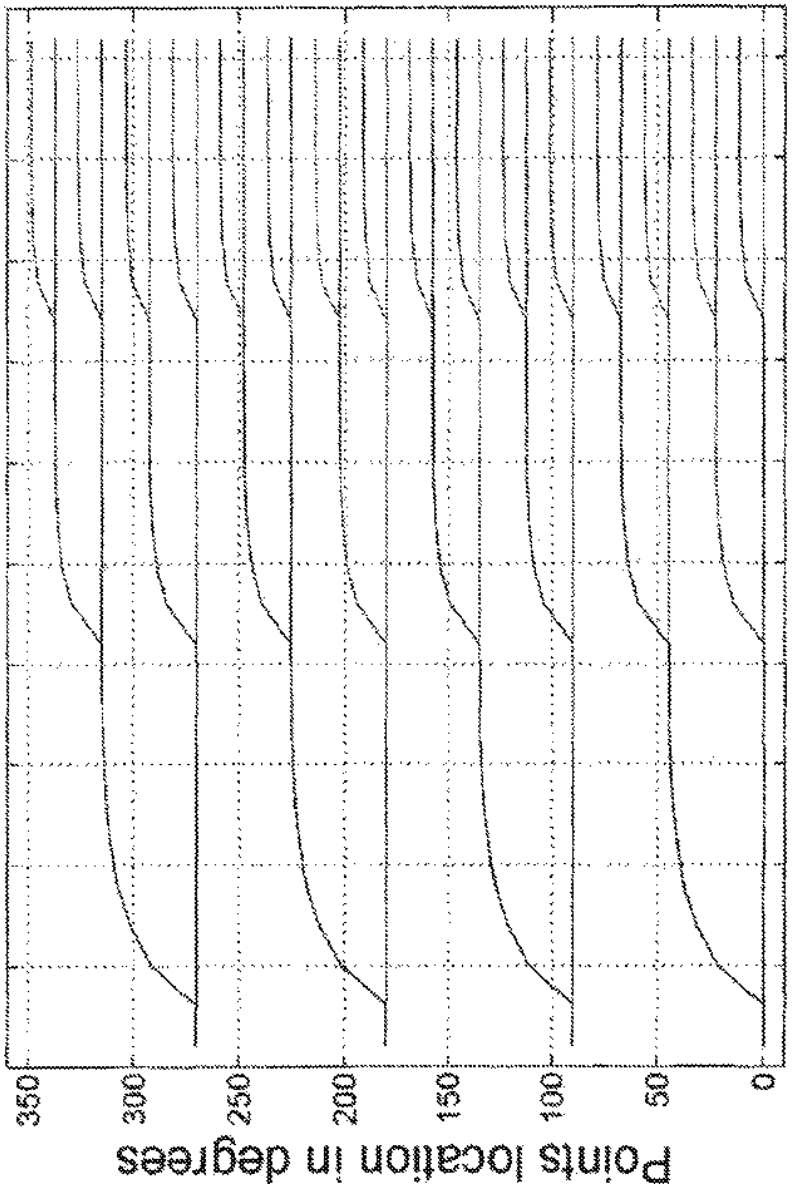

Points location in degrees
350
300
250
200
150
100
50
0
0
2
4
6
8
10
12
14
16
18
20
SNR in dB

FIG. 19

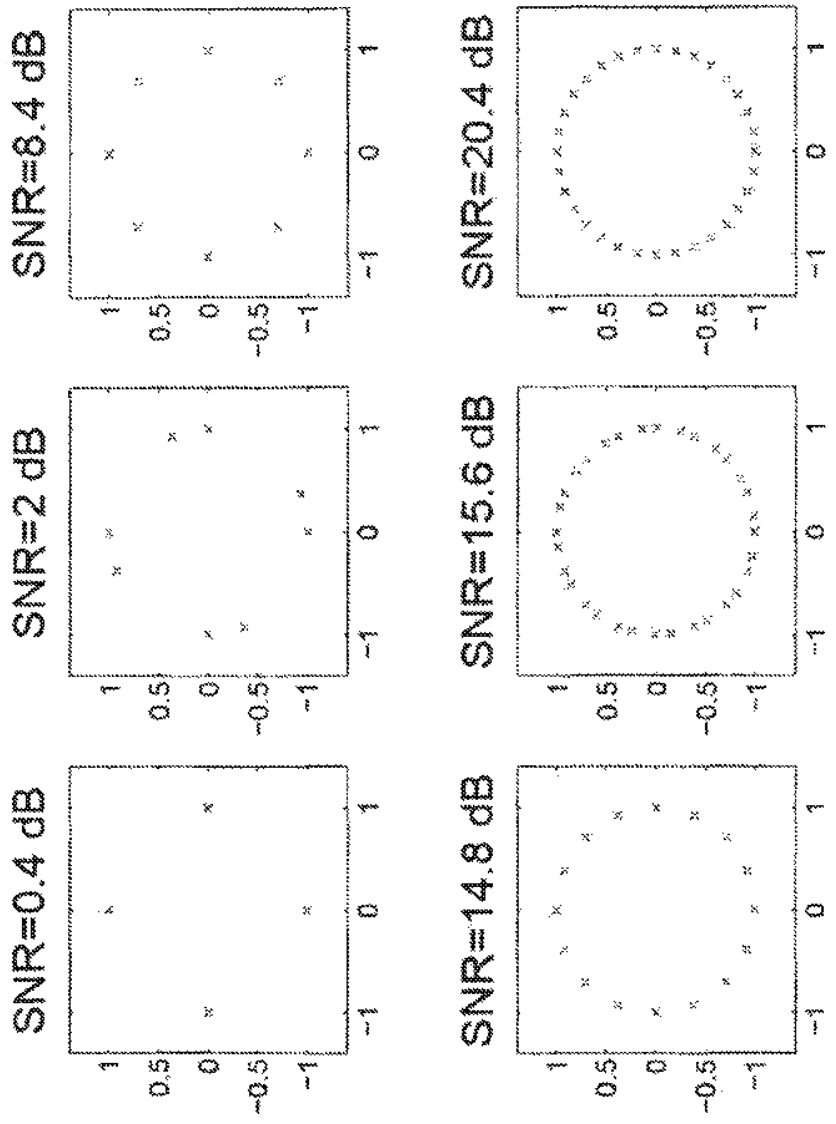


FIG. 20

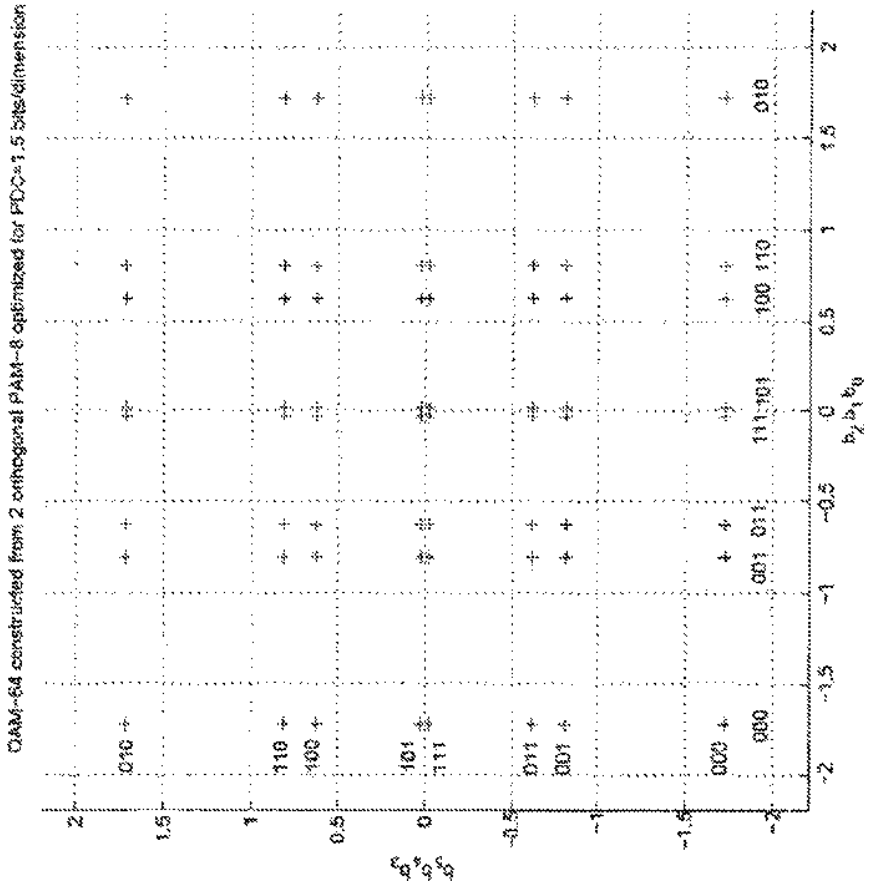

QAM-64 constructed from 2 orthogonal PAM-8 optimized for PDC=1.5 bits/dimension
010
110
100
101
111
011
001
000
010
110
100
101
111
011
001
000
$s_3 s_2 s_1 s_0$
$b_2 b_1 b_0$

FIG. 21

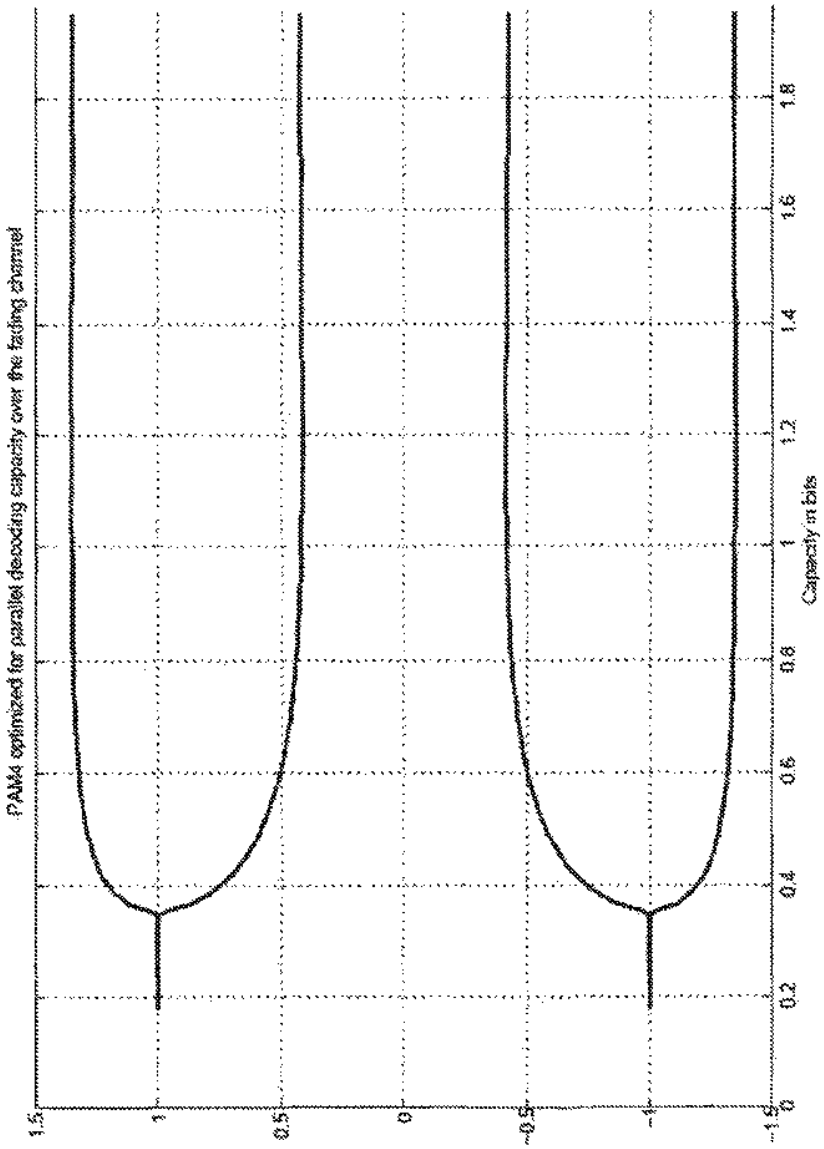

PAM4 optimized for parallel decoding capacity over the fading channel
Capacity in bits
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8

FIG. 22a

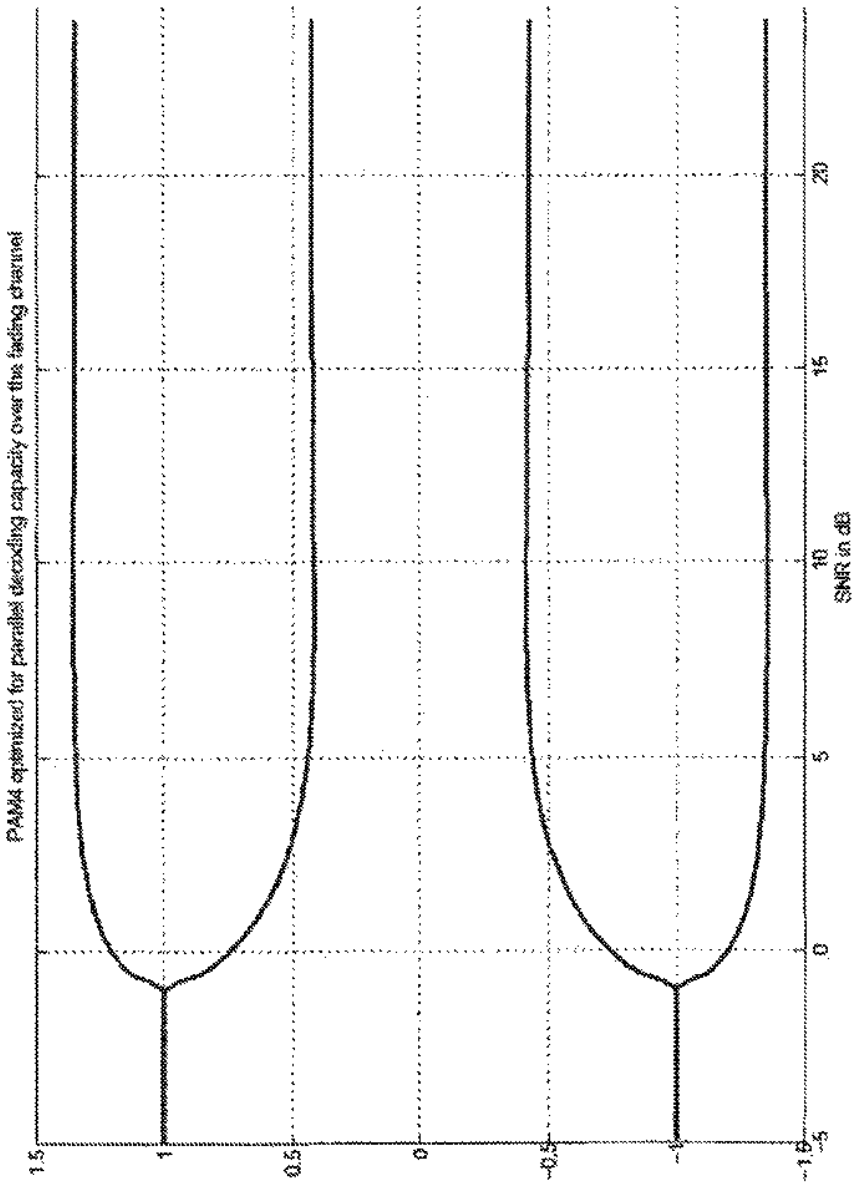

PAM4 optimized for parallel decoding capacity over the fading channel
1.5
1
0.5
0
-0.5
-1
-1.5
-5
0
5
10
15
20
SNR in dB

FIG. 22b

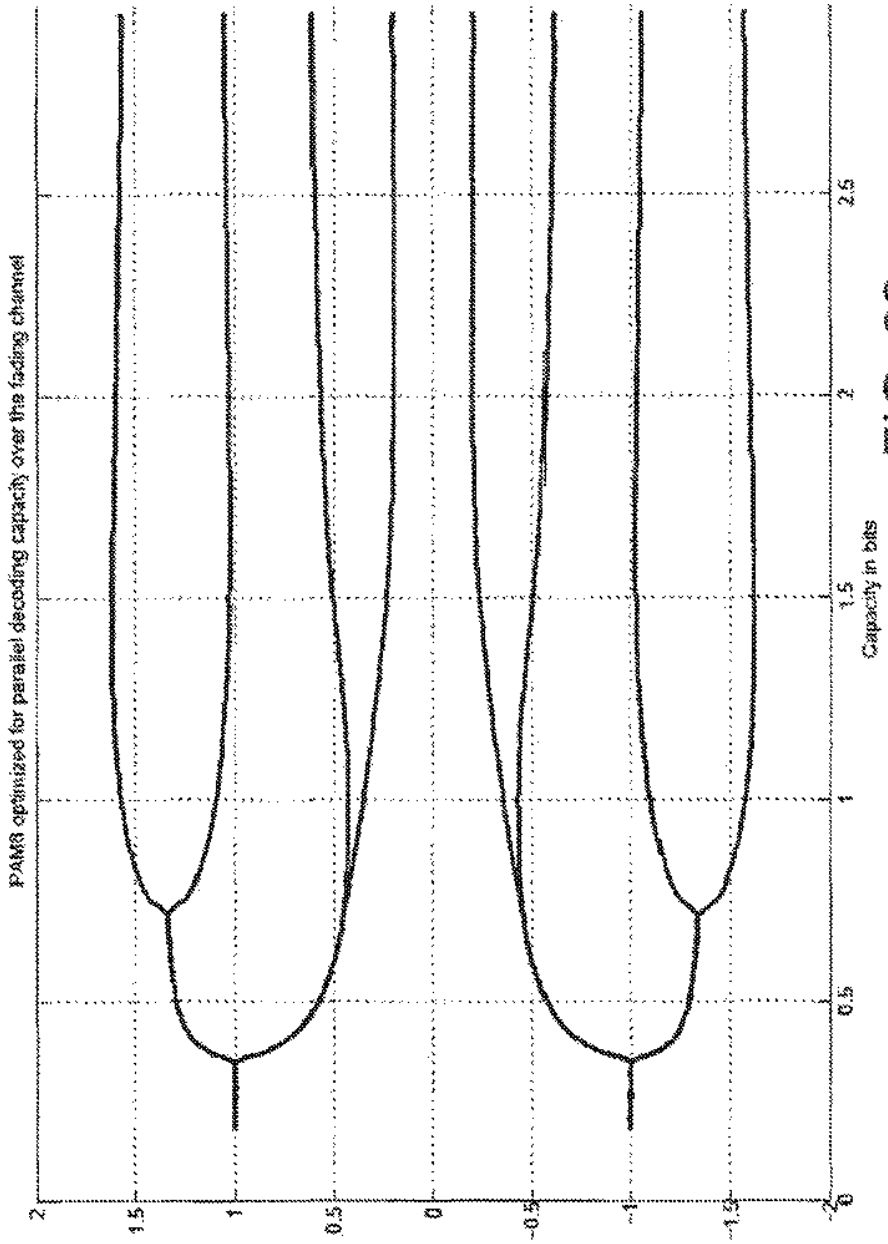

PAMR optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5

FIG. 23a

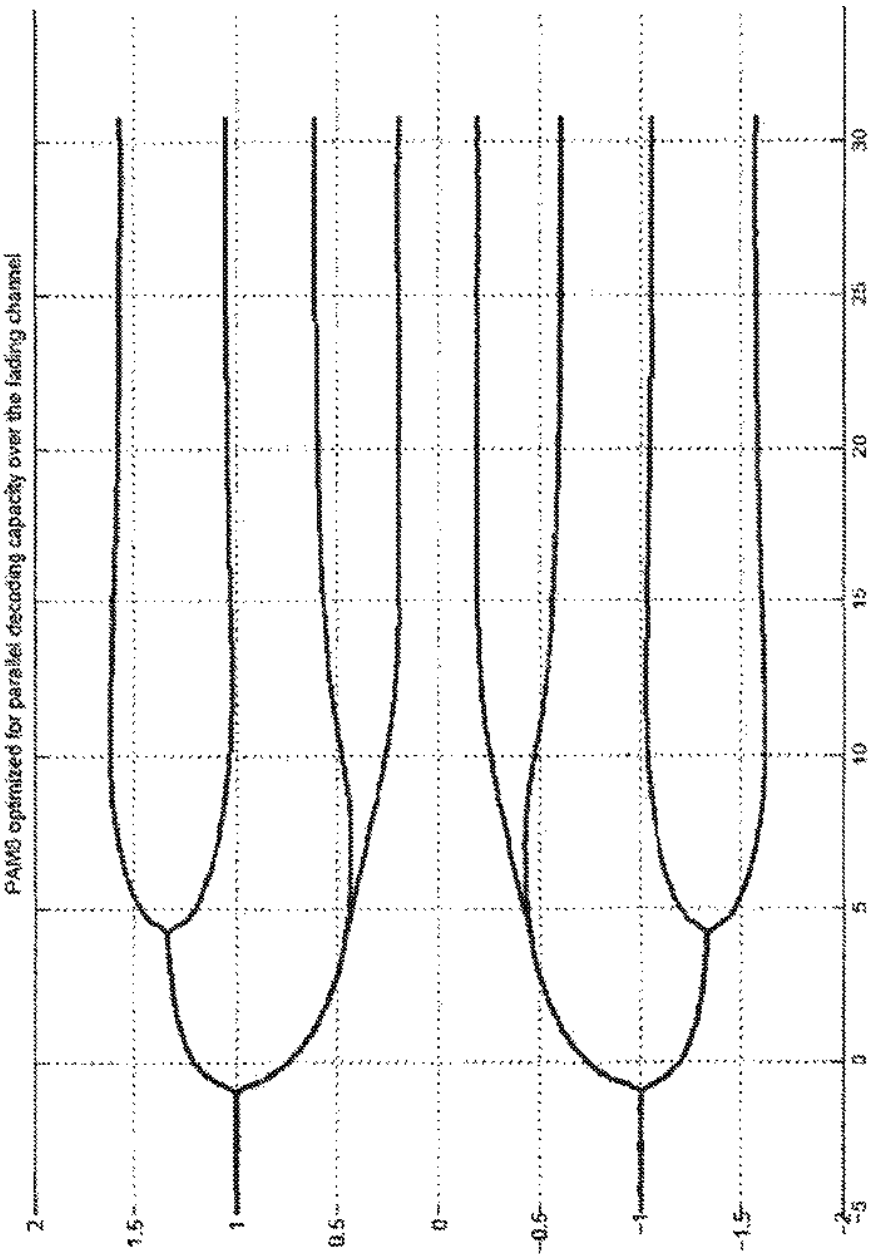

PAM8 optimized for parallel decoding capacity over the fading channel
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-5
0
5
10
15
20
25
30
SNR in dB

FIG. 23b

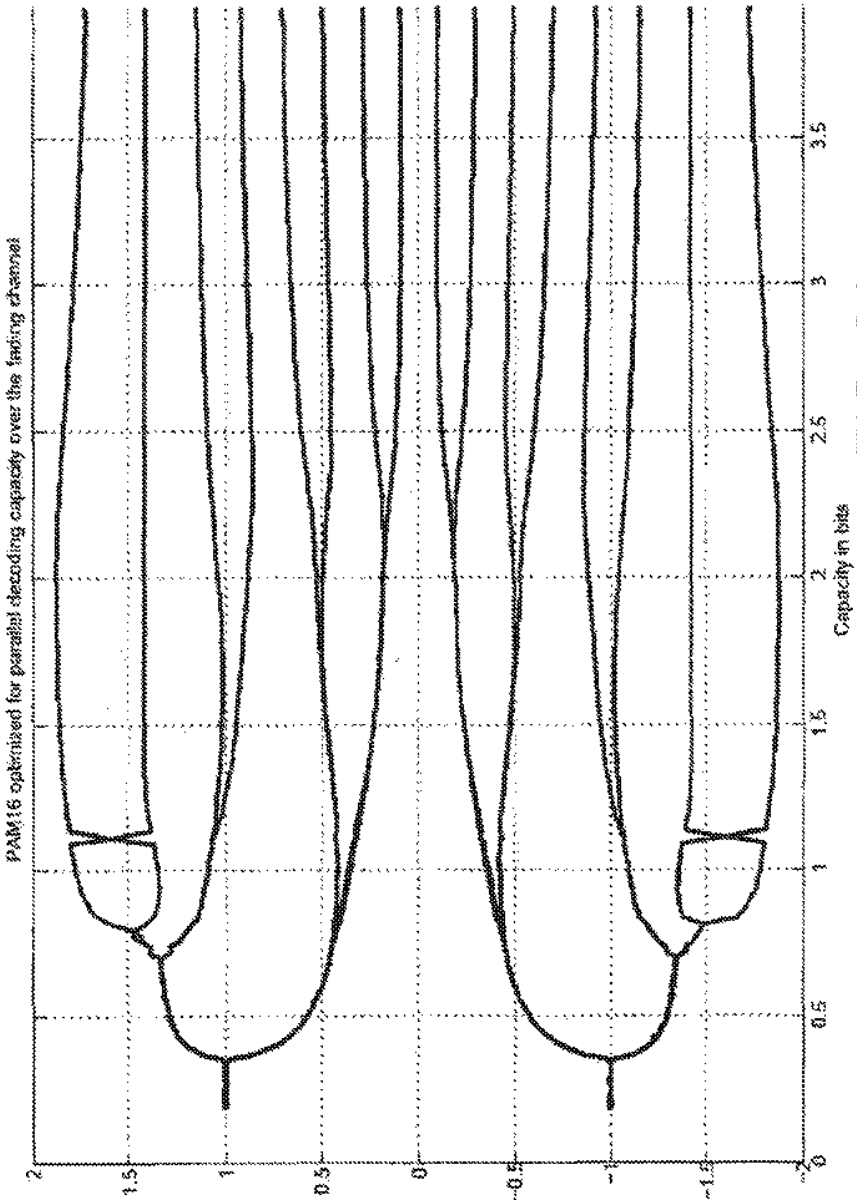

PAM16 optimized for parallel decoding capacity over the fading channel
Capacity in bits
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3
3.5

FIG. 24a

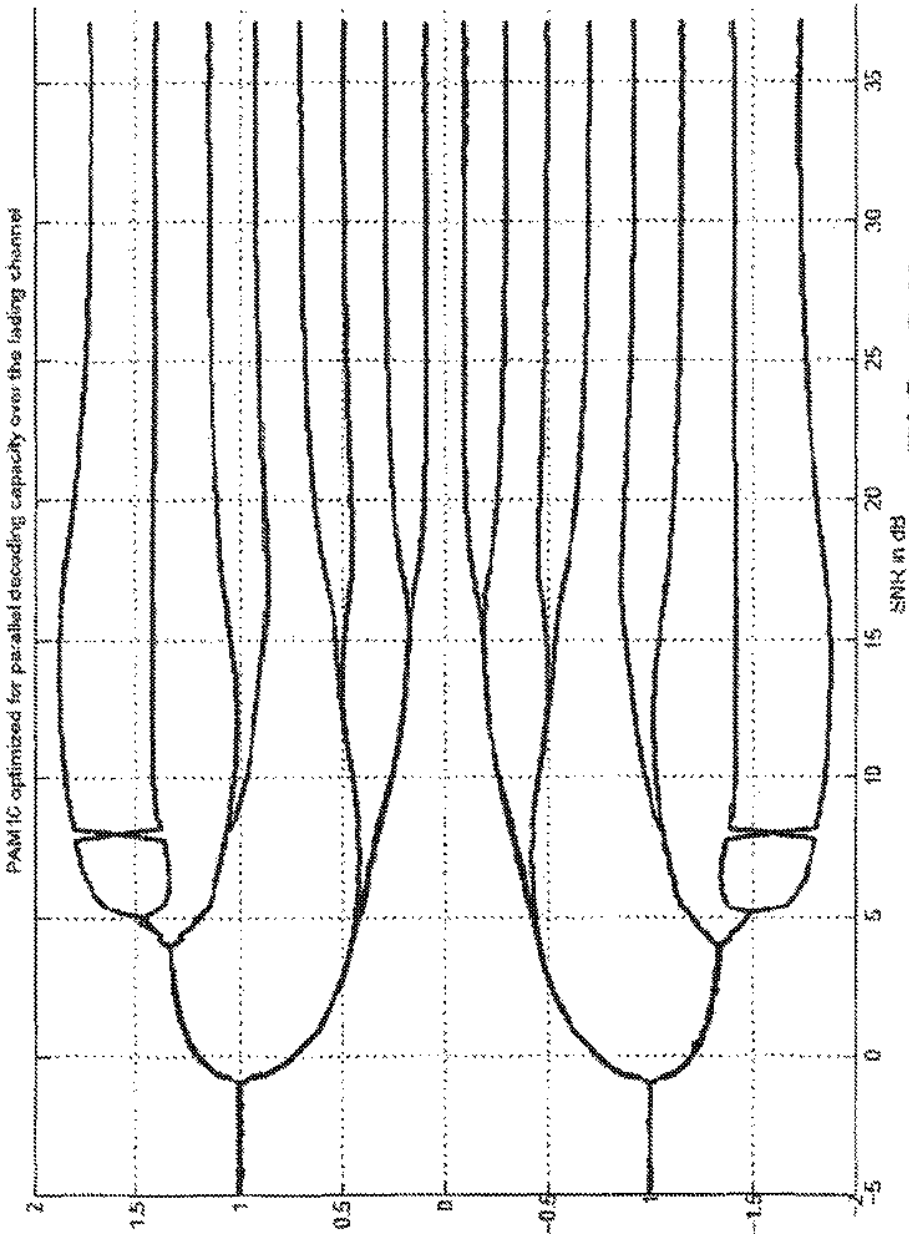

PAM10 optimized for parallel decoding capacity over the fading channel
SNR in dB
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
35
30
25
20
15
10
5
0
-5

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

EPO FORM 1503 03.82 (P04C01)



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	NAGUIB 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, XP031140978, 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
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D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P04C01)

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 August 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
August 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 "Foreign Patent Documents" and "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

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
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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 09/14/2020

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER
BOCURE, TESFALDET

ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
09/14/2020	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 07/09/2020.

☐ 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) 1,9-11,19-22 and 30 is/are rejected.

8) ☑ Claim(s) 2-8,12-18 and 23-29 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 _____ 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 (Final Office Action-) is in response to the amendment and remarks filed on 07/09/2020. Claims 5, 15 and 26 have been amended and the remaining claims 1-4, 6-14, 16-25 and 27-30 are as originally presented. The detail office action to the pending claims 1-30 are as shown below.

Request for Prioritized Examination (TrackOne)

3. The certification and request for prioritized examination (TrackOne) Under 37 CFR 1.102(e) filed on 12/27/2019 has been approved by the petition office on 02/04/2020 and also is noted by the Examiner.

Information Disclosure Statement

4. The information disclosure statements (IDSs) submitted on 07/10/2020 and 08/10/2020 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.

Response to Amendment/Remarks

6. In response to the amendments and Applicant's remarks received on 07/09/2020, the following actions have been taken:

- The objection to the specification for minor informality indicated in the non-final office action mailed on 04/09/2020 is withdrawn.
- The rejections to Claims 22-30 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 indicated in the non-final office action mailed on 04/09/2020 is withdrawn.
- After reviewing Applicant's remarks and further revisiting the MPEP for "Undue Breath Rejection in section 2173.04 and 2111.01 (II)", the rejection to claims 1-30 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 (single means rejection) indicated in the non-final office action mailed on 04/09/2020 is withdrawn.

7. In response to Applicant's argument with respect to the rejections of claims 1, 10, 11, 20, 21 and 22 Under 35 U.S.C. §102(e) over Eberlein that:

> 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 vl .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----:
>
> 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 in view of Eberlein must be withdrawn.

Examiner kindly disagree to the Applicant's assertion that the disclosed signal constellation points in the same quadrant shown in figure 3A in the system of Eberlein are not the same as the claimed "constellation points having identical locations and different labels" in the independent claims of the instant application. Under broad and reasonable interpretation of the claimed subject matter, without bringing the specification in to the claim (see MPEP 2111.01 (II)), the signal points in figure 3A in the system of Eberlein are located in the same quadrant having labelling 0010, 0011, 0001 and 0000. The first two digits 00 defines the location of the quadrant where the encoded bits are located, the same quadrant which is equivalent to the claimed same location, and the remaining different logical representation of 1's and 0's representing the encoded bits having different logical representations of 10, 11, 01, 00, claimed different labelling. Under a broadest reasonable interpretation (BRI), words of the claim (constellation points having identical locations and different labels) must be given their plain meaning, unless such meaning is inconsistent with the specification, therefore, Examiner

still believes that the claimed identical location is equivalent to the identical quadrant shown in in figure 3A of Eberlein.

8. In response to Applicant's argument with respect to the rejections of claims 9,19 and 30 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).that:

> 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).

Regarding Applicant argument to the rejection of claims 9, 19 and 30, the same response applies to those claims as applied to the independent claims 1, 11 and 22 indicated above in par. 7.

Claim Rejections - 35 USC § 102

9. 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. (US patent publication number 2008/0200114-A1, Eberlein hereinafter).

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

10. 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. (US patent publication number 2008/0200114-A1, Eberlein hereinafter) in view of Walker et al. (US patent publication number 2005/011581-A1, Walker hereinafter).

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.

Allowable Subject Matter

13. Claims 2-8, 12-18 and 22-28 are objected to as being dependent upon a rejected base claim, but would be allowable if rewritten in independent form including all of the limitations of the base claim and any intervening claims.

Conclusion

14. **THIS ACTION IS MADE FINAL.** Applicant is reminded of the extension of time policy as set forth in 37 CFR 1.136(a).

A shortened statutory period for reply to this final action is set to expire THREE MONTHS from the mailing date of this action. In the event a first reply is filed within TWO MONTHS of the mailing date of this final action and the advisory action is not mailed until after the end of the THREE-MONTH shortened statutory period, then the shortened statutory period will expire on the date the advisory action is mailed, and any extension fee pursuant to 37 CFR 1.136(a) will be calculated from the mailing date of the advisory action. In no event, however, will the statutory period for reply expire later than SIX MONTHS from the mailing date of this final action.

15. 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

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 search updated	09/09/2020	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
375, 455, 329, 332, 370	EAST and WEST search updated ALL ($8.ccls.)	09/09/2020	TB

Search Notes	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

* See search history printout included with this form or the SEARCH NOTES box below to determine the scope of the search.

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 search updated	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
EAST and WEST search updated	09/09/2020	TB

Interference Search			
US Class/CPC Symbol	**US Subclass/CPC Group**	**Date**	**Examiner**

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	09/09/2020							
	1	✓	✓							
	2	✓	O							
	3	✓	O							
	4	✓	O							
	5	✓	O							
	6	✓	O							
	7	✓	O							
	8	✓	O							
	9	✓	✓							
	10	✓	✓							
	11	✓	✓							
	12	✓	O							
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	14	✓	O							
	15	✓	O							
	16	✓	O							
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	24	✓	O							
	25	✓	O							
	26	✓	O							
	27	✓	O							
	28	✓	O							
	29	✓	O							
	30	✓	✓							

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.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

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.	

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
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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	/TESFALDET BOCURE/	Date Considered	09/05/2020

*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: 2020090920:45

Prior Art Searches

Query	DB	Hits	Op.	Plur.	Thes.	Date
((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	330	OR	YES		09-09-2020
((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	19	OR	YES		09-09-2020
(((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)	PGPB, USPT, USOC, EPAB, JPAB, DWPI, TDBD, FPRS	374	OR	YES		09-09-2020
(((((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		09-09-2020
(((((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	PGPB, USPT, USOC, EPAB,	22	OR	YES		09-09-2020

amplitude adj2 modulat$4)))) and @ad<=20070605)	JPAB, DWPI, TDBD, FPRS					
(((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	19	OR	YES		09-09-2020
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((((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	45	OR	YES		09-09-2020
((((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or	PGPB, USPT,	359	OR	YES		09-09-2020

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.	USOC, EPAB, JPAB, DWPI, TDBD, FPRS					
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"20050268206" | "20060085720" | "20060165190" | "20070211822" | "20200137595" | "20200145276" | "20200145844" | "20200145845" | "20200145846" | "20200145848" | "20200145849" | "20200145850" | "20200162942" | "20050141627" | "20050143004" | "20050169400" | "20060276145" | "20100077275" | "20200221321" | "7123663" | "7359426" | "7539261" | "7599420" | "7907641" | "7908541" | "8031793" | "8102947" | "8111770" | "8160121" | "8171383" | "8199847" | "8428162" | "8483145").PN.	USOC, EPAB, JPAB, DWPI, TDBD, FPRS					

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	2020/04/06 10:49
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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L3	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
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	2020/04/06 10:49
L5	7	L4 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L6	328	L4 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

L7	4	L6 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L8	5	L6 and L1	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
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	2020/04/06 10:49
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	2020/04/06 10:49
L11	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	2020/04/06 10:49
L12	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L13	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L14	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	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

		amplitude near3 modulation))) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 (codes or code))				
L15	7	L14 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L16	328	L14 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L17	4	L16 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L18	5	L16 and L11	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L19	4	L16 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	2020/04/06 10:49
L20	43	L16 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	2020/04/06 10:49
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		"6084915" | "6115415" | "6157678" | "6603801" | "6606355" | "6611554" | "6665831" | "7212582" | "7245666" | "7376203" | "7620067" | "7978777" | "8265175" | "8270511" | "8842761" | "9191148" | "9385832" | "9743290" | "9743292" | "9887870" | "9967127").PN.				
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L23	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	2020/04/06 10:49
L24	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L25	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L26	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	2020/04/06 10:49
L27	7	L26 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L28	328	L26 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2020/04/06 10:49

			EPO; JPO; DERWENT; IBM_TDB			
L29	4	L28 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L30	5	L28 and L23	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L31	4	L28 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	2020/04/06 10:49
L32	43	L28 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	2020/04/06 10:49
L33	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	2020/04/06 10:49
L34	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L35	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L36	421	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3	US-PGPUB; USPAT; USOCR;	OR	ON	2020/04/06 10:49

		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))	FPRS; EPO; JPO; DERWENT; IBM_TDB			
L37	7	L36 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L38	328	L36 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L39	4	L38 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L40	5	L38 and L33	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L41	4	L38 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	2020/04/06 10:49
L42	43	L38 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	2020/04/06 10:49
L43	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" |	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

		"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.				
L44	4	L38 and L43	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L45	374	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
L46	67	L45 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S1	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 20:28
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
S12	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
S13	0	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or	US-PGPUB; USPAT;	OR	ON	2020/04/05 22:12

		(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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			
S14	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) 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
S15	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
S16	7	S15 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S17	328	S15 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
S18	4	S17 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S19	5	S17 and S12	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S20	4	S17 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;	OR	ON	2020/04/05 22:12

			IBM_TDB			
S21	43	S17 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
S22	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
S23	4	S17 and S22	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:14
S24	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	2020/04/06 10:08
S25	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	2020/04/06 10:08
S26	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	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2020/04/06 10:08

		near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	EPO; JPO; DERWENT; IBM_TDB			
S27	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	2020/04/06 10:08
S28	7	S27 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S29	328	S27 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S30	4	S29 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S31	5	S29 and S24	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S32	4	S29 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	2020/04/06 10:08
S33	43	S29 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	2020/04/06 10:08
S34	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;	OR	ON	2020/04/06 10:08

			DERWENT; IBM_TDB			
S35	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S36	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S37	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	2020/04/06 10:08
S38	7	S37 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S39	328	S37 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S40	4	S39 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S41	5	S39 and S34	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S42	4	S39 and (375/261.ccls. or 375/269.ccls. or	US-PGPUB;	OR	ON	2020/04/06

		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.)	USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			10:08
S43	43	S39 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	2020/04/06 10:08
S44	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	2020/04/06 10:08
S45	4	S39 and S44	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S46	374	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:27
S47	67	S46 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:28

4/6/2020 10:55:05 AM
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	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.	

ALL REFERENCES CONSIDERED EXCEPT WHERE LINED THROUGH. /T.B./

Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
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	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.		

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
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	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	/TESFALDET BOCURE/	Date Considered	09/05/2020

*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./

UNITED STATES PATENT AND TRADEMARK OFFICE

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
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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 01/19/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER
BOCURE, TESFALDET

ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
01/19/2021	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

Applicant-Initiated Interview Summary	**Application No.** 16/728,384	**Applicant(s)** Barsoum et al.	
	Examiner TESFALDET BOCURE	**Art Unit** 2633	**AIA (FITF) Status** No

All participants (applicant, applicants representative, PTO personnel):

(1) TESFALDET BOCURE. (3) _____.

(2) Mr. David Baily and Invnetor. (4) _____.

Date of Interview: 12 January 2021.

Type: ☑ Telephonic ☐ Video Conference
☐ Personal [copy given to: ☐ applicant ☐ applicant's representative]

Exhibit shown or demonstration conducted: ☐ Yes ☑ No.
If Yes, brief description: _____.

Issues Discussed ☐101 ☐112 ☑102 ☐103 ☐Others
(For each of the checked box(es) above, please describe below the issue and detailed description of the discussion)

Claim(s) discussed: 1,11 and 22.

Identification of prior art discussed: Eberlein (US patent publication No. 2008/.0200114-A1).

Substance of Interview
(For each issue discussed, provide a detailed description and indicate if agreement was reached. Some topics may include: identification or clarification of a reference or a portion thereof, claim interpretation, proposed amendments, arguments of any applied references etc...)

See Continuation Sheet.

Applicant recordation instructions: The formal written reply to the last Office action must include the substance of the interview. (See MPEP section 713.04). If a reply to the last Office action has already been filed, applicant is given a non-extendable period of the longer of one month or thirty days from this interview date, or the mailing date of this interview summary form, whichever is later, to file a statement of the substance of the interview.

Examiner recordation instructions: Examiners must summarize the substance of any interview of record. A complete and proper recordation of the substance of an interview should include the items listed in MPEP 713.04 for complete and proper recordation including the identification of the general thrust of each argument or issue discussed, a general indication of any other pertinent matters discussed regarding patentability and the general results or outcome of the interview, to include an indication as to whether or not agreement was reached on the issues raised.

☐ Attachment

/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	

Summary of Record of Interview Requirements

Manual of Patent Examining Procedure (MPEP), Section 713.04, Substance of Interview Must be Made of Record
A complete written statement as to the substance of any face-to-face, video conference, or telephone interview with regard to an application must be made of record in the application whether or not an agreement with the examiner was reached at the interview.

Title 37 Code of Federal Regulations (CFR) 1.133 Interviews
Paragraph (b)

In every instance where reconsideration is requested in view of an interview with an examiner, a complete written statement of the reasons presented at the interview as warranting favorable action must be filed by the applicant. An interview does not remove the necessity for reply to Office action as specified in §§ 1.111, 1.135. (35 U.S.C. 132)

37 CFR §1.2 Business to be transacted in writing.
All business with the Patent or Trademark Office should be transacted in writing. The personal attendance of applicants or their attorneys or agents at the Patent and Trademark Office is unnecessary. The action of the Patent and Trademark Office will be based exclusively on the written record in the Office. No attention will be paid to any alleged oral promise, stipulation, or understanding in relation to which there is disagreement or doubt.

The action of the Patent and Trademark Office cannot be based exclusively on the written record in the Office if that record is itself incomplete through the failure to record the substance of interviews.

It is the responsibility of the applicant or the attorney or agent to make the substance of an interview of record in the application file, unless the examiner indicates he or she will do so. It is the examiners responsibility to see that such a record is made and to correct material inaccuracies which bear directly on the question of patentability.

Examiners must complete an Interview Summary Form for each interview held where a matter of substance has been discussed during the interview by checking the appropriate boxes and filling in the blanks. Discussions regarding only procedural matters, directed solely to restriction requirements for which interview recordation is otherwise provided for in Section 812.01 of the Manual of Patent Examining Procedure, or pointing out typographical errors or unreadable script in Office actions or the like, are excluded from the interview recordation procedures below. Where the substance of an interview is completely recorded in an Examiners Amendment, no separate Interview Summary Record is required.

The Interview Summary Form shall be given an appropriate Paper No., placed in the right hand portion of the file, and listed on the "Contents" section of the file wrapper. In a personal interview, a duplicate of the Form is given to the applicant (or attorney or agent) at the conclusion of the interview. In the case of a telephone or video-conference interview, the copy is mailed to the applicants correspondence address either with or prior to the next official communication. If additional correspondence from the examiner is not likely before an allowance or if other circumstances dictate, the Form should be mailed promptly after the interview rather than with the next official communication.

The Form provides for recordation of the following information:
- Application Number (Series Code and Serial Number)
- Name of applicant
- Name of examiner
- Date of interview
- Type of interview (telephonic, video-conference, or personal)
- Name of participant(s) (applicant, attorney or agent, examiner, other PTO personnel, etc.)
- An indication whether or not an exhibit was shown or a demonstration conducted
- An identification of the specific prior art discussed
- An indication whether an agreement was reached and if so, a description of the general nature of the agreement (may be by attachment of a copy of amendments or claims agreed as being allowable). Note: Agreement as to allowability is tentative and does not restrict further action by the examiner to the contrary.
- The signature of the examiner who conducted the interview (if Form is not an attachment to a signed Office action)

It is desirable that the examiner orally remind the applicant of his or her obligation to record the substance of the interview of each case. It should be noted, however, that the Interview Summary Form will not normally be considered a complete and proper recordation of the interview unless it includes, or is supplemented by the applicant or the examiner to include, all of the applicable items required below concerning the substance of the interview.

A complete and proper recordation of the substance of any interview should include at least the following applicable items:
1) A brief description of the nature of any exhibit shown or any demonstration conducted,-
2) an identification of the claims discussed,
3) an identification of the specific prior art discussed,
4) an identification of the principal proposed amendments of a substantive nature discussed, unless these are already described on the Interview Summary Form completed by the Examiner,
5) a brief identification of the general thrust of the principal arguments presented to the examiner,
(The identification of arguments need not be lengthy or elaborate. A verbatim or highly detailed description of the arguments is not required. The identification of the arguments is sufficient if the general nature or thrust of the principal arguments made to the examiner can be understood in the context of the application file. Of course, the applicant may desire to emphasize and fully describe those arguments which he or she feels were or might be persuasive to the examiner.)
6) a general indication of any other pertinent matters discussed, and
7) if appropriate, the general results or outcome of the interview unless already described in the Interview Summary Form completed by the examiner.

Examiners are expected to carefully review the applicants record of the substance of an interview. If the record is not complete and accurate, the examiner will give the applicant an extendable one month time period to correct the record.

Examiner to Check for Accuracy

If the claims are allowable for other reasons of record, the examiner should send a letter setting forth the examiners version of the statement attributed to him or her. If the record is complete and accurate, the examiner should place the indication, Interview Record OK on the paper recording the substance of the interview along with the date and the examiners initials.

Continuation of Substance of Interview including description of the general nature of what was agreed to if an agreement was reached, or any other comments: The interview conducted with the Examiner on 01/12/2021 and in the presence of Dr. Chris Jones (inventor), Mr. David Loo (representative of Constellation Designs) and Mr. David Bailey (counsel for Constellation Designs), the proposed interview agenda as shown and disclosed in fig. 13b and claimed limitation in the independent claims 1, 11 and 22, specifically 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” and was discussed.
Examiner still believes that the prior art of record Eberlein (US patent publication number 2008/ .0200114-A1) under broad and reasonable interpretation of the claimed subject matter reads on the claimed limitations as indicated in the final office action mailed on 09/14/2020. Applicant and his representatives will send a proposed amendment to reflect to the subject matter disclosed in figure 13b and Examiner agreed to look at it.

PTO/SB/30EFS (02-18)
Approved for use through 11/30/2020. 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.

REQUEST FOR CONTINUED EXAMINATION(RCE)TRANSMITTAL
(Submitted Only via EFS-Web)

Application Number	16728384	Filing Date	2019-12-27	Docket Number (if applicable)	C44-00999.CON15	Art Unit	2633
First Named Inventor	Maged F. Barsoum			Examiner Name	Tesfaldet Bocure		

This is a Request for Continued Examination (RCE) under 37 CFR 1.114 of the above-identified application.
Request for Continued Examination (RCE) practice under 37 CFR 1.114 does not apply to any utility or plant application filed prior to June 8, 1995, or to any design application. The Instruction Sheet for this form is located at WWW.USPTO.GOV

SUBMISSION REQUIRED UNDER 37 CFR 1.114

Note: If the RCE is proper, any previously filed unentered amendments and amendments enclosed with the RCE will be entered in the order in which they were filed unless applicant instructs otherwise. If applicant does not wish to have any previously filed unentered amendment(s) entered, applicant must request non-entry of such amendment(s).

☐ Previously submitted. If a final Office action is outstanding, any amendments filed after the final Office action may be considered as a submission even if this box is not checked.

- ☐ Consider the arguments in the Appeal Brief or Reply Brief previously filed on _____
- ☐ Other _____

☒ Enclosed

- ☒ Amendment/Reply
- ☐ Information Disclosure Statement (IDS)
- ☐ Affidavit(s)/ Declaration(s)
- ☒ Other Request for Track 1, Petition for Extention of Time

MISCELLANEOUS

☐ Suspension of action on the above-identified application is requested under 37 CFR 1.103(c) for a period of months _____ (Period of suspension shall not exceed 3 months; Fee under 37 CFR 1.17(i) required)

☐ Other _____

FEES

☒ **The RCE fee under 37 CFR 1.17(e) is required by 37 CFR 1.114 when the RCE is filed.**
The Director is hereby authorized to charge any underpayment of fees, or credit any overpayments, to Deposit Account No 504407

SIGNATURE OF APPLICANT, ATTORNEY, OR AGENT REQUIRED

☒ Patent Practitioner Signature

☐ Applicant Signature

PTO/SB/30EFS (02-18)
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U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

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Signature of Registered U.S. Patent Practitioner			
Signature	/David J. Bailey/	Date (YYYY-MM-DD)	2021-03-05
Name	David J. Bailey	Registration Number	59632

This collection of information is required by 37 CFR 1.114. 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 1.11 and 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.

If you need assistance in completing the form, call 1-800-PTO-9199 and select option 2.

PTO/AIA/22 (10-20)
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U.S. Patent and Trademark Office: U.S. DEPARTMENT OF COMMERCE
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PETITION FOR EXTENSION OF TIME UNDER 37 CFR 1.136(a)	Docket Number (Optional) C44-00999.CON15
Application Number 16/728,384	Filed December 27, 2019
For Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	
Art Unit 2633	Examiner Tesfaldet Bocure

This is a request under the provisions of 37 CFR 1.136(a) to extend the period for filing a reply in the above-identified application.

The requested extension and fee are as follows (check time period desired and enter the appropriate fee below):

		Fee	Small Entity Fee	Micro Entity Fee	
☐	One month (37 CFR 1.17(a)(1))	$220	$110	$55	$
☐	Two months (37 CFR 1.17(a)(2))	$640	$320	$160	$
☑	Three months (37 CFR 1.17(a)(3))	$1,480	$740	$370	$ 1480
☐	Four months (37 CFR 1.17(a)(4))	$2,320	$1,160	$580	$
☐	Five months (37 CFR 1.17(a)(5))	$3,160	$1,580	$790	$

☐ Applicant asserts small entity status. See 37 CFR 1.27.

☐ Applicant certifies micro entity status. See 37 CFR 1.29.
Form PTO/SB/15A or B or equivalent must either be enclosed or have been submitted previously.

☐ A check in the amount of the fee is enclosed.

☐ Payment by credit card. Form PTO-2038 is attached.

☐ The Director has already been authorized to charge fees in this application to a Deposit Account

☑ The Director is hereby authorized to charge any fees which may be required, or credit any overpayment, to Deposit Account Number 50-4407

☑ Payment made via EFS-Web.

WARNING: Information on this form may become public. Credit card information should not be included on this form. Provide credit card information and authorization on PTO-2038.

I am the

☐ applicant.

☐ attorney or agent of record. Registration number _____.

☑ attorney or agent acting under 37 CFR 1.34. Registration number 59,632.

/David J. Bailey/ — Signature

March 5, 2021 — Date

David J. Bailey — Typed or printed name

949.852.0000 — Telephone Number

NOTE: This form must be signed in accordance with 37 CFR 1.33. See 37 CFR 1.4 for signature requirements and certifications. Submit multiple forms if more than one signature is required, see below*.

☑ * Total of 1 forms are submitted.

This collection of information is required by 37 CFR 1.136(a). 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 1.11 and 1.14. This collection is estimated to take 6 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: Mail Stop PCT, Commissioner for Patents, P.O. Box 1450, Alexandria, VA 22313-1450.

Doc Code: TRACK1.REQ
Document Description: TrackOne Request

PTO/AIA/424 (04-14)

CERTIFICATION AND REQUEST FOR PRIORITIZED EXAMINATION UNDER 37 CFR 1.102(e) (Page 1 of 1)

First Named Inventor:	Maged F. Barsoum	Nonprovisional Application Number (if known):	16/728,384
Title of Invention:	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations		

APPLICANT HEREBY CERTIFIES THE FOLLOWING AND REQUESTS PRIORITIZED EXAMINATION FOR THE ABOVE-IDENTIFIED APPLICATION.

1. The processing fee set forth in 37 CFR 1.17(i)(1) and the prioritized examination fee set forth in 37 CFR 1.17(c) have been filed with the request. The publication fee requirement is met because that fee, set forth in 37 CFR 1.18(d), is currently $0. The basic filing fee, search fee, and examination fee are filed with the request or have been already been paid. I understand that any required excess claims fees or application size fee must be paid for the application.

2. I understand that the application may not contain, or be amended to contain, more than four independent claims, more than thirty total claims, or any multiple dependent claims, and that any request for an extension of time will cause an outstanding Track I request to be dismissed.

3. The applicable box is checked below:

I. ☐ **Original Application (Track One) - Prioritized Examination under § 1.102(e)(1)**

i. (a) The application is an original nonprovisional utility application filed under 35 U.S.C. 111(a). This certification and request is being filed with the utility application via EFS-Web.

---OR---

(b) The application is an original nonprovisional plant application filed under 35 U.S.C. 111(a). This certification and request is being filed with the plant application in paper.

ii. An executed inventor's oath or declaration under 37 CFR 1.63 or 37 CFR 1.64 for each inventor, or the application data sheet meeting the conditions specified in 37 CFR 1.53(f)(3)(i) is filed with the application.

II. ☑ **Request for Continued Examination - Prioritized Examination under § 1.102(e)(2)**

i. A request for continued examination has been filed with, or prior to, this form.
ii. If the application is a utility application, this certification and request is being filed via EFS-Web.
iii. The application is an original nonprovisional utility application filed under 35 U.S.C. 111(a), or is a national stage entry under 35 U.S.C. 371.
iv. This certification and request is being filed prior to the mailing of a first Office action responsive to the request for continued examination.
v. No prior request for continued examination has been granted prioritized examination status under 37 CFR 1.102(e)(2).

Signature /David J. Bailey/	Date March 5, 2021
Name (Print/Typed) David J. Bailey	Practitioner Registration Number 59,632

Note: *This form must be signed in accordance with 37 CFR 1.33. See 37 CFR 1.4(d) for signature requirements and certifications. Submit multiple forms if more than one signature is required.**

☑ *Total of 1 forms are submitted.

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 March 5, 2021 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 RCE
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806
March 5, 2021

Commissioner:

In response to the Office action mailed September 14, 2020, please amend the above-identified application as follows:

Amendments to the Claims are reflected in the listing of claims which begins on page 2 of this paper.

Remarks/Arguments begin on page 19 of this paper.

Amendments to the Claims:

This listing of claims will replace all prior versions, and listings, of claims in the application:

Listing of Claims:

1. (Currently Amended) 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 constellation points at a plurality of unique point locations, where:

the plurality of unique point locations are unequally spaced;

the constellation points each have a location and a different label;

and

the locations of at least two of the constellation points are the same ~~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. (Currently Amended) The communication system of claim 1, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations;

the plurality of ~~non-uniform~~ unequally spaced symbol constellations includes a plurality of ~~non-uniform~~ unequally spaced symbol 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~~ unequally spaced symbol

constellations of the first type include at least two constellation points having ~~identical~~ locations that are the same and different labels;

the receiver is capable of selecting an LDPC code rate and the ~~non-uniform~~ unequally spaced symbol constellation as a pair from a plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is only included in one of the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs.

3. (Currently Amended) The communication system of claim 2, wherein the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs includes:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 1/3 and a sixty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 5/8 and a two-hundred-and-fifty-six-point ~~non-uniform~~ unequally spaced symbol constellation of the first type; and

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type.

4. (Currently Amended) 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~~ unequally spaced symbol constellation pair that includes an LDPC code rate that is below 1/3 and a sixty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point ~~non-uniform~~ unequally spaced symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellations of the first type.

5. (Currently Amended) The communication system of claim 4, wherein:

the plurality of ~~non-uniform~~ unequally spaced symbol constellations include a plurality of ~~non-uniform~~ unequally spaced symbol 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 an ~~non-uniform~~ unequally spaced symbol constellation of the second type ~~share an identical~~ share locations that are the same;

the demapper is also capable of determining likelihoods using the demodulated signal and an ~~non-uniform~~ unequally spaced symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs include:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type; and

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type.

6. (Currently Amended) The communication system of claim 1, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellations when operated at the same SNR.

7. (Currently Amended) 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~~ unequally spaced symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

8. (Currently Amended) The communication system of claim 1, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced 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. (Cancelled)

10. (Currently Amended) 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. (Currently Amended) 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 constellation points at a plurality of unique point locations, where:

the plurality of unique point locations are unequally spaced;

the constellation points each have a location and a different label; and

the locations of at least two of the constellation points are the same~~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. (Currently Amended) The communication system of claim 11, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations;

the plurality of ~~non-uniform~~ unequally spaced symbol constellations includes a plurality of ~~non-uniform~~ unequally spaced symbol 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~~ unequally spaced symbol constellations of the first type include at least two constellation points having ~~identical~~ locations that are the same and different labels;

the receiver selects an LDPC code rate and the ~~non-uniform~~ unequally spaced symbol constellation as a pair from a plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is only included in one of the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs.

13. (Currently Amended) The communication system of claim 12, wherein the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs includes:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 1/3 and a sixty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 5/8 and a two-hundred-and-fifty-six-point ~~non-uniform~~ unequally spaced symbol constellation of the first type; and

multiple LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type.

14. (Currently Amended) 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~~ unequally spaced symbol constellation pair that includes an LDPC code

rate that is below 1/3 and a sixty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point ~~non-uniform~~ unequally spaced symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellations of the first type.

15. (Currently Amended) The communication system of claim 14, wherein:

the plurality of ~~non-uniform~~ unequally spaced symbol constellations includes a plurality of ~~non-uniform~~ unequally spaced symbol 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 an ~~non-uniform~~ unequally spaced symbol constellation of the second type ~~share an identical~~ have locations that are the same;

the demapper also determines likelihoods using the demodulated signal and an ~~non-uniform~~ unequally spaced symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs includes:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type; and

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type.

16. (Currently Amended) The communication system of claim 11, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations provides greater parallel decoding capacity when operated at a symbol constellation operating SNR than the other plurality of ~~non-uniform~~ unequally spaced symbol constellations when operated at the same SNR.

17. (Currently Amended) The communication system of claim 16, wherein the symbol constellation operating SNR is a channel SNR where the receiver uses the respective ~~non-uniform~~ unequally spaced symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

18. (Currently Amended) The communication system of claim 11, wherein

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced 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. (Cancelled)

20. (Currently Amended) 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. (Currently Amended) The communication system of claim [[11]]22, 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. (Currently Amended) 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 to transform the received signals into received bits, and the symbol constellation [[that]] includes constellation points at a plurality of unique point locations, where:

the plurality of unique point locations are unequally spaced;

the constellation points each have a location and a different label; and

the locations of at least two of the constellation points are the same ~~having identical locations and different labels~~.

23. (Currently Amended) The communication system of claim 22, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations;

the plurality of ~~non-uniform~~ unequally spaced symbol constellations includes a plurality of ~~non-uniform~~ unequally spaced symbol 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~~ unequally spaced 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~~ unequally spaced symbol constellation as a pair from a plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is only included in one of the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs.

24. (Currently Amended) The communication system of claim 23, wherein the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs includes:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced symbol constellation pairs that each include an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type.

25. (Currently Amended) 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~~ unequally spaced symbol constellation pair that includes an LDPC code rate that is below 1/3 and a sixty-four-point ~~non-uniform~~ unequally spaced symbol constellation of the first type;

equal to or less than 15.42dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 5/8 and a two-hundred-fifty-six-point ~~non-uniform~~ unequally spaced symbol constellation of the first type; and

equal to or less than 21.52dB while using an LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pair that includes an LDPC code rate below 7/10 and a one-thousand-twenty-four-point ~~non-uniform~~ unequally spaced symbol constellations of the first type.

26. (Currently Amended) The communication system of claim 25, wherein:

the plurality of ~~non-uniform~~ unequally spaced symbol constellations includes a plurality of ~~non-uniform~~ unequally spaced symbol 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 an ~~non-uniform~~ unequally spaced symbol constellation of the second type ~~share an identical~~ have locations that are the same;

the receiver also transforms received signals into received bits using an ~~non-uniform~~ unequally spaced symbol constellation of the second type; and

the plurality of predetermined LDPC code rate and ~~non-uniform~~ unequally spaced symbol constellation pairs include:

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type;

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type; and

multiple LDPC code rate and ~~non-uniform~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced 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~~ unequally spaced symbol constellation of the second type.

27. (Currently Amended) The communication system of claim 22, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations is characterized by the assignment of labels and spacing of the constellation points such that each of the plurality of ~~non-uniform~~ unequally spaced symbol constellations provides greater parallel decoding capacity when operated at a symbol constellation operating SNR than the other plurality of ~~non-uniform~~ unequally spaced symbol constellations when operated at the same SNR.

28. (Currently Amended) The communication system of claim 27, wherein the symbol constellation operating SNR is a channel SNR where the receiver uses the respective ~~non-uniform~~ unequally spaced symbol constellation to receive data at a frame error rate (FER) of 10^{-2} or lower.

29. (Currently Amended) The communication system of claim 22, wherein:

the ~~non-uniform~~ symbol constellation is selected from a plurality of ~~non-uniform~~ unequally spaced symbol constellations; and

each of the plurality of ~~non-uniform~~ unequally spaced 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. (Cancelled)

31. (New) The communication system of claim 1, wherein:

the symbol constellation is selected from a plurality of symbol constellations each having a same number of constellation points;

a first of the plurality of symbol constellations includes constellation points at a first set of unique point locations that are unequally spaced;

a second of the plurality of symbol constellations includes constellation points at a second set of unique point locations that are unequally spaced; and

the number of unique point locations in the first set of unique point locations is different from the number of unique point locations in the second set of unique point locations.

32. (New) The communication system of claim 11, wherein:

the symbol constellation is selected from a plurality of symbol constellations each having a same number of constellation points;

a first of the plurality of symbol constellations includes constellation points at a first set of unique point locations that are unequally spaced;

a second of the plurality of symbol constellations includes constellation points at a second set of unique point locations that are unequally spaced; and

the number of unique point locations in the first set of unique point locations is different from the number of unique point locations in the second set of unique point locations.

33. (New) The communication system of claim 22, wherein:

the symbol constellation is selected from a plurality of symbol constellations each having a same number of constellation points;

a first of the plurality of symbol constellations includes constellation points at a first set of unique point locations that are unequally spaced;

a second of the plurality of symbol constellations includes constellation points at a second set of unique point locations that are unequally spaced; and

the number of unique point locations in the first set of unique point locations is different from the number of unique point locations in the second set of unique point locations.

REMARKS/ARGUMENTS

INTERVIEW SUMMARY

Applicant wishes to thank Examiner Bocure for extending the courtesy of an Examiner Interview on January 12, 2021 with respect to U.S. Patent Application Serial Nos. 16/728,384 and 16/728,397. The interview was attended by the inventor Dr. Christopher Jones, David Loo (Chief Operating Officer and General Counsel of Applicant), and David Bailey (Counsel for Applicant). During the interview, Dr. Jones discussed his and his co-inventor's work developing constellations including unequally spaced symbol constellations that include overlapping constellation points. Dr. Jones discussed the difference between the hierarchical modulations described in U.S. Patent Publication No. 2008/0200114 to Eberlein et al. (Eberlein) and the claimed unequally spaced symbol constellations. Dr. Jones noted that the hierarchical modulation schemes described in Eberlein are incompatible with the use of overlapping constellations points having locations that are the same. Specifically, the goal of a hierarchical modulation scheme is to enable multiple layers that can each be decoded above a specific SNR. Therefore, the points in each additional layer must occupy unique point locations that are not shared with other constellation points. Dr. Jones provided a discussion of the benefits that unequally spaced symbol constellations having constellation point locations that are the same can provide at appropriate SNR operating points. Dr. Jones also attempted to clarify the manner in which constellation points can overlap. Specifically, Dr. Jones explained that the claimed constellations have a number of constellation points that is greater than the number of unique point locations occupied by the constellations points of the symbol constellations. Therefore, some of the constellation points are overlapping meaning that they have different labels, but their locations are the same. Dr. Jones noted that an important result of he and his co-inventor's work is that unequally spaced symbol constellations can practically achieve the same performance improvements as can be theoretically achieved by probabilistically shaped constellations and that, in many instances, unequally spaced symbol constellations utilize constellation points having locations that are the same and different labels to achieve these performance

improvements. Examiner Bocure acknowledged that many of the previously pending claims were already clearly differentiated compared to the prior art of record including the hierarchical modulation schemes of Eberlein. Examiner Bocure further acknowledged Applicant's remarks regarding the patentability of previously pending claims 1, 11, and 22 and suggested submitting amendments to at least claims 1, 11, and 22 further clarifying the manner in which a symbol constellation can have points at a plurality of unique point locations and that some of the constellation points withing the constellation can have locations that are the same (i.e. they are located at the same unique point location) and different labels. No agreement was reached with respect to the allowability of claims 1, 11, and 22.

In order to advance prosecution, as suggested Applicant submits herewith clarifying amendments to claims 1, 11, and 22.

SUMMARY OF THE OFFICE ACTION

The Office action dated September 14, 2020 (the "Office action") contains the following objections and rejections:

- claims 1, 10, 11, 20, 21 and 22 are rejected under pre-AIA 35 U.S.C. §102(e) as being anticipated by Eberlein;
- 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); and
- claims 2-8, 12-18, and 23-29 were objected to as being dependent upon a rejected base claim, but would be allowable if rewritten in independent form.

SUMMARY OF THE AMENDMENTS TO THE CLAIMS

Claims 1-8, 10-18, 20-29, and 31-33 are currently pending. Claims 1-8, 10-18, and 20-29 are currently amended. Claims 31-33 are newly added and claims 9, 19, and 30 were cancelled to retain Track 1 eligibility of the above referenced application. Claims 1, 11, and 22 are amended for clarity as discussed above. Claims 2-8, 10, 12-18, 20, 21, and 23-29 are amended to conform to the amendments to claims 1, 11, and 22. Support

for the amendments to claims 1, 11, and 22 and newly added claims 31-33 is discussed further below.

Claim 1 has been amended to recite "a symbol constellation that includes constellation points at a plurality of unique point locations, where: the plurality of unique point locations are unequally spaced; the constellation points each have a location and a different label; and the locations of at least two of the constellation points are the same".

Support for the symbol constellations recited in claim 1 can be found throughout the specification and figures of the above referenced application. The significance of using overlapping constellation points in unequally spaced constellations was first observed in U.S. Provisional Application Serial No. 60/933,319 ('319 provisional), which forms the basis of the earliest priority date claimed in the above referenced application. Specifically, the '319 provisional notes that non-uniform equiprobable constellations χ (i.e. geometrically shaped constellations) can achieve equivalent performance as probabilistically shaped constellations S using the same number or a larger number of constellation points (i.e. $|\mathcal{X}| \geq |S|$).

> *Proposition 1 (Shaping):* For any given alphabet $\mathcal{S}$ with rational $p(s) \forall s \in \mathcal{S}$ there exists an alphabet $\mathcal{X}$ with $p(x) = \frac{1}{|\mathcal{X}|}$ such that $I(X;Y) \geq I(S;Y)$.
>
> *Proof:* Allow $|\mathcal{X}| \geq |\mathcal{S}|$. Define $|\mathcal{S}|$ subgroups $\mathcal{G}_i \in \Re^n$ and let $g_i = s_i$ for $i = 1, \ldots, |\mathcal{S}|$. Map elements from $\mathcal{X}$ into $\mathcal{G}$ such that the ith subgroup has $\frac{|\mathcal{G}_i|}{|\mathcal{X}|} = p(s_i)$. This yields $I(X;Y) = I(G;Y) = I(S;Y)$. Releasing the constraint that elements in $\mathcal{X}$ precisely map to $|\mathcal{S}|$ possible elements in $\mathcal{G}$ permits the inequality. ■
>
> *Corollary 2:* The cardinality, $|\mathcal{X}|$, of the set of equiprobable points $\mathcal{X}$ is no less than the lowest common multiple of the denominators of the rational fractions that define $p(s_i)$ for $i = 1, \ldots, |\mathcal{S}|$.

The proof from the '319 provisional notes that "[r]eleasing the constraint that elements in χ precisely map to $|S|$ possible elements" enabled the inventors to show that

"any gain in capacity that can be achieved by probabilistic shaping can also be achieved by equiprobable but non-uniformly spaced signaling i.e. geometrical shaping." Stated another way, the '319 provisional shows that practical (i.e. "real world") communication systems can be constructed using unequally spaced symbol constellations having multiple overlapping constellation points, which can achieve comparable performance to uniform constellations in which the constellation points are not transmitted with equal probability (i.e. previous approaches "that attempt to approximate the capacity achieving Gaussian input distribution by transmitting some constellation points with greater frequency than others, an operation which poses several practical difficulties.").

The above referenced application expands upon the disclosure provided in the '319 provisional by providing various examples of symbol constellations that include constellation points at a number of unique point locations, where the plurality of unique point locations are unequally spaced, and the locations of at least two of the constellation points are the same. The benefits of using symbol constellations in which the number of constellation points is greater than the number of unique point locations of the symbol constellation is observed with respect to PAM-4 constellations in paragraph [0072] noting "[a]t low SNRs, the PD capacity optimized PAM-4 constellations have only 2 unique points". Furthermore, it is demonstrated that symbol constellations with the characteristics of the symbol constellation recited in claim 1 can achieve improved performance over uniform symbol constellations that maximize the minimum distance between the constellation points (in the degrees of freedom available to the constellation) at specific SNRs.

The specification of the above referenced application provides numerous examples of the symbol constellation recited in claim 1, including the symbol constellations that can be observed in Fig. 12d (reproduced below with annotations).

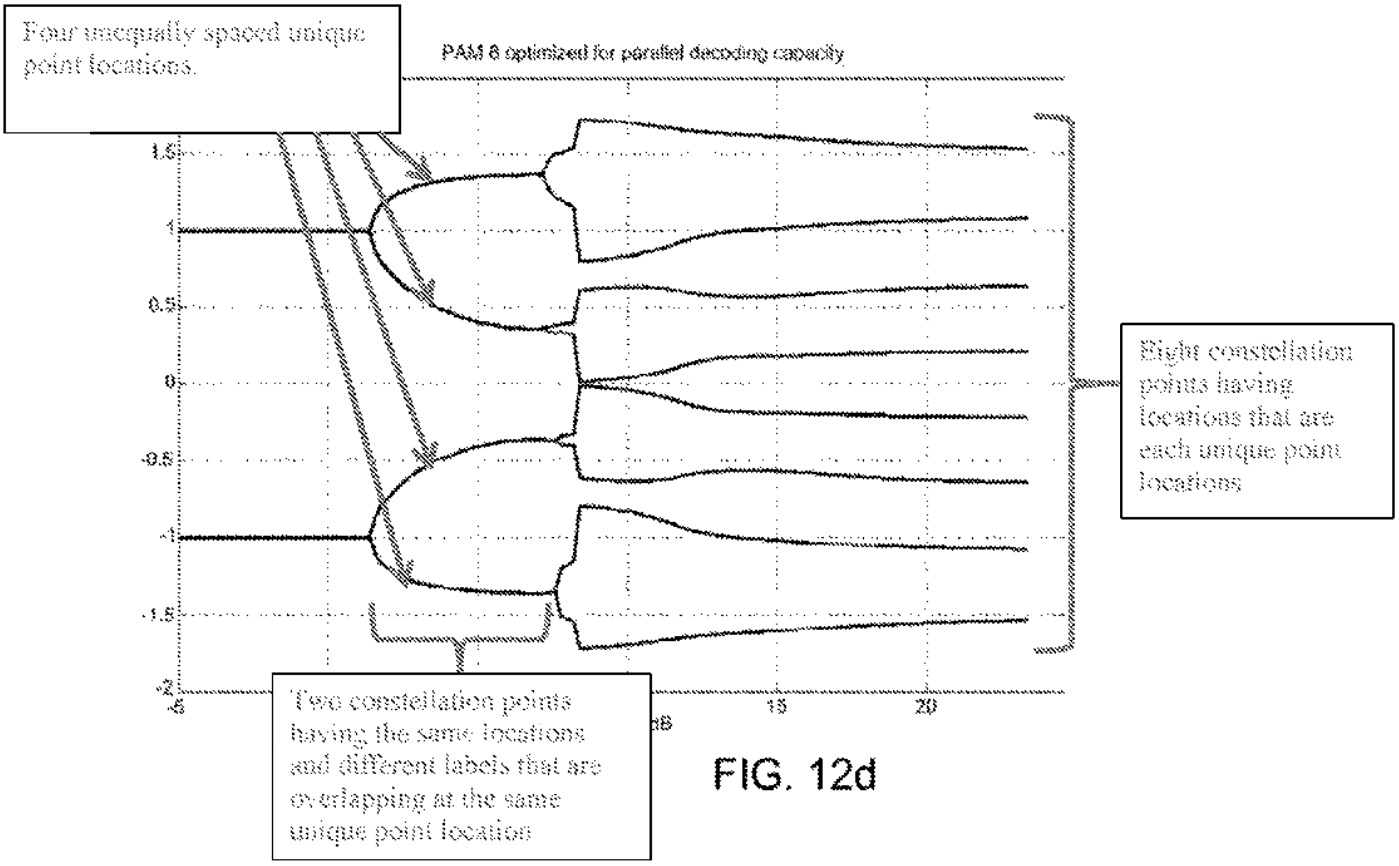


FIG. 12d

As SNR decreases, the locations of the eight non-overlapping constellation points of the PD capacity optimized PAM-8 constellations merge into four unique point locations that are unequally spaced (i.e. four sets of two constellation points that each occupy one of the four unique point locations that are unequally spaced).

Amplitudes of PAM-8 constellations for specific SNRs are provided in Fig. 13b, which is reproduced below including annotations indicating constellation points having locations that are the same 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

At an SNR of 5.27 dB, the PAM-8 constellation has four unique point locations and includes four sets of two constellation points having locations that are the same and different labels.

Similar examples of the symbol constellation recited in claim 1 can be observed in Figs. 14d, 15b, 16d, 17b, 22b and 23b. Symbol constellations that use phase shift keying in which the locations of at least two of the constellation points are the same can be observed in Figs. 19, and 20. The specification of the above referenced application also discloses that one-dimensional unequally spaced 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 the same 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, unequally spaced 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.").

In light of the above, Applicant respectfully submits that support for the amendments to claim 1 can be found both in the specification of the '319 provisional and the specification and figures of the present application. Accordingly, Applicant respectfully submits that no new matter is added by way of the amendments to claim 1. Furthermore, Applicant respectfully submits that the amendments to claims 11 and 22 are also supported by the specification of the '319 provisional and the specification and figures of the present application.

With specific regard to the new claims, claims 31, 32, and 33 recite that the symbol constellation is selected from a plurality of symbol constellations each having a same number of constellation points, where at least two of the symbol constellations have the different numbers of unique point locations. Support for claims 32, 34, and 36 can be found throughout the specification including paragraphs [0043], [0044], [0078], [0079], and [0082], and Figs. 12d and 13b.

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.

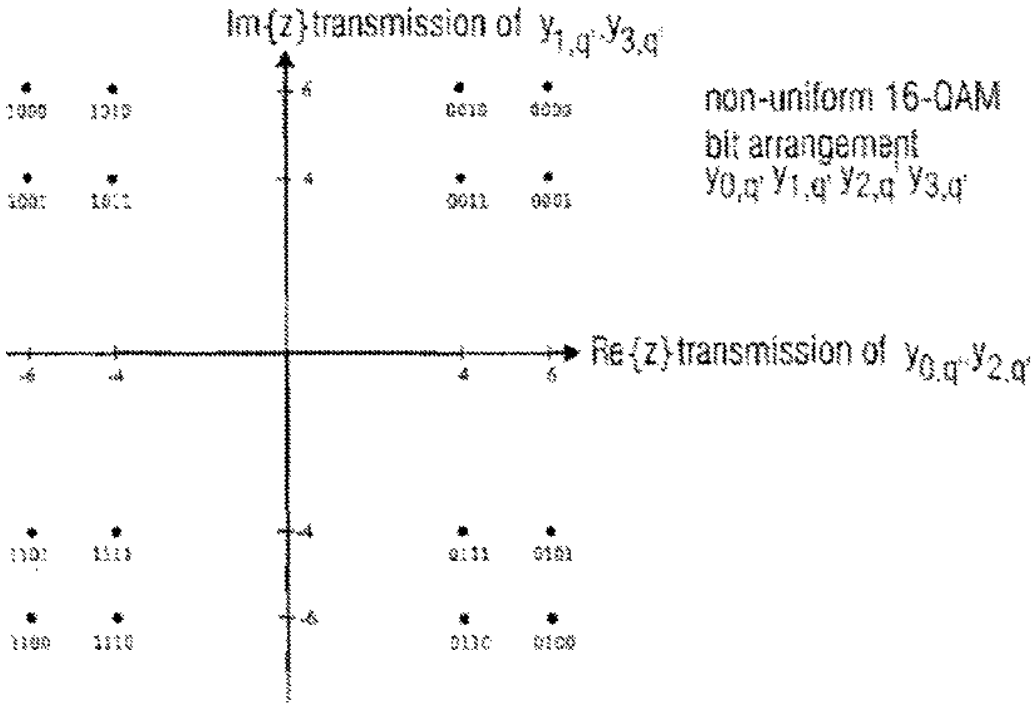


FIGURE 3A

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, p. 4-6. 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 4. Applicant respectfully submits that Fig. 3A of Eberlein above clearly shows the four points in the first quadrant have different unique point locations that are not the same. While the constellation points do have different labeling, they do not occupy unique point locations that are the same. As discussed by Dr. Jones, the fact that the constellation points in the first quadrant

occupy different unique point locations enables the communication system to transmit two separate layers of data using the hierarchical modulation scheme.

Accordingly, Applicant both maintains that Eberlein fails to anticipate previously pending claims 1, 11, and 22 and further submits that Eberlein fails to disclose currently amended claims 1, 11, and 22. In light of the clarifying amendments to claims 1, 11, and 22, Applicant respectfully requests withdrawal of the rejection of claims 1, 11, and 22. Furthermore, Applicant respectfully submits that claims 10, 20, and 21 are allowable for reasons including the reasons outlined above with respect to amended claims 1, and 11. Accordingly, Applicant also respectfully requests withdrawal of the rejections of claims 10, 20, and 21.

OBJECTIONS TO CLAIMS 2-8, 12-18, AND 23 - 28

Claims 2-8, 12-18, and 23-28 are objected to as being dependent upon a rejected base claim, but are indicated as allowable if rewritten in independent form including all of the limitations of the base claim and any intervening claims. As such Applicant respectfully submits that claims 2-8, 12-18, and 23-28, remain allowable.

PATENTABILITY OF CLAIM 29

The Office action does not contain any rejection or objections to currently pending claim 29. Applicant suspect that claim 29 is implicitly objected to as being dependent upon a rejected base claim, but allowable if rewritten in independent form including all of the limitations of the base claim and any intervening claims. Accordingly, Applicant respectfully submits that claim 29 remains allowable.

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) based upon Eberlein in view of Walker. Applicant respectfully submits that amended claims 9, 19, and 30 are allowable over the combined teachings of Eberlein and Walker. In order to retain Track One status, however, Applicant has cancelled claims 9, 19, and 30.

NEW CLAIMS 31-33

Applicant respectfully submits that new claims 31-33 are allowable over the prior art of record taken alone and in combination for reasons including those recited above with respect to amended claims 1, 11, and 22

DISCUSSION OF ADDITIONAL PRIOR ART REFERENCES

Applicant respectfully submits that the currently pending claims are patentable with respect to the prior art of record taken alone and in combination. In addition to discussing Eberlein and Walker, the above remarks provide a general basis for understanding the distinction between hierarchical modulation schemes and the various symbol constellations incorporating constellation points having locations that are the same recited in the currently pending claims. The following remarks discuss the patentability of the currently pending claims with respect to a number of other types of symbol constellations. Accordingly, Applicant specifically directs the Examiner's attention to the references and classes of constellation identified in the remarks below and submits that the claimed symbol constellations are patentable in view of the referenced prior art constellations.

Sommer and Fettweis

The '319 provisional notes the teachings of Dirk Sommer and Gerhard Fettweis, "Signal Shaping by Non-Uniform QAM for AWGN Channels and Applications Using Turbo Coding", ITG Conference on Source and Channel Coding, Jan. 2000, pp. 81-86 (hereafter "Sommer and Fettweis", previously submitted via Information Disclosure Statement). Sommer and Fettweis disclose non-uniform constellations designed to achieve "a Gaussian distribution of the output signal". *Id.* at p. 81. Sommer and Fettweis describe a process for constructing a non-uniform constellation "using N-ASK, $N=2^m$, we split the Gaussian cumulative distribution function (CDF) into N sections of *equal probability*". *Id.* Accordingly, it is a requirement of the non-uniform constellation of Sommer and Fettweis that the constellation points be assigned to different point locations. Furthermore, the constellations of Sommer and Fettweis are independent of the signal-to-noise ratio within the channel (see, for example, claim 8).

Applicant respectfully submits that attempting to assign multiple constellation points to the same point location would be contrary to the principle of operation of the non-uniform constellations described by Sommer and Fettweis, which rely upon dividing a Gaussian cumulative distribution function into a number of equiprobable sections corresponding to the number of constellation points of the non-uniform constellation. Accordingly, Applicant submits that it would not be obvious to modify the non-uniform constellations described by Sommer and Fettweis with a goal of developing "a symbol constellation that includes constellation points at a plurality of unique point locations, where: the plurality of unique point locations are unequally spaced; the constellation points each have a location and a different label; and the locations of at least two of the constellation points are the same" as recited in claim 1.

Applicant further questions the extent to which a POSITA would rely upon the teachings of Sommer and Fettweis in attempting to develop new unequally spaced symbol constellations given Sommer and Fettweis' conclusions regarding the usefulness of their results. Specifically, Sommer and Fettweis note that "[c]ompared to other shaping methods, the achieved gain is lower... shaping gain is exhibited only when the constellations are used with code rates R = 0.5... Furthermore, the enlarged signal point constellation, entailing a considerably higher peak-to-average power ratio, can be prohibitive in some applications." As Sommer and Fettweis note, "[t]he main Advantage is the simplicity of the method." Accordingly, Applicant respectfully submits that a POSITA would most likely have sought to utilize one of the constellations described by Sommer and Fettweis (which do not include constellation points where the locations of at least two of the constellation points are the same) for its simplicity or would have attempted to utilize a different (more complex) shaping method that Sommer and Fettweis report would be likely to achieve better performance.

De Gaudenzi et al. and DVB-S2

Riccardo De Gaudenzi at the European Space Agency and a number of co-authors have published extensively on the design of Amplitude Phase Shift Keyed (APSK) constellations including a number of concentric rings of uniformly spaced constellation

points using a technique involving selecting the ratios of the concentric rings to maximize mutual information. The technique described by De Gaudenzi et al. is specific to APSK constellations and yields what is reported as being a "small but significant improvement" over APSK constellations designed to maximize minimum distance. De Gaudenzi, Riccardo et al., "Turbo-Coded APSK Modulations Design for Satellite Broadband Communications", International Journal of Satellite Communications and Networking, vol. 24, No. 4, Jul. 1, 2006, pp. 261-281, 279. Applicants believe that the constellations proposed by Dr. De Gaudenzi and his co-authors are the basis of the APSK constellations adopted in the DVB-S2 standard. A number of publications co-authored by Dr. De Gaudenzi and the DVB-S2 standard were previously submitted via Information Disclosure Statements. An example of an APSK optimized by De Gaudenzi et al. in the above referenced paper is reproduced below.

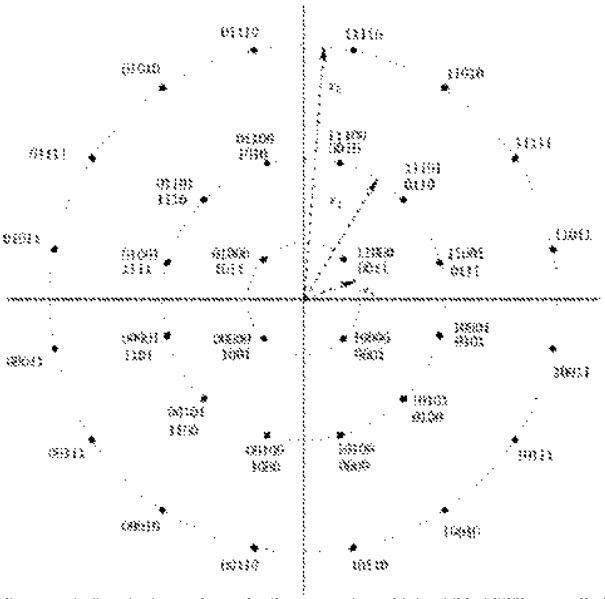


Figure 1. Parametric description and pseudo-Gray mapping of 16 and 32-APSK constellations with $\alpha_1 = 3$, $\alpha_2 = 15$, $\phi_2 = 0$ and $\alpha_1 = 4$, $\alpha_2 = 12$, $\alpha_3 = 16$, $\phi_2 = 0$, $\phi_3 = \pi/16$ respectively. For the first two rings: mapping below corresponds to 4+12 APSK, mapping above to 4+12+16-APSK.

The method proposed by De Gaudenzi et al. does not provide a mechanism for modifying the relative spacing of constellation points on the concentric rings. Quite the contrary, De Gaudenzi et al. teach away from using anything other than rings of uniformly spaced constellation points noting that the relative phase of the constellation points has no significant impact on the performance of the constellation. *Id.* at p. 268 ("This absence of dependency is justified by the fact that the optimum constellation separates the rings by a distance larger than the number of points in the ring itself, so that the relative phase ϕ has no significant impact...").

Accordingly, Applicant respectfully submits that the techniques disclosed in the prior art for designing APSK constellations provide no mechanism for designing unequally spaced symbol constellations including at least two constellation points having different labels and locations that are the same. Furthermore, a POSITA would have been taught away from attempting to construct an unequally spaced APSK constellation including at least two constellation points having different labels and locations that are the same, because De Gaudenzi et al. specifically teach that modifying relative phase of constellation points within an APSK optimized for mutual information does not improve performance. *Id.*

Probabilistic Shaping using Non-Unique Mappings

In 2005, researchers at Munich University of Technology published a paper reporting a shaping gain achieved by assigning a "non-uniform amount of distinct bit labels to the signal points [of an equally spaced symbol constellation]". Schreckenbach, F., and P. Henkel. "Signal shaping using non-unique symbol mappings." *Proc. 43rd Annual Allerton Conf. on Commun., Cont., and Comp., Monticello, IL.* 2005, p. 1 (hereafter "Schreckenbach and Henkel"). The constellations utilized by Schreckenbach and Henkel, when assigning multiple labels to point locations, were equally spaced constellations. Shaping gains were obtained by modifying the number of unique labels assigned to each of the equally spaced constellation points. Accordingly, the approach of Schreckenbach and Henkel was designed to obtain the benefits of probabilistic shaping through assignment of multiple equiprobable labelings to individual constellation points.

Schreckenbach and Henkel do not provide any teachings with respect to assignment of multiple equiprobable labelings in symbol constellations, where the constellation points occupy unique point locations that are unequally spaced.

Applicant respectfully submits that a POSITA would be unable to adapt the method of Schreckenbach and Henkel to design unequally spaced symbol constellations including at least two constellation points locations that are the same and different labels. The method of Schreckenbach and Henkel applies "a greedy label filling method to optimize the probability distribution of the signal point and a binary switching algorithm to optimize the bit mapping." *Id.* at p. 9. The probability distribution that is optimized by Schreckenbach and Henkel's method assumes equal spacing of constellation points. In effect, Schreckenbach and Henkel are attempting to assign labels to individual constellation points within an equally spaced symbol constellation to approximate the probability distribution required to achieve optimal probabilistic shaping. *Id.* at p. 4 ("To achieve a probability distribution close to the one given by the value of λ, we assign a non-uniform amount of distinct labels to the signal points and optimize the number of labels per signal point."). Schreckenbach and Henkel do not contemplate using the disclosed method with unequally spaced symbol constellations and do not provide any information that would enable a POSITA to determine the appropriate probability distribution to utilize as the basis for attempting to assign labels to point locations that are unequally spaced.

CONCLUSION

In view of the foregoing amendments and remarks, Applicant respectfully requests the withdraw of the rejections of the currently pending claims 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 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:	David J. Bailey/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:				
REQUEST FOR PRIORITIZED EXAMINATION	1817	1	4200	4200
Pages:				
Claims:				
Miscellaneous-Filing:				
PUBL. FEE- EARLY, VOLUNTARY, OR NORMAL	1504	1	0	0
PROCESSING FEE, EXCEPT PROV. APPLS.	1830	1	140	140
Petition:				
Patent-Appeals-and-Interference:				

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Post-Allowance-and-Post-Issuance:				
Extension-of-Time:				
Extension - 3 months with $0 paid	1253	1	1480	1480
Miscellaneous:				
RCE- 1ST REQUEST	1801	1	1360	1360
		Total in USD ($)		**7180**

Electronic Acknowledgement Receipt	
EFS ID:	42107302
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
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Payment was successfully received in RAM	$7180
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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	Request for Continued Examination (RCE)	00999con15rce.pdf	85411 691bc35c18d92054ebcd7b59bcbf800171f48e12	no	2
Warnings:					
This is not a USPTO supplied RCE SB30 form.					
Information:					
2	Extension of Time	00999con15extoftime.pdf	129329 1823bc0b99d071f88c71570c4f883b31935afd3f	no	1
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Information:					
4		00999con15amendment.pdf	558476 df02da9e3b6298efa7f18cbdadd8ee81eb0200cb	yes	33

Multipart Description/PDF files in .zip description

Document Description	Start	End
Amendment Submitted/Entered with Filing of CPA/RCE	1	1
Claims	2	18
Applicant Arguments/Remarks Made in an Amendment	19	33

Warnings:

Information:

5	Fee Worksheet (SB06)	fee-info.pdf	39573 0f317af9c064321107dbced549cd2adbf7d88ab0	no	2
Warnings:					
Information:					
			Total Files Size (in bytes):	939034	

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/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

FOR	(Column 1) NUMBER FILED	(Column 2) 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))					
* If the difference in column 1 is less than zero, enter "0" in column 2.				TOTAL	

APPLICATION AS AMENDED - PART II

AMENDMENT	03/05/2021	(Column 1) CLAIMS REMAINING AFTER AMENDMENT		(Column 2) HIGHEST NUMBER PREVIOUSLY PAID FOR	(Column 3) PRESENT EXTRA		RATE ($)	ADDITIONAL FEE ($)
	Total (37 CFR 1.16(i))	* 29	Minus	** 30	= 0		x $100 =	0
	Independent (37 CFR 1.16(h))	* 3	Minus	*** 3	= 0		x $480 =	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) CLAIMS REMAINING AFTER AMENDMENT		(Column 2) HIGHEST NUMBER PREVIOUSLY PAID FOR	(Column 3) 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.	LIE
** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 20, enter "20".	/TARA J WITCHER/
*** 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.

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
		10848989	B2	2020-11-24	Barsoum et al.	
		10848990	B2	2020-11-24	Barsoum et al.	
		10863370	B2	2020-12-08	Barsoum et al.	
		10887780	B2	2021-01-05	Barsoum et al.	

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]
		CLEVORN ET AL., "Iterative Decoding of BICM with Non-Regular Signal Constellation Sets", January 2004, Retrieved from: https://www.researchgate.net/publication/228849016_Iterative_decoding_of_BICM_with_non-regular_signal_constellation_sets, 8 pgs.	
		DE GAUDENZI ET AL., "Adaptive coding and modulation for satellite broadband networks: From theory to practice", International Journal of Satellite Communications, Vol. 28, pp. 59-111, 2010, published online March 27, 2009, 53 pgs.	
		KAYHAN ET AL., "Constellation Design for Transmission over Nonlinear Satellite Channels", arXiv, October 5, 2012, arXiv: 1210.1762v1, 7 pgs.	
		LIOLIS ET AL., "On 64-APSK Constellation Design Optimization", 2008 10th International Workshop on Signal Processing for Space Communications, Conference Date: October 6-8, 2008, Rhodes Island, Greece, 7 pgs.	
		SCHRECKENBACH, FRANK, "Iterative Decoding of Bit-Interleaved Coded Modulation", Dissertation, Technische Universität München, 2007, 169 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	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 Acknowledgement Receipt	
EFS ID:	42121608
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:	08-MAR-2021
Filing Date:	27-DEC-2019
Time Stamp:	20:28:51
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	clevornetal.pdf	638295 981afb01bc99fa3ea555f1d1a420238ec6c1b986	no	8
Warnings:					

Information:					
2	Non Patent Literature	degaudenzi53pgs.pdf	1292960 bb730c667c1ed5e0292ca88d47ee3bd7b62847e2	no	54
Warnings:					
Information:					
3	Non Patent Literature	kayhanetal.pdf	520529 538e527c59ff5f08dd2306fa903cd1852ca7b575	no	7
Warnings:					
Information:					
4	Non Patent Literature	liolisetal.pdf	375197 10c50c8634b86b6e7a963cecc8fcc07139febd31	no	7
Warnings:					
Information:					
5	Non Patent Literature	schreckenbach169pgs.pdf	1285966 e7e732261006bbac922e67d688805a974ceb2f4a	no	169
Warnings:					
Information:					
6	Transmittal Letter	00999con15idsxtml.pdf	120987 b9c7d6d68da4aeefdab7de19710ca4c8b4d62637	no	2
Warnings:					
Information:					
7	Information Disclosure Statement (IDS) Form (SB08)	00999con15ids1449.pdf	143424 64b7ac3b1b38d9f8b880d8e74a8b0d0c8f954344	no	2
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					
			Total Files Size (in bytes):	4377358	

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.

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 March 8, 2021 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(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
March 8, 2021

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

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 03/09/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER
BOCURE, TESFALDET

ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
03/09/2021	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 BRIAN W BROWN	**Art Unit** OPET	**AIA (FITF) Status** No

1. THE REQUEST FILED 05 March 2021 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 BRIAN BROWN at (571)272-5338. In his/her absence, calls may be directed to Petition Help Desk at (571) 272-3282.

/BRIAN W BROWN/ Petitions Examiner, OPET	

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

NOTICE OF ALLOWANCE AND FEE(S) DUE

71897 7590 03/19/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER	
BOCURE, TESFALDET	
ART UNIT	PAPER NUMBER
2633	

DATE MAILED: 03/19/2021

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

TITLE OF INVENTION: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

APPLN. TYPE	ENTITY STATUS	ISSUE FEE DUE	PUBLICATION FEE DUE	PREV. PAID ISSUE FEE	TOTAL FEE(S) DUE	DATE DUE
nonprovisional	UNDISCOUNTED	$1200	$0.00	$0.00	$1200	06/21/2021

THE APPLICATION IDENTIFIED ABOVE HAS BEEN EXAMINED AND IS ALLOWED FOR ISSUANCE AS A PATENT. PROSECUTION ON THE MERITS IS CLOSED. THIS NOTICE OF ALLOWANCE IS NOT A GRANT OF PATENT RIGHTS. THIS APPLICATION IS SUBJECT TO WITHDRAWAL FROM ISSUE AT THE INITIATIVE OF THE OFFICE OR UPON PETITION BY THE APPLICANT. SEE 37 CFR 1.313 AND MPEP 1308.

THE ISSUE FEE AND PUBLICATION FEE (IF REQUIRED) MUST BE PAID WITHIN THREE MONTHS FROM THE MAILING DATE OF THIS NOTICE OR THIS APPLICATION SHALL BE REGARDED AS ABANDONED. THIS STATUTORY PERIOD CANNOT BE EXTENDED. SEE 35 U.S.C. 151. THE ISSUE FEE DUE INDICATED ABOVE DOES NOT REFLECT A CREDIT FOR ANY PREVIOUSLY PAID ISSUE FEE IN THIS APPLICATION. IF AN ISSUE FEE HAS PREVIOUSLY BEEN PAID IN THIS APPLICATION (AS SHOWN ABOVE), THE RETURN OF PART B OF THIS FORM WILL BE CONSIDERED A REQUEST TO REAPPLY THE PREVIOUSLY PAID ISSUE FEE TOWARD THE ISSUE FEE NOW DUE.

HOW TO REPLY TO THIS NOTICE:

I. Review the ENTITY STATUS shown above. If the ENTITY STATUS is shown as SMALL or MICRO, verify whether entitlement to that entity status still applies.

If the ENTITY STATUS is the same as shown above, pay the TOTAL FEE(S) DUE shown above.

If the ENTITY STATUS is changed from that shown above, on PART B - FEE(S) TRANSMITTAL, complete section number 5 titled "Change in Entity Status (from status indicated above)".

For purposes of this notice, small entity fees are 1/2 the amount of undiscounted fees, and micro entity fees are 1/2 the amount of small entity fees.

II. PART B - FEE(S) TRANSMITTAL, or its equivalent, must be completed and returned to the United States Patent and Trademark Office (USPTO) with your ISSUE FEE and PUBLICATION FEE (if required). If you are charging the fee(s) to your deposit account, section "4b" of Part B - Fee(s) Transmittal should be completed and an extra copy of the form should be submitted. If an equivalent of Part B is filed, a request to reapply a previously paid issue fee must be clearly made, and delays in processing may occur due to the difficulty in recognizing the paper as an equivalent of Part B.

III. All communications regarding this application must give the application number. Please direct all communications prior to issuance to Mail Stop ISSUE FEE unless advised to the contrary.

IMPORTANT REMINDER: Maintenance fees are due in utility patents issuing on applications filed on or after Dec. 12, 1980. It is patentee's responsibility to ensure timely payment of maintenance fees when due. More information is available at www.uspto.gov/PatentMaintenanceFees.

PART B - FEE(S) TRANSMITTAL

Complete and send this form, together with applicable fee(s), by mail or fax, or via EFS-Web.

By mail, send to: Mail Stop ISSUE FEE
Commissioner for Patents
P.O. Box 1450
Alexandria, Virginia 22313-1450

By fax, send to: (571)-273-2885

INSTRUCTIONS: This form should be used for transmitting the ISSUE FEE and PUBLICATION FEE (if required). Blocks 1 through 5 should be completed where appropriate. All further correspondence including the Patent, advance orders and notification of maintenance fees will be mailed to the current correspondence address as indicated unless corrected below or directed otherwise in Block 1, by (a) specifying a new correspondence address; and/or (b) indicating a separate "FEE ADDRESS" for maintenance fee notifications.

CURRENT CORRESPONDENCE ADDRESS (Note: Use Block 1 for any change of address)

71897 7590 03/19/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

Note: A certificate of mailing can only be used for domestic mailings of the Fee(s) Transmittal. This certificate cannot be used for any other accompanying papers. Each additional paper, such as an assignment or formal drawing, must have its own certificate of mailing or transmission.

Certificate of Mailing or Transmission

I hereby certify that this Fee(s) Transmittal is being deposited with the United States Postal Service with sufficient postage for first class mail in an envelope addressed to the Mail Stop ISSUE FEE address above, or being transmitted to the USPTO via EFS-Web or by facsimile to (571) 273-2885, on the date below.

(Typed or printed name)
(Signature)
(Date)

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

TITLE OF INVENTION: Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

APPLN. TYPE	ENTITY STATUS	ISSUE FEE DUE	PUBLICATION FEE DUE	PREV. PAID ISSUE FEE	TOTAL FEE(S) DUE	DATE DUE
nonprovisional	UNDISCOUNTED	$1200	$0.00	$0.00	$1200	06/21/2021

EXAMINER	ART UNIT	CLASS-SUBCLASS
BOCURE, TESFALDET	2633	375-262000

1. Change of correspondence address or indication of "Fee Address" (37 CFR 1.363).

❑ Change of correspondence address (or Change of Correspondence Address form PTO/SB/122) attached.

❑ "Fee Address" indication (or "Fee Address" Indication form PTO/SB/47; Rev 03-09 or more recent) attached. **Use of a Customer Number is required.**

2. For printing on the patent front page, list
(1) The names of up to 3 registered patent attorneys or agents OR, alternatively,
(2) The name of a single firm (having as a member a registered attorney or agent) and the names of up to 2 registered patent attorneys or agents. If no name is listed, no name will be printed.

1_____
2_____
3_____

3. ASSIGNEE NAME AND RESIDENCE DATA TO BE PRINTED ON THE PATENT (print or type)

PLEASE NOTE: Unless an assignee is identified below, no assignee data will appear on the patent. If an assignee is identified below, the document must have been previously recorded, or filed for recordation, as set forth in 37 CFR 3.11 and 37 CFR 3.81(a). Completion of this form is NOT a substitute for filing an assignment.

(A) NAME OF ASSIGNEE (B) RESIDENCE: (CITY and STATE OR COUNTRY)

Please check the appropriate assignee category or categories (will not be printed on the patent) : ❑ Individual ❑ Corporation or other private group entity ❑ Government

4a. Fees submitted: ❑Issue Fee ❑Publication Fee (if required) ❑Advance Order - # of Copies _____

4b. Method of Payment: *(Please first reapply any previously paid fee shown above)*

❑ Electronic Payment via EFS-Web ❑ Enclosed check ❑ Non-electronic payment by credit card (Attach form PTO-2038)

❑ The Director is hereby authorized to charge the required fee(s), any deficiency, or credit any overpayment to Deposit Account No. _____

5. **Change in Entity Status** (from status indicated above)

❑ Applicant certifying micro entity status. See 37 CFR 1.29

❑ Applicant asserting small entity status. See 37 CFR 1.27

❑ Applicant changing to regular undiscounted fee status.

NOTE: Absent a valid certification of Micro Entity Status (see forms PTO/SB/15A and 15B), issue fee payment in the micro entity amount will not be accepted at the risk of application abandonment.
NOTE: If the application was previously under micro entity status, checking this box will be taken to be a notification of loss of entitlement to micro entity status.
NOTE: Checking this box will be taken to be a notification of loss of entitlement to small or micro entity status, as applicable.

NOTE: This form must be signed in accordance with 37 CFR 1.31 and 1.33. See 37 CFR 1.4 for signature requirements and certifications.

Authorized Signature _____ Date _____

Typed or printed name _____ Registration No. _____

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 03/19/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER	
BOCURE, TESFALDET	
ART UNIT	PAPER NUMBER
2633	

DATE MAILED: 03/19/2021

Determination of Patent Term Adjustment under 35 U.S.C. 154 (b)
(Applications filed on or after May 29, 2000)

The Office has discontinued providing a Patent Term Adjustment (PTA) calculation with the Notice of Allowance.

Section 1(h)(2) of the AIA Technical Corrections Act amended 35 U.S.C. 154(b)(3)(B)(i) to eliminate the requirement that the Office provide a patent term adjustment determination with the notice of allowance. See Revisions to Patent Term Adjustment, 78 Fed. Reg. 19416, 19417 (Apr. 1, 2013). Therefore, the Office is no longer providing an initial patent term adjustment determination with the notice of allowance. The Office will continue to provide a patent term adjustment determination with the Issue Notification Letter that is mailed to applicant approximately three weeks prior to the issue date of the patent, and will include the patent term adjustment on the patent. Any request for reconsideration of the patent term adjustment determination (or reinstatement of patent term adjustment) should follow the process outlined in 37 CFR 1.705.

Any questions regarding the Patent Term Extension or Adjustment determination should be directed to the Office of Patent Legal Administration at (571)-272-7702. Questions relating to issue and publication fee payments should be directed to the Customer Service Center of the Office of Patent Publication at 1-(888)-786-0101 or (571)-272-4200.

OMB Clearance and PRA Burden Statement for PTOL-85 Part B

The Paperwork Reduction Act (PRA) of 1995 requires Federal agencies to obtain Office of Management and Budget approval before requesting most types of information from the public. When OMB approves an agency request to collect information from the public, OMB (i) provides a valid OMB Control Number and expiration date for the agency to display on the instrument that will be used to collect the information and (ii) requires the agency to inform the public about the OMB Control Number's legal significance in accordance with 5 CFR 1320.5(b).

The information collected by PTOL-85 Part B is required by 37 CFR 1.311. 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 30 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, Virginia 22313-1450. DO NOT SEND FEES OR COMPLETED FORMS TO THIS ADDRESS. SEND TO: Commissioner for Patents, P.O. Box 1450, Alexandria, Virginia 22313-1450. 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.

Privacy Act Statement

The Privacy Act of 1974 (P.L. 93-579) requires that you be given certain information in connection with your submission of the attached form related to a patent application or patent. Accordingly, pursuant to the requirements of the Act, please be advised that: (1) the general authority for the collection of this information is 35 U.S.C. 2(b)(2); (2) furnishing of the information solicited is voluntary; and (3) the principal purpose for which the information is used by the U.S. Patent and Trademark Office is to process and/or examine your submission related to a patent application or patent. If you do not furnish the requested information, the U.S. Patent and Trademark Office may not be able to process and/or examine your submission, which may result in termination of proceedings or abandonment of the application or expiration of the patent.

The information provided by you in this form will be subject to the following routine uses:

1. The information on this form will be treated confidentially to the extent allowed under the Freedom of Information Act (5 U.S.C. 552) and the Privacy Act (5 U.S.C 552a). Records from this system of records may be disclosed to the Department of Justice to determine whether disclosure of these records is required by the Freedom of Information Act.
2. A record from this system of records may be disclosed, as a routine use, in the course of presenting evidence to a court, magistrate, or administrative tribunal, including disclosures to opposing counsel in the course of settlement negotiations.
3. A record in this system of records may be disclosed, as a routine use, to a Member of Congress submitting a request involving an individual, to whom the record pertains, when the individual has requested assistance from the Member with respect to the subject matter of the record.
4. A record in this system of records may be disclosed, as a routine use, to a contractor of the Agency having need for the information in order to perform a contract. Recipients of information shall be required to comply with the requirements of the Privacy Act of 1974, as amended, pursuant to 5 U.S.C. 552a(m).
5. A record related to an International Application filed under the Patent Cooperation Treaty in this system of records may be disclosed, as a routine use, to the International Bureau of the World Intellectual Property Organization, pursuant to the Patent Cooperation Treaty.
6. A record in this system of records may be disclosed, as a routine use, to another federal agency for purposes of National Security review (35 U.S.C. 181) and for review pursuant to the Atomic Energy Act (42 U.S.C. 218(c)).
7. A record from this system of records may be disclosed, as a routine use, to the Administrator, General Services, or his/her designee, during an inspection of records conducted by GSA as part of that agency's responsibility to recommend improvements in records management practices and programs, under authority of 44 U.S.C. 2904 and 2906. Such disclosure shall be made in accordance with the GSA regulations governing inspection of records for this purpose, and any other relevant (i.e., GSA or Commerce) directive. Such disclosure shall not be used to make determinations about individuals.
8. A record from this system of records may be disclosed, as a routine use, to the public after either publication of the application pursuant to 35 U.S.C. 122(b) or issuance of a patent pursuant to 35 U.S.C. 151. Further, a record may be disclosed, subject to the limitations of 37 CFR 1.14, as a routine use, to the public if the record was filed in an application which became abandoned or in which the proceedings were terminated and which application is referenced by either a published application, an application open to public inspection or an issued patent.
9. A record from this system of records may be disclosed, as a routine use, to a Federal, State, or local law enforcement agency, if the USPTO becomes aware of a violation or potential violation of law or regulation.

Notice of Allowability	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--

All claims being allowable, PROSECUTION ON THE MERITS IS (OR REMAINS) CLOSED in this application. If not included herewith (or previously mailed), a Notice of Allowance (PTOL-85) or other appropriate communication will be mailed in due course. **THIS NOTICE OF ALLOWABILITY IS NOT A GRANT OF PATENT RIGHTS.** This application is subject to withdrawal from issue at the initiative of the Office or upon petition by the applicant. See 37 CFR 1.313 and MPEP 1308.

1. ☑ This communication is responsive to 03/5/2021.
   ☐ A declaration(s)/affidavit(s) under **37 CFR 1.130(b)** was/were filed on _____.

2. ☐ 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.

3. ☑ The allowed claim(s) is/are 1-8,10-18,20-29 and 31-33 . As a result of the allowed claim(s), 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.**

4. ☐ 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)).

   * Certified copies not received: _____ .

Applicant has THREE MONTHS FROM THE "MAILING DATE" of this communication to file a reply complying with the requirements noted below. Failure to timely comply will result in ABANDONMENT of this application.
THIS THREE-MONTH PERIOD IS NOT EXTENDABLE.

5. ☐ CORRECTED DRAWINGS (as "replacement sheets") must be submitted.
   ☐ including changes required by the attached Examiner's Amendment / Comment or in the Office action of Paper No./Mail Date _____ .

   Identifying indicia such as the application number (see 37 CFR 1.84(c)) should be written on the drawings in the front (not the back) of each sheet. Replacement sheet(s) should be labeled as such in the header according to 37 CFR 1.121(d).

6. ☐ DEPOSIT OF and/or INFORMATION about the deposit of BIOLOGICAL MATERIAL must be submitted. Note the attached Examiner's comment regarding REQUIREMENT FOR THE DEPOSIT OF BIOLOGICAL MATERIAL.

Attachment(s)

1. ☑ Notice of References Cited (PTO-892)
2. ☑ Information Disclosure Statements (PTO/SB/08), Paper No./Mail Date _____.
3. ☐ Examiner's Comment Regarding Requirement for Deposit of Biological Material _____.
4. ☐ Interview Summary (PTO-413), Paper No./Mail Date. _____.
5. ☑ Examiner's Amendment/Comment
6. ☑ Examiner's Statement of Reasons for Allowance
7. ☐ Other _____.

/TESFALDET BOCURE/
Primary Examiner, Art Unit 2633

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 (Notice of Allowance is in response to the amendments and remarks filed on 03/05/2021. Claims 9, 19 and 30 have been cancelled, claims 1-8, 10-18, 20-29 have been amended and new claims 31-33 have been added by the amendment of 03/05/2021. The pending claims 1-8, 10-18, 20-29 and 31-33 are allowed.

Request for Prioritized Examination (TrackOne)

3. The certification and request for prioritized examination (TrackOne) Under 37 CFR 1.102(e) filed on 03/05/2021 has been granted by the petition office on 03/10/2021 and is also noted by the Examiner.

Continued Examination Under 37 CFR 1.114

4. A request for continued examination under 37 CFR 1.114, including the fee set forth in 37 CFR 1.17(e), was filed in this application after final rejection. Since this application is eligible for continued examination under 37 CFR 1.114, and the fee set forth in 37 CFR 1.17(e) has been timely paid, the finality of the previous Office action has been withdrawn pursuant to 37 CFR 1.114. Applicant's submission filed on 03/05/2021 has been entered.

Information Disclosure Statement

5. The information disclosure statement (IDS) submitted on 03/08/2021 is in compliance with the provisions of 37 CFR 1.97. Accordingly, the information disclosure statement is being considered by the examiner. Attached with this correspondence is the initialed copy of the IDS.

Drawings

6. The drawings were received on 12/27/2019. These drawings are accepted by the Examiner.

Response to Amendment/Remarks

8. In response to the amendments and Applicant's remarks received on 07/09/2020, the following actions have been taken:

- The rejections to claims 1, 10, 11, 20, 21 and 22 under pre-AIA 35 U.S.C. 102(e) as being anticipated by Eberlein et al. (US patent publication number 2008/0200114-A1, Eberlein hereinafter) indicated in in the ***final office action*** mailed on 09/14/2020 is withdrawn.
- The rejection to claims 9, 19 and 30 under pre-AIA 35 U.S.C. 103(a) as being unpatentable over Eberlein et al. (US patent publication number 2008/0200114-A1 Eberlein hereinafter) in view of Walker et al. (US patent publication number

2005/011581-A1, Walker hereinafter) indicated in in the ***final office action*** mailed on 09/14/2020 is withdrawn.

Reasons for Allowance

9. The following is an examiner's statement of reasons for allowance: the claimed subject matters in claims 1-8, 10-18, 20-29 and 31-33 are allowed because the arts of record fail to teach or fairly suggest the claimed "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 symbol constellation to transform the received signals into received bits, and the symbol constellation includes constellation points at a plurality of unique point locations, where: the plurality of unique point locations are unequally spaced; the constellation points each have a location and a different label; and the locations of at least two of the constellation points are the same having identical locations and different labels" ***and in combinations of each and every claimed limitations*** in claims 1, 11 and 22 as disclosed and shown in figures 4b, 12d and 13b.

Any comments considered necessary by applicant must be submitted no later than the payment of the issue fee and, to avoid processing delays, should preferably accompany the issue fee. Such submissions should be clearly labeled "Comments on Statement of Reasons for Allowance."

Conclusion

10. The prior art made of record and not relied upon is considered pertinent to applicant's disclosure. Non-patent publication "Shaping Gain of LDPC Coded-QAM Transmitting Systems with Non-Uniform Constellation" published to Minhan Zheng et al. discloses a design of non-uniform constellation in use for signal transmission and reception comprising a receiver decoding the received non-uniform constellation signal using LPDC.

11. 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 https://ppair-my.uspto.gov/pair/PrivatePair. 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					
	B					
	C					
	D					
	E					
	F					
	G					
	H					
	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	Minhan Zheng, "Shaping Gain of LDPC Coded-QAM Transmitting Systems with Non-Uniform Constellation", IEEE, Retrieved March 13, 2021, pages 1-9.
	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	09/09/2020	03/13/2021						
1	1	✓	✓	=						
2	2	✓	O	=						
3	3	✓	O	=						
4	4	✓	O	=						
5	5	✓	O	=						
6	6	✓	O	=						
7	7	✓	O	=						
8	8	✓	O	=						
-	9	✓	✓	-						
9	10	✓	✓	=						
11	11	✓	✓	=						
12	12	✓	O	=						
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27	27	✓	O	=						
28	28	✓	O	=						
29	29	✓	O	=						
-	30	✓	✓	-						
10	31			=						
20	32			=						
30	33			=						

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 search updated	09/09/2020	TB
CPC search updated		

CPC Combination Sets - Searched*		
Symbol	**Date**	**Examiner**
CPC combination: (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
Combination updated	03/13/2021	TB

/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	

Search Notes	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

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
375, 455, 329, 332, 370	EAST and WEST search updated ALL ($8.ccls.)	09/09/2020	TB
375, 455, 329, 332, 370	Search updated All ($8.ccls.)	03/13/2021	

* See search history printout included with this form or the SEARCH NOTES box below to determine the scope of the search.

/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	

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 search updated	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
EAST and WEST search updated	09/09/2020	TB
EAST, Invnetor's Name, IP5, Global dossier search updated	03/13/2021	TB

Interference Search			
US Class/CPC Symbol	**US Subclass/CPC Group**	**Date**	**Examiner**
375	340	03/13/2021	TB
CPC	(H04W24/02|H04B15/00|H04L1/0003|H04L1/0009| H04L27/3405|H04L27/3483).CPC.	03/13/2021	TB

/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	

Issue Classification	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

CPC						
Symbol					Type	Version
H04W	/	24	/	02	F	2013-01-01
H04L	/	1	/	0009	I	2013-01-01
H04L	/	27	/	3405	I	2013-01-01
H04B	/	17	/	336	I	2015-01-15
H03M	/	13	/	255	I	2013-01-01
H03M	/	13	/	6325	I	2013-01-01
H04L	/	27	/	3809	I	2013-01-01
H04L	/	1	/	0003	I	2013-01-01
H04B	/	15	/	00	I	2013-01-01
H04L	/	27	/	3483	I	2013-01-01

CPC Combination Sets							
Symbol					Type	Set	Ranking	Version
	/		/					

NONE		Total Claims Allowed:	
(Assistant Examiner)	(Date)	30	
/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	13 March 2021	O.G. Print Claim(s)	O.G. Print Figure
(Primary Examiner)	(Date)	1	4b and 13b

U.S. Patent and Trademark Office — Part of Paper No.: 20210313

Issue Classification	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

INTERNATIONAL CLASSIFICATION				
CLAIMED				
H04L	/	27	/	06

NON-CLAIMED				
	/		/	

US ORIGINAL CLASSIFICATION	
CLASS	SUBCLASS
375	340

CROSS REFERENCES(S)						
CLASS	SUBCLASS (ONE SUBCLASS PER BLOCK)					

NONE		Total Claims Allowed:	
(Assistant Examiner)	(Date)	30	
/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	13 March 2021	O.G. Print Claim(s)	O.G. Print Figure
(Primary Examiner)	(Date)	1	4b and 13b

Issue Classification	Application/Control No.	Applicant(s)/Patent Under Reexamination
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	Examiner	**Art Unit**
	TESFALDET BOCURE	2633

☐ Claims renumbered in the same order as presented by applicant ☐ CPA ☐ T.D. ☐ R.1.47

CLAIMS

Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original
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-	9	18	18	27	27										

NONE		**Total Claims Allowed:**	
(Assistant Examiner)	(Date)	30	
/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	13 March 2021	O.G. Print Claim(s)	O.G. Print Figure
(Primary Examiner)	(Date)	1	4b and 13b

U.S. Patent and Trademark Office Part of Paper No.: 20210313

Web Search History

date, time	web site	search string
3/13/2021 7:25:34 PM	Google	(nonuniform constellation) (receiver)
3/13/2021 7:27:40 PM	Google	(hierarchical constellation) (PAM)
3/13/2021 7:28:28 PM	Google	(non uniform constellation) (PAM)
3/13/2021 7:29:11 PM	Google	(geometrically shaped constellation) (PAM)

EAST Search History

EAST Search History (Prior Art)

Ref #	Hits	Search Query	DBs	Default Operator	Plurals	Time Stamp
L1	340	((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	2021/03/13 18:22
L2	2	((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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L3	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L4	525	(((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	2021/03/13 18:22
L5	7	L4 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22

L6	416	L4 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L7	4	L6 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L8	23	L6 and L1	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
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	2021/03/13 18:22
L10	45	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	2021/03/13 18:22
L11	340	((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	2021/03/13 18:22
L12	2	((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	2021/03/13 18:22

L13	23	((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	2021/03/13 18:22
L14	525	(((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	2021/03/13 18:22
L15	7	L14 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L16	416	L14 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L17	4	L16 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L18	23	L16 and L11	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L19	4	L16 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	2021/03/13 18:22

L20	45	L16 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	2021/03/13 18:22
L21	155	("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	2021/03/13 18:22
L22	4	L16 and L21	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L23	340	((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	2021/03/13 18:22
L24	2	((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	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22

		near2 keying near2 modulation)) and (((low near2 densit$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))				
L25	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L26	525	(((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	2021/03/13 18:22
L27	7	L26 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L28	416	L26 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L29	4	L28 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L30	23	L28 and L23	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L31	4	L28 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	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2021/03/13 18:22

		375/332.ccls. or 375/358.ccls. or 329/103.ccls. or 332/104.ccls. 329/304.ccls. 329/305.ccls.)	EPO; JPO; DERWENT; IBM_TDB			
L32	45	L28 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	2021/03/13 18:22
L33	340	((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	2021/03/13 18:22
L34	2	((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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L35	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L36	525	(((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	2021/03/13 18:22
L37	7	L36 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2021/03/13 18:22

			EPO; JPO; DERWENT; IBM_TDB			
L38	416	L36 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L39	4	L38 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L40	23	L38 and L33	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L41	4	L38 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	2021/03/13 18:22
L42	45	L38 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	2021/03/13 18:22
L43	155	("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"	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22

		| "6084915" | "6115415" | "6157678" | "6603801" | "6606355" | "6611554" | "6665831" | "7212582" | "7245666" | "7376203" | "7620067" | "7978777" | "8265175" | "8270511" | "8842761" | "9191148" | "9385832" | "9743290" | "9743292" | "9887870" | "9967127").PN.				
L44	4	L38 and L43	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L45	431	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L46	67	L45 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L47	340	((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	2021/03/13 18:22
L48	2	((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	2021/03/13 18:22
L49	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or	US-PGPUB; USPAT;	OR	ON	2021/03/13 18:22

		(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)	USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			
L50	525	(((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	2021/03/13 18:22
L51	7	L50 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L52	416	L50 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:22
L53	4	L52 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L54	23	L52 and L47	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L55	4	L52 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	2021/03/13 18:23
L56	45	L52 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;	OR	ON	2021/03/13 18:23

			EPO; JPO; DERWENT; IBM_TDB			
L57	340	((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	2021/03/13 18:23
L58	2	((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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L59	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L60	525	(((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	2021/03/13 18:23
L61	7	L60 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L62	416	L60 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2021/03/13 18:23

			EPO; JPO; DERWENT; IBM_TDB			
L63	4	L62 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L64	23	L62 and L57	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L65	4	L62 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	2021/03/13 18:23
L66	45	L62 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	2021/03/13 18:23
L67	155	("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	2021/03/13 18:23

L68	4	L62 and L67	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L69	340	((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	2021/03/13 18:23
L70	2	((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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L71	23	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L72	525	(((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	2021/03/13 18:23
L73	7	L72 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23

L74	416	L72 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L75	4	L74 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L76	23	L74 and L69	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L77	4	L74 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	2021/03/13 18:23
L78	45	L74 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	2021/03/13 18:23
L79	340	((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	2021/03/13 18:23
L80	2	((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	2021/03/13 18:23

L81	23	((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	2021/03/13 18:23
L82	525	(((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	2021/03/13 18:23
L83	7	L82 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L84	416	L82 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L85	4	L84 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L86	23	L84 and L79	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L87	4	L84 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	2021/03/13 18:23

L88	45	L84 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	2021/03/13 18:23
L89	155	("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	2021/03/13 18:23
L90	4	L84 and L89	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L91	431	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:23
L92	67	L91 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2021/03/13 18:23

			DERWENT; IBM_TDB			
L93	7	("10848989" | "10848990" | "10863370" | "10887780").PN.	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:35
L94	4	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation))))) and L93	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2021/03/13 18:37
S1	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 20:28
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)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2020/04/05 20:33

		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))	DERWENT; IBM_TDB			
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
S12	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

S13	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S14	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) 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
S15	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
S16	7	S15 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S17	328	S15 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
S18	4	S17 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12

S19	5	S17 and S12	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:12
S20	4	S17 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
S21	43	S17 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
S22	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
S23	4	S17 and S22	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/05 22:14
S24	306	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.)	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2020/04/06 10:08

			EPO; JPO; DERWENT; IBM_TDB			
S25	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S26	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S27	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	2020/04/06 10:08
S28	7	S27 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S29	328	S27 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S30	4	S29 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS;	OR	ON	2020/04/06 10:08

			EPO; JPO; DERWENT; IBM_TDB			
S31	5	S29 and S24	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S32	4	S29 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	2020/04/06 10:08
S33	43	S29 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	2020/04/06 10:08
S34	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	2020/04/06 10:08
S35	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S36	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S37	421	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or	US-PGPUB; USPAT;	OR	ON	2020/04/06 10:08

		((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))	USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			
S38	7	S37 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S39	328	S37 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S40	4	S39 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S41	5	S39 and S34	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S42	4	S39 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	2020/04/06 10:08
S43	43	S39 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	2020/04/06 10:08
S44	154	("10149179" | "10524139" | "10530629" | "20020044597" | "20020106010" | "20030231715" | "20040022179" | "20040066738" | "20040161050" | "20040258177" | "20050180531" |	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2020/04/06 10:08

		"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.	DERWENT; IBM_TDB			
S45	4	S39 and S44	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:08
S46	374	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:27
S47	67	S46 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:28
S48	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	2020/04/06 10:49
S49	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	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2020/04/06 10:49

		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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	DERWENT; IBM_TDB			
S50	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S51	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	2020/04/06 10:49
S52	7	S51 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S53	328	S51 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S54	4	S53 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S55	5	S53 and S48	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2020/04/06 10:49

			DERWENT; IBM_TDB			
S56	4	S53 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	2020/04/06 10:49
S57	43	S53 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	2020/04/06 10:49
S58	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	2020/04/06 10:49
S59	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S60	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S61	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	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

		near2 parity near2 check) or LPDC) near3 (codes or code))				
S62	7	S61 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S63	328	S61 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S64	4	S63 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S65	5	S63 and S58	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S66	4	S63 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	2020/04/06 10:49
S67	43	S63 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	2020/04/06 10:49
S68	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" |	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

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S69	4	S63 and S68	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S70	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	2020/04/06 10:49
S71	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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S72	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S73	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	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

		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))				
S74	7	S73 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S75	328	S73 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S76	4	S75 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S77	5	S75 and S70	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S78	4	S75 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	2020/04/06 10:49
S79	43	S75 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	2020/04/06 10:49
S80	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	2020/04/06 10:49
S81	0	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or	US-PGPUB; USPAT;	OR	ON	2020/04/06 10:49

		(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$3 near2 parity near2 check) or LPDC) near3 code near3 (rate or rates))	USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			
S82	5	((Constellation near3 Designs).as. or (Barsoum near3 Maged near3 F).inv. or (Jones near3 Christopher near3 R).inv.) and ((PSK or (phase near2 shift near2 keying near2 modulation)) and (symbol near3 (constellations or constellation$))) and ((low near2 densit$3 near2 parity near2 check) or LPDC)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S83	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	2020/04/06 10:49
S84	7	S83 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S85	328	S83 and (demapper or demapping) and (decoder or decoding)	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S86	4	S85 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S87	5	S85 and S80	US-PGPUB; USPAT;	OR	ON	2020/04/06 10:49

			USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB			
S88	4	S85 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	2020/04/06 10:49
S89	43	S85 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	2020/04/06 10:49
S90	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	2020/04/06 10:49
S91	4	S85 and S90	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49
S92	374	(((non adj2 uniform) or hierarchical or uneven or unequal or unequally or ((geometrically or geometrical) near3 shaped)) near4 constellations) and ((orthogonally or orthogonal or	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO;	OR	ON	2020/04/06 10:49

		orthogonalized) same ((PSK or (phase near2 shift near2 keying near2 (modulating or modulation))) or (QAM or (quadrature near2 amplitude near3 (modulating or modulation)))))	DERWENT; IBM_TDB			
S93	67	S92 and (@ad<="20070605")	US-PGPUB; USPAT; USOCR; FPRS; EPO; JPO; DERWENT; IBM_TDB	OR	ON	2020/04/06 10:49

3/13/2021 6:38:02 PM
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Doc code: IDS
Doc description: Information Disclosure Statement (IDS) Field

PTO/SB/08a (01-10)
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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
		10848989	B2	2020-11-24	Barsoum et al.	
		10848990	B2	2020-11-24	Barsoum et al.	
		10863370	B2	2020-12-08	Barsoum et al.	
		10887780	B2	2021-01-05	Barsoum et al.	

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]
		CLEVORN ET AL., "Iterative Decoding of BICM with Non-Regular Signal Constellation Sets", January 2004, Retrieved from: https://www.researchgate.net/publication/228849016_Iterative_decoding_of_BICM_with_non-regular_signal_constellation_sets, 8 pgs.	
		DE GAUDENZI ET AL., "Adaptive coding and modulation for satellite broadband networks: From theory to practice", International Journal of Satellite Communications, Vol. 28, pp. 59-111, 2010, published online March 27, 2009, 53 pgs.	
		KAYHAN ET AL., "Constellation Design for Transmission over Nonlinear Satellite Channels", arXiv, October 5, 2012, arXiv: 1210.1762v1, 7 pgs.	
		LIOLIS ET AL., "On 64-APSK Constellation Design Optimization", 2008 10th International Workshop on Signal Processing for Space Communications, Conference Date: October 6-8, 2008, Rhodes Island, Greece, 7 pgs.	
		SCHRECKENBACH, FRANK, "Iterative Decoding of Bit-Interleaved Coded Modulation", Dissertation, Technische Universität München, 2007, 169 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
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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	/TESFALDET BOCURE/	Date Considered	03/13/2021

*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.

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Members in Patent Family (31 currently shown) ()

IP Office Type

- [x] All IP5 Offices
  - [x] EPO
  - [x] SIPO
  - [x] KIPO
  - [x] JPO
  - [x] USPTO
- [ ] Non-IP5 Offices

Office	Application	Applicant	Title	App. Dat
JP	2010511344 View Dossier ()	1) Bar *** **** F., 2) Jones Christopher Richard, 3) SHOBAYASHI, Masayuki, 4) HAYASHI, Kazuyoshi,	Methodology and method and apparatus for signaling with capacity optimized constellations	06/05/20
EP	18212572 View Dossier ()	1) Constellation Designs, LLC,	METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS	06/05/20
EP	08795885 View Dossier ()	1) Constellation Designs, Inc.,	METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS	06/05/20
US	12156989 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	METHODOLOGY AND METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS	06/05/20
US	PCT/US08/65994 View Dossier ()	1) BARSOUM, Maged, F., 2) JONES, Christopher, Richard,	DESIGN METHODOLOGY AND METHOD AND APPARATUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS	06/05/20
US	PCT/US09/69881 View Dossier ()	1) CONSTELLATION DESIGNS, INC., 2) BARSOUM, Maged, F., 3) JONES, Christopher, Richard,	METHODS AND APPARATUSES FOR SIGNALING WITH GEOMETRIC CONSTELLATIONS	12/30/20

⊞ ()	Office ⬍ ()	Application ⬍ ()	Applicant	Title ⬍ ()	App. Dat
⊞ ()	US	12650532 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	METHODS AND APPARATUSES FOR SIGNALING WITH GEOMETRIC CONSTELLATIONS	12/30/20
⊞ ()	US	13118921 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christoper R.	DESIGN METHODOLOGY AND METHOD AND APPARTUS FOR SIGNALING WITH CAPACITY OPTIMIZED CONSTELLATIONS	05/31/20
⊞ ()	US	13179383 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	METHODS AND APPARATUSES FOR SIGNALING WITH GEOMETRIC CONSTELLATIONS IN A RALEIGH FADING CHANNEL	07/08/20
⊞ ()	US	13608838 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	METHODS AND APPARATUSES FOR SIGNALING WITH GEOMETRIC CONSTELLATIONS	09/10/20
⊞ ()	US	13618630 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Design Methodology and Method and Apparatus for Signaling with Capacity Optimized Constellations	09/14/20
⊞ ()	US	14491731 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Design Methodology and Method and Apparatus for Signaling with Capacity Optimized Constellations	09/19/20

⊞ ()	Office ⬍ ()	Application ⬍ ()	Applicant	Title ⬍ ()	App. Dat
⊞ ()	US	14943003 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Methods and Apparatuses for Signaling with Geometric Constellations in a Raleigh Fading Channel	11/16/20
⊞ ()	US	15200800 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Design Methodology and Method and Apparatus for Signaling with Capacity Optimized Constellations	07/01/20
⊞ ()	US	15682475 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Systems and Methods for Transmitting Data Using Parallel Decode Capacity Optimized Symbol Constellations	08/21/20
⊞ ()	US	15682512 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Methods and Apparatuses for Signaling with Geometric Constellations in a Rayleigh Fading Channel	08/21/20
⊞ ()	US	15826579 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Methods and Apparatuses for Signaling with Geometric Constellations	11/29/20
⊞ ()	US	16206991 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Design Methodology and Method and Apparatus for Signaling with Capacity Optimized Constellations	11/30/20

⊞ ()	Office ⬍ ()	Application ⬍ ()	Applicant	Title ⬍ ()	App. Dat
⊞ ()	US	16517497 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Methods and Apparatuses for Signaling with Geometric Constellations in a Rayleigh Fading Channel	07/19/20
⊞ ()	US	16724047 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Transmitters Incorporating Uniform and Non-Uniform Constellations with Rings	12/20/20
⊞ ()	US	16724038 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Receivers Incorporating Uniform and Non-Uniform Constellations with Rings	12/20/20
⊞ ()	US	16726065 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Transmitters Incorporating Non-Uniform Multidimensional Constellations and Code Rate Pairs	12/23/20
⊞ ()	US	16726037 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Receivers Incorporating Non-Uniform Multidimensional Constellations and Code Rate Pairs	12/23/20

⊞ ()	Office ⬍ ()	Application ⬍ ()	Applicant	Title ⬍ ()	App. Dat
⊞ ()	US	16728397 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Transmitters Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	12/27/20
⊞ ()	US	16729216 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Receivers Incorporating Uniform and Non-Uniform Constellations and Adaptive Selection	12/27/20
⊞ ()	US	16729178 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Transmitters Incorporating Unequally Spaced Constellations that Provide Reduced SNR Requirements as Compared to Equally Spaced Constellations	12/27/20
⊞ ()	US	16728384 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations	12/27/20
⊞ ()	US	16729171 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	RECEIVERS INCORPORATING UNEQUALLY SPACED CONSTELLATIONS THAT PROVIDE REDUCED SNR REQUIREMENTS AS COMPARED TO EQUALLY SPACED CONSTELLATIONS	12/27/20

⊞ ()	Office ⇕ ()	Application ⇕ ()	Applicant	Title ⇕ ()	App. Date ⇕
⊞ ()	US	16729226 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Transmitters Incorporating Uniform and Non-Uniform Constellations and Adaptive Selection	12/27/2019
⊞ ()	US	16734261 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	Methods and Apparatuses for Signaling with Geometric Constellations	01/03/2020
⊞ ()	US	16752332 View Dossier ()	1) Barsoum, Maged F. 2) Jones, Christopher R.	METHODS AND APPARATUSES FOR SIGNALING WITH GEOMETRIC CONSTELLATIONS IN A RAYLEIGH FADING CHANNEL	01/24/2020

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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

Details	Results | Date & Time
64	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	16 | Mar 13, 2021, 7:41:18 PM
63	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*))	126 | Mar 13, 2021, 7:40:23 PM
62	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	8 | Mar 13, 2021, 7:39:51 PM
61	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND ((maximum OR minimize OR minimizing) NEAR/4 (signal NEAR/2 noise NEAR/2 ratio) OR SNR)) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	0 | Mar 13, 2021, 7:39:40 PM
60	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	8 | Mar 13, 2021, 7:39:25 PM
59	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	8 | Mar 13, 2021, 7:38:31 PM
58	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	8 | Mar 13, 2021, 7:33:57 PM
57	(((nonuniform) OR hierarchical OR uneven OR unequal OR unequally OR (geometrical* NEAR/3 shaped)) NEAR/4 constellation) AND ((maximum OR minimize OR minimizing) NEAR/4 (signal NEAR/2 noise NEAR/2 ratio) OR SNR)) AND (QAM OR (quadrature NEAR/2 amplitude NEAR/2 modulat*)) AND (low NEAR/2 density NEAR/2 parity NEAR/2 check)	0 | Mar 13, 2021, 7:33:36 PM

Bibliographic data: US 2020221321 (A1)

Receivers incorporating uniform and non-uniform constellations with rings

Publication date:	9 July 2020	
Inventor(s):	BARSOUM MAGED F [US]; JONES CHRISTOPHER R [US]	
Applicant(s):	CONSTELLATION DESIGNS LLC [US]	
Classifications:	**International:**	***H04W24/02***; ***H04B15/00***; ***H04L1/00***; ***H04L27/34***
	Cooperative:	H03M13/255 (US); H03M13/6325 (US); H04B15/00 (US); H04B17/336 (US); H04L1/0003 (EP,US); H04L1/0009 (EP,US); H04L27/3405 (EP,US); H04L27/3483 (US); H04L27/3809 (US); H04W24/02 (US); Y02D30/50 (EP)
Application number:	US201916724038 20191220	
Priority number(s):	US201916724038 20191220	
	US201816206991 20181130	
	US201715682475 20170821	
	US201615200800 20160701	
	US201414491731 20140919	
	US201213618630 20120914	
	US201113118921 20110531	
	US20080156989 20080605	
	US20070933319P 20070605	

Abstract of US 2020221321 (A1)
Communication systems are described that use unequally spaced constellations that have increased capacity compared to conventional constellations operating within a similar SNR band. One embodiment is a digital communications system including a transmitter transmitting signals via a communication channel, the transmitter including a coder capable of receiving user bits and outputting encoded bits at a rate, a mapper capable of mapping encoded bits to symbols in a constellation, and a modulator capable of generating a modulated signal for transmission via the communication channel using symbols generated by the mapper, wherein the constellation is unequally spaced and characterizable by assignment of locations and labels of constellation points to maximize parallel decode capacity of the constellation at a given signal-to-noise ratio so that the constellation provides a given capacity at a reduced signal-to-noise ratio compared to a uniform constellation that maximizes the minimum distance between constellation points of the uniform constellation.

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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 03/30/2021

KPPB LLP
2190 S. Towne Centre Place
Suite 300
Anaheim, CA 92806

EXAMINER
BOCURE, TESFALDET

ART UNIT	PAPER NUMBER
2633	

NOTIFICATION DATE	DELIVERY MODE
03/30/2021	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

UNITED STATES DEPARTMENT OF COMMERCE
U.S. Patent and Trademark Office
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APPLICATION NO./ CONTROL NO.	FILING DATE	FIRST NAMED INVENTOR/ PATENT IN REEXAMINATION	ATTORNEY DOCKET NO.
16/728,384	12/27/2019	Barsoum et al.	C44-00999.CON15

	EXAMINER	
KPPB LLP 2190 S. Towne Centre Place Suite 300 Anaheim, CA 92806	TESFALDET BOCURE	
	ART UNIT	PAPER
	2633	20210325

DATE MAILED: _____

Please find below and/or attached an Office communication concerning this application or proceeding.

Commissioner for Patents

Applicant's representative Mr. David J. Bailey (Reg. #59,632) called Examiner on 03/24/2021 regarding the reasons for allowance mailed on 03/19/2021, where a deleted claimed limitation, having identical locations and different labels, was included in the reasons for allowance. Here below is the corrected reasons for allowance excluding the deleted limitation " having identical locations and different labels".

Reasons for Allowance
8. The following is an examiner's statement of reasons for allowance: the claimed subject matters in claims 1-8, 10-18, 20 -29 and 31-33 are allowed because the arts of record fail to teach or fairly suggest the claimed "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 symbol constellation to transform the received signals into received bits, and the symbol constellation includes constellation points at a plurality of unique point locations, where: the plurality of unique point locations are unequally spaced; the constellation points each have a location and a different label; and the locations of at least two of the constellation points are the same" and in combinations of each and every claimed limitations in claims 1, 11 and 22 as disclosed and shown in figures 4b, 12d and 13b.
Any comments considered necessary by applicant must be submitted no later than the payment of the issue fee and, to avoid processing delays, should preferably accompany the issue fee. Such submissions should be clearly labeled " Comments on Statement of Reasons for Allowance."

/TESFALDET BOCURE/
Primary Examiner, Art Unit 2633

PTO-90C (Rev.04-03)

Applicant-Initiated Interview Summary	**Application No.** 16/728,384	**Applicant(s)** Barsoum et al.		
	Examiner TESFALDET BOCURE	**Art Unit** 2633	**AIA (First Inventor to File) Status** No	**Page** 1 of 1

All Participants (applicant, applicants representative, PTO personnel)	**Title**	**Type**
TESFALDET BOCURE	Primary Examiner	Telephonic
David J. Bailey (Reg. #59,632)	Attorney of Record	

Date of Interview: 25 March 2021

Issues Discussed:

Other

Correction to the "Reasons for Allowance" mailed on 03/19/2021.

/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	

Applicant is reminded that a complete written statement as to the substance of the interview must be made of record in the application file. It is the applicants responsibility to provide the written statement, unless the interview was initiated by the Examiner and the Examiner has indicated that a written summary will be provided. See MPEP 713.04
Please further see:
MPEP 713.04
Title 37 Code of Federal Regulations (CFR) § 1.133 Interviews, paragraph (b)
37 CFR § 1.2 Business to be transacted in writing

Applicant recordation instructions: The formal written reply to the last Office action must include the substance of the interview. (See MPEP section 713.04). If a reply to the last Office action has already been filed, applicant is given a non-extendable period of the longer of one month or thirty days from this interview date, or the mailing date of this interview summary form, whichever is later, to file a statement of the substance of the interview.

Examiner recordation instructions: Examiners must summarize the substance of any interview of record. A complete and proper recordation of the substance of an interview should include the items listed in MPEP 713.04 for complete and proper recordation including the identification of the general thrust of each argument or issue discussed, a general indication of any other pertinent matters discussed regarding patentability and the general results or outcome of the interview, to include an indication as to whether or not agreement was reached on the issues raised.

PART B – FEE(S) TRANSMITTAL

Complete and send this form, together with the applicable fee(s), to: Mail **Mail Stop ISSUE FEE**
Commissioner for Patents
P.O. Box 1450
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or Fax **(571)-273-2885**

INSTRUCTIONS: This form should be used for transmitting the ISSUE FEE and PUBLICATION FEE (if required). Blocks 1 through 5 should be completed where appropriate. All further correspondence including the Patent, advance orders and notification of maintenance fees will be mailed to the current correspondence address as indicated unless corrected below or directed otherwise in Block 1, by (a) specifying a new correspondence address; and/or (b) indicating a separate "FEE ADDRESS" for maintenance fee notifications.

CURRENT CORRESPONDENCE ADDRESS (Note: Use Block 1 for any change of address)

71897 7590 03/19/2021

KPPB LLP
2190 S. Towne Centre Place, Suite 300
Anaheim, CA 92806

Note: A certificate of mailing can only be used for domestic mailings of the Fee(s) Transmittal. This certificate cannot be used for any other accompanying papers. Each additional paper, such as an assignment or formal drawing, must have its own certificate of mailing or transmission.

Certificate of Mailing or Transmission

I hereby certify that this Fee(s) Transmittal is being deposited with the United States Postal Service with sufficient postage for first class mail in an envelope addressed to the Mail Stop ISSUE FEE address above, EFS-Web transmitted, or facsimile transmitted to the USPTO (571) 273-2885, on the date indicated below.

Katrina Lee	(Depositor's Name)
/Katrina Lee/	(Signature)
April 9, 2021	(Date)

APPLICATION NO	FILING DATE	FIRST NAMED INVENTOR	ATTORNEY DOCKET NO	CONFIRMATION NO
16/728,384	December 27, 2019	Maged F. Barsoum	C44-00999.CON15	6350

TITLE OF INVENTION:
Receivers Incorporating Non-Uniform Constellations with Overlapping Constellation Point Locations

APPLN. TYPE	ENTITY STATUS	ISSUE FEE DUE	PUBLICATION FEE DUE	PREV. PAID ISSUE FEE	TOTAL FEE(S) DUE	DATE DUE
nonprovisional	undiscounted	$1200	$0	$0	$1200	June 19, 2021

EXAMINER	ART UNIT	CLASS-SUBCLASS
Tesfaldet Bocure	2633	375-262000

1. Change of correspondence address or indication of "Fee Address" (37 CFR 1.363).

☐ Change of correspondence address (or Change of Correspondence Address form PTO/SB/122) attached.

☐ "Fee Address" indication (or "Fee Address" Indication form PTO/SB/47; Rev 03-02 or more recent) attached. **Use of a Customer Number is required.**

2. For printing on the patent front page, list
(1) The names of up to 3 registered patent attorneys or agents OR, alternatively,
(2) The name of a single firm (having as a member a registered attorney or agent) and the names of up to 2 registered attorney or agent. If no name is listed, no name will be printed

1. KPPB LLP
2.
3.

3. ASSIGNEE NAME AND RESIDENCE DATA TO BE PRINTED ON THE PATENT (print or type)
PLEASE NOTE: Unless an assignee is identified below, no assignee data will appear on the patent. If an assignee is identified below, the document has been filed for recordation as set forth in 37 CFR 3.11. Completion of this form is NOT a substitute for filing an assignment.

1a. NAME OF ASSIGNEE (B) RESIDENCE: (CITY and STATE OR COUNTRY)

Constellation Designs, LLC — Anaheim, California

Please check the appropriate assignee categories (will not be printed on the patent): ☐ Individual ☑ Corporation or other private group entity ☐ Government

4a. The following fee(s) are submitted:
☑ Issue Fee
☐ Publication Fee (No small entity discount permitted)
☐ Advance Order - # of Copies

4b. Payment of Fee(s): **(Please first reapply and previously paid fee shown above)**
☐ A check is enclosed
☐ Payment by credit card. Form PTO-2038 is attached
☑ The director is hereby authorized to charge the required fee(s), any deficiency, or credits any overpayment, to Deposit Account Number 50-4407 (enclose an extra copy of this form)

5. Change of Entity Status (from status indicated above)

☐ Applicant certifying micro entity status. See 37 CFR 1.29 — NOTE: Absent a valid certification of Micro Entity Status (see forms PTO/SB/15A and 15B), issue fee payment in the micro entity amount will not be accepted at the risk of application abandonment.

☐ Applicant asserting small entity status. See 37 CFR 1.27 — NOTE: If the application was previously under micro entity status, checking this box will be taken to be a notification of loss of entitlement to micro entity status.

☐ Applicant changing to regular undiscounted fee status. — NOTE: Checking this box will be taken to be notification of loss of entitlement to small or micro entity status, as applicable.

NOTE: This form must be signed in accordance with 37 CFR 1.31 and 1.33. See 37 CFR 1.4 for signature requirements and certifications.

Authorized Signature /Isaac M. Fine/ Date April 9, 2021

Typed or printed name Isaac M. Fine Registration No. 74,274

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:				
UTILITY APPL ISSUE FEE	1501	1	1200	1200

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Extension-of-Time:				
Miscellaneous:				
			Total in USD ($)	**1200**

Electronic Acknowledgement Receipt	
EFS ID:	42417065
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-APR-2021
Filing Date:	27-DEC-2019
Time Stamp:	16:53:54
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	yes
Payment Type	CARD
Payment was successfully received in RAM	$1200
RAM confirmation Number	E202149G54241606
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	Issue Fee Payment (PTO-85B)	00999con15issuefeextml.pdf	107671 1e29c83566051a41e9e0146aaa2f9a733969471c	no	1
Warnings:					
Information:					
2	Fee Worksheet (SB06)	fee-info.pdf	30721 9f7193235066f8a7a8739c39708748087b07194f	no	2
Warnings:					
Information:					
		Total Files Size (in bytes):	138392		

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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.

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
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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

Change(s) applied to document, /P.P/ 4/9/2021

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Issue Classification	Application/Control No.	Applicant(s)/Patent Under Reexamination
	16/728,384	Barsoum et al.
	Examiner	Art Unit
	TESFALDET BOCURE	2633

☐ Claims renumbered in the same order as presented by applicant ☐ CPA ☐ T.D. ☐ R.1.47

CLAIMS

Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original	Final	Original
1	1	9	10	-	19	28	28								
2	2	11	11	19	20	29	29								
3	3	12	12	22 ~~21~~	21	-	30								
4	4	13	13	21 ~~22~~	22	10	31								
5	5	14	14	23	23	20	32								
6	6	15	15	24	24	30	33								
7	7	16	16	25	25										
8	8	17	17	26	26										
-	9	18	18	27	27										

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NONE		Total Claims Allowed:	
(Assistant Examiner)	(Date)	30	
/TESFALDET BOCURE/ Primary Examiner, Art Unit 2633	13 March 2021	O.G. Print Claim(s)	O.G. Print Figure
(Primary Examiner)	(Date)	1	4b and 13b

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ISSUE NOTIFICATION

The projected patent number and issue date are specified above.

Determination of Patent Term Adjustment under 35 U.S.C. 154 (b)
(application filed on or after May 29, 2000)

The Patent Term Adjustment is 0 day(s). Any patent to issue from the above-identified application will include an indication of the adjustment on the front page.

If a Continued Prosecution Application (CPA) was filed in the above-identified application, the filing date that determines Patent Term Adjustment is the filing date of the most recent CPA.

Applicant will be able to obtain more detailed information by accessing the Patent Application Information Retrieval (PAIR) WEB site (http://pair.uspto.gov).

Any questions regarding the Patent Term Extension or Adjustment determination should be directed to the Office of Patent Legal Administration at (571)-272-7702. Questions relating to issue and publication fee payments should be directed to the Application Assistance Unit (AAU) of the Office of Data Management (ODM) at (571)-272-4200.

APPLICANT(s) (Please see PAIR WEB site http://pair.uspto.gov for additional applicants):

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