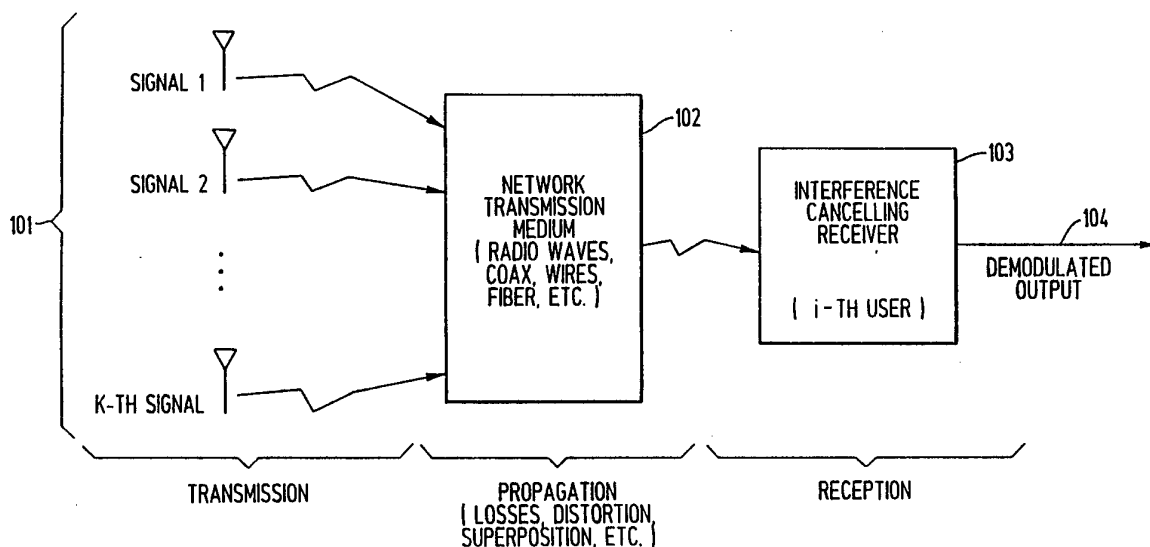




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(54) Title: METHOD AND APPARATUS FOR INTERFERENCE CANCELLATION IN A MULTIPLE ACCESS COMMUNICATION RECEIVER

**(57) Abstract**

A multiaccess communication system for cellular telephony, portable telephones, mobile radio communications, broadcast radio and television, wireless communication networks and the like, including planned use of interference cancelling receiving methods and apparatus which permit multiple users to share the same or overlapping frequency assignments by estimating the undesired signals in a family of such signals so that they may be cancelled to receive any desired signal which is selected from the family. Various cross-coupled phase locked loop receivers, both analog and digital, employing inphase, inphase and quadrature, feed forward and difference amplitude tracking configurations are disclosed and may be employed, by way of example, in the practice of the invention.

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Method and Apparatus for Interference Cancellation in a Multiple Access Communication Receiver

BACKGROUND OF THE INVENTION

This invention relates generally to improvements
5 in methods and apparatus for increasing the capacity of
multiaccess communication systems, such as cellular
telephony, portable telephones, mobile radio communica-
tions, broadcast radio and television, wireless communica-
tion networks and the like, and, more particularly, to a
10 new and improved system for harnessing interfering signals
to increase system capacity for multiple users.

The modern world is awash with multiuser
communications channels. The demand for the existing
spectrum is, as such, relentless. From mobile communica-
15 tions, to cordless phones, to commercial broadcast
stations, systems must make allowances for interferences.

Various approaches to increasing capacity in such
multiaccess communication systems include expanded fre-
quency and/or spatial diversity, using more than one
20 frequency or antenna to mitigate interference. However,
this requires more spectrum or higher cost to cover the
increased number of antennas required.

Code Division Multiple Access (CDMA) has also
been suggested to increase communication system capacity.
25 In CDMA, each signal uses a much larger bandwidth than is
required to transmit the desired information, and a coding
scheme to choose the desired signal from multiple signals
in the same bandwidth. Unfortunately, this requires rather
complicated receivers which must receive much more band-
30 width than is normally required for each user. Some forms
of CDMA are also known as spread spectrum multiple access,

in which the wide-band codes make interfering signals appear to be white noise to the desired signal receiver. In this latter technique, the required bandwidth of the desired signal is greatly expanded. On reception, the
5 desired signal's bandwidth is collapsed back to the information-bearing bandwidth, while the interfering signal's bandwidth remains spread. Hence, in such a modulation technique, the interference is not cancelled or removed. It still reduces the network capacity by contributing to the noise. The information-bearing structure of
10 the interference is not used in any way to mitigate this undesirable effect.

Prior art systems dealing with interference generally tend to cancel interference to receive a single
15 selected signal, rather than harness interfering signals to provide increased capacity in a multiaccess system. Such prior art approaches attempt to mitigate the effect of a generally weaker interference signal on the reception of a generally stronger desired signal. The interference in
20 such prior art system is typically totally discarded. Modulation or information contained in the interference is generally not used in any way in its mitigation. In those instances where an estimate of the interference is used to remove it, prior art systems require a second channel, a
25 so-called "reference" input of the interference, in order to cancel the interference. The prior art may separately demodulate a desired signal, but does not attempt to improve overall network capacity by jointly demodulating multiple, interfering signals.

30 One of the main advantages of wide-deviation, frequency-modulated (FM) signals is the FM threshold. Because of this, the detected signal-to-noise ratio (SNR) is markedly improved if the input is above a certain

received SNR, with the amount and position of the improvement threshold being determined by the bandwidth expansion as given by the ratio of the FM deviation to the modulating signal's bandwidth. In an environment where the main
5 source of "noise" is interfering FM signals, usually co-channel, this effect, known as the FM capture effect, facilitates a simple method of providing multiaccess communications by dividing the service area into geographical regions. The allocation of frequencies for FM
10 transmitters has been traditionally accomplished in this manner. The allowable geographical separation of co-channel and adjacent channel FM transmitters, be they high-power broadcast stations or relatively low-power cellular telephones, has been regulated so that propagation laws
15 will mitigate any interference problems.

However, as the frequency spectrum becomes more crowded, the search for more efficient uses of the available bandwidth intensifies. Complex digital combined modulation and coding schemes have been proposed to allow
20 more robust communications in a denser interference environment. By allowing additional receiver complexity, many of these techniques provide greater spectral efficiency by actively mitigating the interference through advanced coding and modulation. Among these are the
25 aforescribed spread spectrum and other code division multiple access (CDMA) techniques. The addition of interference cancellation at the receiver input of such systems, fed by the decoded output, would allow the resulting system to tolerate still more interference in the channel.

30 A natural target for any technique to improve the spectral efficiency of FM is, of course, the cellular telephone channel. The number of users of cellular telephony has grown dramatically, beyond almost all

expectations, putting an increasing burden on the spectrum allocated for this service. The existing bandwidth allocation for cellular telephony effectively limits the number of subscribers who can be provisioned in any one service area at any one time. Competition with other uses for the frequency spectrum make additional allocations both expensive and difficult to obtain. Reuse of transmission frequencies across small areas, called "cells," is what improves the traffic handling capacity of a cell-based telephone system. However, the ability of the current United States cellular telephone system to handle traffic is constrained not only by the available spectrum, but also by frequency reuse rules designed to maintain call quality in the presence of interfering signals from the surrounding cells. The application of interference-cancelling receivers, in accordance with the present invention, to such cellular channels will decrease the minimum frequency reuse distance, allowing the system to carry more traffic on the same allocated spectrum, i.e., more channels per cell. In addition, such a change in the cellular system could be implemented without the lengthy and expensive regulatory procedures required for changing the modulation schemes or obtaining additional bandwidth.

Even a commercial broadcast station can be considered a multiuser communications system over a large geographical area that includes multiple stations assigned to the same frequency. Most of these systems face much less severe reception environments than the cellular telephone channel. In more general mobile communications applications, the required level of voice quality can be far below cellular telephone standards, e.g., on a police, aircraft, or taxicab radio. Application of interference cancellers to these channels, in accordance with the present invention, would not only allow several users to

5 speak at once, something not possible on current systems, but would also allow the allocated bandwidth to support more user licenses. Application to these mobile channels would be similar to the cellular telephone channel, since the cellular telephone channel is also a mobile channel.

10 For cordless phones the rapid fading nature of the cellular environment, primarily because of multiple signal paths of different doppler shift does not occur. Users of cordless phones in areas of high user density would also benefit from the application of interference
15 cancellers, in accordance with the present invention, since these communications systems operate on a small number of frequencies and are commonly observed to interfere frequently with one another. For commercial broadcast
20 stations, the high signal-to-noise ratios can be taken advantage of, combined with more complicated encoding or modulation at the transmitter. Commercial stations could be placed closer together geographically, and listeners could then choose which station they wanted to hear by
25 selecting, in accordance with the present invention, from the plurality of interference-canceller outputs available to each listener. All these applications would allow more efficient use of the spectrum over geographically distributed areas and satisfy a long existing need for expanded, high quality multiaccess communication systems.

Hence, those concerned with the development and use of multiaccess communication systems have long recognized the need for improved communication systems which simply, reliably and economically enable increased system
30 capacity, with minimal interference, from a given frequency spectrum thereby facilitating enhanced signal reception to a greater number of users. The present invention clearly fulfills all of these needs.

SUMMARY OF THE INVENTION

Briefly, and in general terms, the present invention provides a new and improved multiaccess communication system embodying novel methods and apparatus for
5 extracting a selected signal from a larger family of such signals having some relationship to the selected signal and overlapping predictable electrical characteristics wherein a desired signal to be received is selected from the family of signals, an estimate is made of the magnitude of at
10 least one predetermined electrical parameter in the remaining unselected signals of the family which would normally interfere with the selected signal, and the estimate is used to substantially cancel the unselected signals and facilitate enhanced reception of the desired
15 signal. In this way, any signal in the family of signals may be designated as the desired signal to be received to thereby increase system quality, capacity and availability to an increased number of users. Hence, what would normally be interference is selectively harnessed to
20 increase the availability of overlapping signals to multiple users.

Fading, shadowing, and other propagation effects, prevent the guarantee of a desired signal power level at the receiver either relative to the interference level or
25 absolute. As a result, it is not always possible to capture the desired signal with prior art techniques. However, interference cancellation, in accordance with the current invention, removes interference, guaranteeing enhanced reception of the desired signal, and is robust to
30 these propagation effects. Moreover, planning and designing to allow such interference to increase capacity for multiaccess is entirely new and represents a breakthrough

for expansion of communications systems without the need for expanded frequency and/or spatial diversity.

Network capacity can be increased if multiple signals can overlap in transmission parameters, e.g., frequency or polarization. With the exception of spread spectrum techniques, the prior art often relies on filtering a separable physical transmission parameter, such as frequency or polarization. Inadequate initial separation, e.g., frequency overlap or insufficient cross-polarization isolation, result in system degradation. By not relying on an initial separating transmission parameter and instead removing the interference based on the modulation, propagation and/or information bearing characteristics of the signal, the system of the present invention is freed from such constraints. Signals can overlap in frequency, and need not be separated by polarization.

Moreover, a second "reference" channel input is not needed in the system of the present invention to cancel the interference. In practice, such a channel, which includes only the interference and not the desired signal, is not available with only a single antenna and receiver.

Prior art systems, in spread spectrum multiple access, add additional bandwidth to the desired signal at transmission, necessitating sometimes larger spectrum allocations and always more complex transmission and reception equipment. The system of the present invention uses the inherent information redundancy present in the modulation format, the modulating signal, or the channel path-induced distortions of almost all practical systems to separate interfering signals and, hence, does not add additional bandwidth to the signal at transmission.

In accordance with the invention, the estimation of an interfering waveform is used to cancel interference in the form of undesired signals wherein any or all of the signals may be information bearing signals or constructed signals having known and predictable characteristics, as opposed to "white noise" having totally unpredictable characteristics. Such interference cancellation yields multiple selective access to the pool of signals for extraction of desired signals by multiple users.

By way of example, and not by way of limitation, the aforescribed invention may be embodied in various analog and digital receiver configurations, utilizing cross-coupled phase locked loop technology, in configurations embodying inphase, inphase and quadrature, feed forward and difference amplitude-tracking electrical networks.

The new and improved multiaccess communication system of the present invention is unique in that it harnesses interference to increase system capacity available for plural access by single or multiple users. The improved system provides receivers which are independent of any power hierarchy in that desired signal extraction does not depend on relative position of the desired signal in the power hierarchy. The system can detect the weakest signal in the family of overlapping signals, and the relative power hierarchy need not be stable or preserved. Hence, the present invention expands the useful bandwidth in multiaccess systems and satisfies a long existing need in the communications industry for accurate, reliable, easy to use and economical systems for exploiting the frequency spectrum to provide increased system capacity for multiple users.

These and other objects and advantages of the invention will become apparent from the following more detailed description when taken in conjunction with the accompanying drawings of illustrative embodiments.

5

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is an overall system block diagram of a network system incorporating features of the present invention to achieve increased capacity;

10 FIG. 2 is a block diagram of a simple receiver subsystem embodying the invention;

FIG. 3 is a block diagram illustrating an alternative embodiment of a receiver subsystem in accordance with the invention;

15 FIG. 4 is a block diagram of another embodiment of a receiver subsystem in accordance with the invention;

FIG. 5 is a conceptual block diagram of an FM Multi-User Receiver embodying features of the invention;

20 FIG. 6 is a block diagram illustrating independent interference estimation for plural interfering signals;

FIG. 7 is a block diagram illustrating vector sum estimation in a receiver subsystem;

FIG. 8 is a block diagram illustrating a basic Cross-Coupled Phase Locked Loop receiver configuration;

FIG. 9 is a block diagram illustrating Estimator-Subtractor Interference Cancellation for a receiver;

FIG. 10 is a block diagram illustrating an Optimum (MAP, or Maximum A Posteriori) Receiver for FM with
5 interchannel interference;

FIG. 11 is a block diagram illustrating a CCPLL (cross-coupled phase locked loop) with Amplitude and Leakage Control;

FIG. 12 is a block diagram illustrating a
10 receiver subsystem embodying a Feedforward (Direct Form) Optimum Amplitude Estimator;

FIG. 13 is a block diagram illustrating a receiver subsystem embodying a Feedback (Gradient-Descent Form) Optimum Amplitude Estimator;

FIG. 14 is a block diagram illustrating a
15 receiver subsystem embodying Difference-Amplitude Tracking configuration in accordance with the invention;

FIG. 15 is a basic block diagram illustrating a digital system embodying a presently preferred embodiment
20 of the invention;

FIG. 16 is a block diagram of an analog to digital conversion subsystem suitable for use with the system of Fig. 15;

FIG. 17 is a block diagram illustrating an
25 interference cancelling FM demodulator suitable for use with the system of Fig. 15;

FIG. 18 is a block diagram for a phase detector suitable for use with the system of Fig. 15;

FIG. 19 is a block diagram illustrating the loop gain and filter for the system shown in Fig. 15;

5 FIG. 20 is a block diagram illustrating an amplitude estimator suitable for use with the system of Fig. 15;

10 FIG. 21 is a block diagram illustrating a 1st Order Digital Butterworth Lowpass Filter suitable for use with the system of Fig. 15; and

FIG. 22 is a block diagram illustrating tone detectors and channel selection for the system of Fig. 15.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

15 It has been said that the radio spectrum is like real estate; they are not making any more of it. However, the important quantity to conserve is not necessarily the bandwidth itself, but its practical ability to carry as much information as possible. In geographically distributed systems, wasting the recoverable information present
20 in strong interfering signals is, quite literally, wasting the channel's capacity for carrying information. Even in high-capacity systems where coding techniques allow users to communicate on the same frequency, it is potentially feasible to improve interference resistance by decoding
25 (demodulating), re-encoding, and then cancelling the interference.

In accordance with the invention, it has been discovered, that with the current modulation specification

for the cellular industry, a frequency may even be reused within its own cell. In this way, FM interference cancellation can easily provide at least a doubling of the call-carrying capacity of the allocated spectrum under the
5 current modulation specification. Current results indicate that there are prospects for even greater rewards in terms of increased spectral efficiency.

As the cellular communications industry grows, the need to utilize the available spectrum efficiently becomes increasingly vital. The need to accommodate an
10 ever larger subscriber base has pushed the development of co-frequency channel interference-resistant multiple-access schemes. These schemes must provide greater call capacity for geographic regions, allowing for future expansion.
15 However, they are significantly more complicated than the conventional frequency division multiple access (FDMA) FM techniques. Indeed, the basic receiver design in current systems is at least a decade old and, comparing its performance to that of a CDMA system, is something like
20 comparing today's desktop personal computers to programmable calculators of ten years ago. In December 1989, the United States Federal Communications Commission issued General Docket No. 88-411, "In the Matter of Advanced Technologies for the Public Safety Radio Services" in which
25 they queried whether further advances in analog technology could improve spectral efficiency. One answer to that question is the development of interference-cancelling receivers in accordance with the present invention.

The present invention provides new and improved
30 multiaccess communication systems such as cellular telephony, cordless phones, mobile radio, broadcast radio and the like, embodying novel methods and apparatus for extracting a selected signal from a larger family of such

signals having some relationship to the selected signal and overlapping predictable electrical characteristics wherein a desired signal to be received is selected from the family of signals, an estimate is made of the magnitude of at least one predetermined electrical parameter in the remaining unselected signals of the family which would normally interfere with the selected signal, and the estimate is used to substantially cancel the unselected signals and facilitate enhanced reception of the desired signal. In this way, any signal in the family of signals may be designated as the desired signal to be received to thereby increase system quality, capacity, and availability to an increased number of users. Hence, what would normally be interference is selectively harnessed to increase the availability of overlapping signals to multiple users.

In accordance with the invention, the estimation of an interfering waveform is used to cancel interference in the form of undesired signals wherein any or all of the signals may be information bearing signals or constructed signals having known and predictable characteristics, as opposed to "white noise" having maximum unpredictability. Such selective interference cancellation yields multiple access to the pool of undesired signals for selective extraction of desired signals by multiple users. The improved system provides receivers which are independent of any power hierarchy in that the desired signal extraction does not depend on relative position of the desired signal in the power hierarchy. The system can detect the weakest signal in the family of overlapping signals, and the relative power hierarchy need not be stable or preserved.

By way of example, and not by way of limitation, the aforescribed invention may be embodied in various

analog and digital receiver configurations, utilizing cross-coupled phase locked loop technology, in configurations embodying inphase, inphase and quadrature, feedforward and difference amplitude-tracking electrical
5 networks.

Referring now to Fig. 1 of the drawings, there is shown a network system which achieves increased capacity by planned use of interference cancelling receivers to receive each user's desired signal in the presence of interference.
10 By way of example, a plurality of antennas 101 propagate corresponding signals (Signal 1, Signal 2 ... Signal K) via any suitable network transmission medium 102, as by radio waves, coaxial cable, wires, fiber optics, or the like, characterized by the usual losses, distortion, superposi-
15 tion, etc. The resultant signals are received by an interference cancelling receiver 103, or a plurality of such receivers and demodulated output for a selected signal is available on line 104 for the i th user in a family of multiple users.

20 In Fig. 2 of the drawings, interference estimates are subtracted from a single input channel. In this regard, an estimate of the interference is provided as negative feedback, over line 106, from an estimator/demodulator subsystem 105 to a summing junction 107
25 where it is effectively subtracted from the input to the junction on line 108. The resultant is directed to the demodulation subsystem 105 over line 109. In the most general form, demodulation and estimation subsystems can be either combined or separate processes.

30 Referring now specifically to Fig. 3, wherein like reference numerals denote like or corresponding parts throughout the drawings, there is shown an alternative

embodiment wherein the demodulation and estimation processes are separate. When they are fully separated, or only demodulation information from the desired signal is used in the estimation process, a demodulator is only
5 needed for the desired user's signal. In the subsystem of Fig. 3, the demodulator 105a, for the i th user receives input over line 109a from the summing junction 107, while the interference estimator 105b for all users except the i th user, receives such input over line 109b. The negative
10 feedback to the junction 107, over line 106, is provided solely by the estimator 105b. A further alternative embodiment, which does not use demodulator information, will be subsequently described in connection with the subsystem shown in Fig. 9 of the drawings.

15 As best observed in Fig. 4, in still another alternative embodiment, demodulation and estimation are functionally combined, as is the case in our cross-coupled FM interference cancelling receivers. In this embodiment, a demodulator/estimator 105c, 105d is needed for each
20 interfering signal.

In accordance with the invention, cancellation of interfering frequency-modulation (FM) signals has been developed with emphasis towards applications on the cellular telephone channel as an important example of a
25 multiple access communications system. Results in the field of interference channels are shown to motivate the estimation and subtraction of undesired interfering signals. Hence, this invention is best understood from an examination of the relative optimality of the current FM
30 techniques in known interference channels, before pursuing the estimation and subtracting of interfering FM signals.

As previously indicated, one of the main advantages of wide-deviation, FM signals is the FM threshold effect. Because of this, the detected SNR is markedly improved if the input is above a certain received SNR, with the amount and position of the improvement threshold being determined by the bandwidth expansion as given by the ratio of the FM deviation to the modulating signal's bandwidth. In an environment where the main source of "noise" is interfering FM signals, usually co-channel, this effect, known as the FM capture effect, facilitates a simple method of providing multiaccess communications by dividing the service area into geographical regions. The allocation of frequencies for FM transmitters has been traditionally accomplished in this manner. The allowable geographical separation of co-channel and adjacent channel FM transmitters, be they high-power broadcast stations or low-power cellular telephones, has been regulated so that propagation laws will mitigate any interference problems.

As the frequency spectrum becomes more crowded, the search for more efficient uses of the available bandwidth intensifies. In accordance with one aspect of the invention, the capture-effect phenomenon of FM reception is exploited to produce simple interference-cancelling receivers with a cross-coupled topology. Phase-locked loop receivers cross-coupled with amplitude-tracking loops to estimate the FM signals are utilized in various embodiments of the invention. The function of these cross-coupled phase-locked loop (CCPLL) interference cancelling receivers is explained below. New interference cancellers inspired by optimal estimation and the CCPLL topology are described, resulting in simpler receivers than those in prior art.

Some preferred embodiments for FM interference cancelling receivers, in accordance with the invention, are:

- 5 1. Difference-amplitude tracking Inphase-only and Inphase-and-Quadrature forms of the CCPLL interference cancelling receiver, as will subsequently be described in greater detail in connection with the embodiments of Figs. 14-22;
- 10 2. Feedforward Inphase-only and Inphase-Quadrature forms of the CCPLL, e.g. as in the embodiment of Fig. 11;
3. Feedback Inphase-only form of the CCPLL, as in the embodiment of Fig. 12; and
- 15 4. A variation of the CCPLL with amplitude and leakage control as in Fig. 11. This approach may be also implemented in any of the form variants in categories 1 through 3, as well as the feedback Inphase-and-Quadrature form.

20 FM interference-cancelling receivers are considered for increasing the frequency reuse in a cellular telephone system. Interference mitigation in the cellular environment is seen to require tracking of the desired signal during time intervals when it is not the strongest signal present. Use of interference cancelling in conjunction with dynamic frequency-allocation algorithms is viewed
25 as a way of improving spectrum efficiency. Performance of such interference cancellers indicates possibilities for greatly increased frequency reuse. The known ability of an FM broadcast receiver to unambiguously receive one station
30 or another at the same frequency in areas of interference leads one to conclude that there exist situations where, in

accordance with the present invention, both mutually interfering signals could potentially be received with the right receiver. Referring now more particularly to Fig. 5, demodulated FM information, enhanced by the capture effect, is provided as output over line 104a from a capture effect receiver 100a and first is remodulated as an estimate of the interference. This estimate is then subtracted at junction 112 from the input to a second capture effect receiver 100b, leaving the originally weaker signal now the stronger and allowing its capture by the second receiver.

By using the remodulated phase estimates, the amplitude component of each interfering signal is estimated, producing an estimate of the entire interfering signal. By tracking both the amplitude and phase of the interfering signals, the full separation of the interfering signals can be exploited. This can be contrasted with tracking only the phase, as is done in conventional broadcast FM receivers. Phase-only tracking can be seen as projecting the resultant of the interfering signals onto the surface of a sphere of constant amplitude in signal space. The loss of sometimes essential side information contained in the resultant's amplitude plagues these receivers. Such "power-division multiplexing" of FM signals has been proposed before, but the separation methods involved tuning a frequency domain "trap" filter, and are, therefore, fundamentally different from the coherent methods described here.

As previously indicated, the estimation of an interfering waveform is used to cancel interference in the form of undesired signals wherein any or all of the signals may be information bearing signals or constructed signals having known and predictable characteristics, as opposed to "white noise". Such selective interference cancellation

yields multiple access to the pool of undesired signals for selective extraction of desired signals by multiple users.

There are two primary approaches for expanding the interference cancelling receiver to accommodate multiple interfering signals. These are by forming independent estimates of each interfering signal and adding them together to cancel out the sum total interference, as shown in Fig. 6 of the drawings, and by estimating the vector sum of the interference as a single estimate for each set of interfering signals, as illustrated in Fig. 7 of the drawings. Both of these methods differ from successive cancellation of interfering signals in that a "power hierarchy" need not be maintained.

Power hierarchy is maintained by direct measurement and tracking, with or without additional power control. In this regard, successive interference cancelling relies on transmitted power control to maintain a strict power hierarchy. Interference cancelling receivers, in accordance with the invention, do not require a strict power hierarchy because they directly measure and track the amplitudes of the received signals. However, power control can still be added on top of this tracking to boost the power of signals that are shadowed from the receiver, with rapidly varying power levels being tracked by the receiver. Such successive interference cancelling receivers fitted with the direct measurement and tracking of power are considered within the scope of the present invention.

Referring again to Figs. 6 and 7 of the drawings, there are shown two methods for expanding interference cancelling receivers, in accordance with this invention, to, by way of example, three interfering signals in a family comprising signals S, I1 and I2.

In Fig. 6, independent interference estimators 105d are provided for each of the interference signals from the family of signals. Fig. 6 illustrating only one of the three possible branches of this interference cancelling receiver. The other two branches (not shown) would differ only in that they would have interference estimators for signals I1 and S or I2 and S.

In Fig. 7, rather than providing separate estimators for each interfering signal, a vector sum interference estimator 105e is provided. Again, as in the case of Fig. 6, only one of the three branches is shown for purposes of simplified illustration.

In accordance with the invention, a receiver can also estimate interfering signals by a means of phase estimation. In this connection, the phase-locked loop is not the only way to estimate the phase of an interfering signal. Other phase estimators would produce equivalent systems. For example, systems with frequency-locked loops together with inphase and quadrature amplitude estimation would produce the same result. Moreover, non-tracking signal phase estimators such as are used in digital phase modulated signal reception could also be used to estimate the phase of the interference.

A receiver can also estimate interfering signals by a means of amplitude estimation. The correlators employed in our preferred embodiment are not the only way to estimate the amplitude of an interfering signal. Other amplitude estimators would produce equivalent systems. For example, the several different types of the analog FM interference canceller differ only in the amplitude estimators (feedforward inphase-only, feedforward inphase and quadrature, difference amplitude tracking inphase-only,

difference amplitude tracking inphase and quadrature, feedback inphase-only). Other amplitude estimation means, such as filtering for a pilot tone and detecting its power, can be used to produce an equivalent system.

5 Still another possible receiver separates interfering signals by means of estimating the propagation loss. Estimation of the amplitudes of the received signals in our interference cancelling receivers is, in fact, a way of estimating the channel propagation loss of each signal.
10 Other methods of channel estimation would be equivalent. Interfering signals would then be separated by identifying the channel loss characteristics associated with each of them.

 For example, other methods of channel estimation
15 includes estimation of the channel impulse response or transfer function, as is done in equalization, and separating interfering signals using the fact that they have passed through different propagation channels. Another method would be estimation of the received signal power and
20 separation of the signals by use of the different fading and shadowing environments. Both of these estimation techniques could be used in a tracking mode, where the estimate is updated over time using previous estimates or, in a non-tracking mode, where new estimates are formed as
25 needed without regard for the previous estimate. Prior art contains many examples of equalization and channel estimation. In accordance with the invention, the channel estimate is not being used only to improve a single desired signal, but also as an identifying characteristic to
30 separate and estimate multiple interfering signals for cancellation.

Referring now to Fig. 8 of the drawings, there is shown a basic generalized configuration for a CCPLL. The CCPLL interference canceller shown is an example of subtracting an estimate of interfering signals from the incoming corrupted signal to produce a clean output. The CCPLL configuration includes a junction 113a, phase locked loop 114a and gain control 115a cross-coupled to a junction 113b, phase locked loop 114b and gain control 115b.

As observed in Fig. 9, the estimator-subtractor system is quite general, similar to Fig. 2 previously described, and can be applied to many different interference situations including FM multipath, co-channel FM interference, and AM-FM interference. Getting the best possible estimate of the interference environment is the most important and one of the most challenging aspects in developing one of these interference cancellers. Estimation approaches vary from those that utilize extensive a priori knowledge of the input to those that make as few restrictions as possible on the input signals.

The CCPLL interference canceller is based on the topology of a noncausal optimum (Maximum a Posteriori (MAP)) FM receiver for suppression of interchannel interference with constant modulus (no AM component). Given an input signal mix:

25

$$v(t) = s_1(t) + s_2(t) + n(t),$$

where:

$$s_1(t) = A_1 \cos(\omega_1 t + x_1(t))$$

$$s_2(t) = A_2 \cos(\omega_2 t + x_2(t))$$

$n(t)$ is white Gaussian noise with zero mean and normalized two-sided power spectral density of 1 W/Hz.

$x_1(t)$ is the desired-signal angle modulation (here assumed a Gaussian process), for FM.

$x_1(t) = \int^t m_1(u) du$, where $m_1(t)$ is the desired-signal frequency modulation information.

$x_2(t)$ is the interfering signal angle modulation (also assumed Gaussian).

A_1 and A_2 are the moduli of the two signals and are assumed constant.

In the MAP receiver, estimates of $x_1(t)$ and $x_2(t)$ are:

$$\hat{x}_1 = k_1 \int_0^T R_{x_1} \frac{\partial \hat{s}_1}{\partial x_1} (v - \bar{s}_2) du$$

$$\hat{x}_2 = k_2 \int_0^T R_{x_2} \frac{\partial \hat{s}_2}{\partial x_2} (v - \bar{s}_1) du,$$

where $k_{1,2}$ are constants, $R_{x_{1,2}}(t, u)$ are the impulse responses of lowpass filters, and $\bar{s}_{1,2}$ represent the minimum mean-square error (MMSE) estimates of $s_1(t)$ and $s_2(t)$. Referring now to Fig. 10 of the drawings, there is shown a correlator implementation of the optimum receiver for x_1 . Since for FM, $m_{1,2}(t) = \dot{x}_{1,2}(t)$, where $m_{1,2}(t)$ represents the desired information, the phase-controlled oscillator 117 in Fig. 10 can be replaced, in practice, with a voltage-controlled oscillator (VCO). The result is a representation of the optimum receiver for $m_{1,2}(t)$ similar to a phase-locked loop (PLL). This optimum receiver topology inspires the use of the phase-locked loop structure as a causal approximation. However, even with noncausal correlation loops, the receiver is optimum only when the generated estimates are the MMSE estimates; i.e., $\hat{s}_{1,2}(t) = \bar{s}_{1,2}(t)$.

The CCPLL system utilizes two important characteristics of the phase-locked loop (PLL) to produce its estimate:

1. A PLL is an unbiased estimator of phase.

2. PLL's, like other FM demodulators, exhibit FM capture effect and nonlinearly enhance the strongest signal when more than one signal is present.

The characteristics (loop bandwidths, gain control, etc.) of the CCPLL interference canceller can be fine-tuned to best accommodate different types of interferers. However, this requires a priori knowledge of both the input desired signal and the interfering signals. Modifications of the system to include amplitude-tracking loops inspired by Least Mean Squares (LMS) adaptive filters have been reported to produce a slightly more complex system that is fairly robust in the presence of co-channel AM-FM and pulsed RF (FM) interferers. The performance of CCPLL interference cancellers has been reported against these types of interferers both in simulation and by experiment, providing as much as 20 dB of interference suppression.

The addition of LMS-type amplitude-tracking loops to the CCPLL interference canceller transforms the device from one based primarily on the FM capture effect into a more standard, adaptive noise-cancelling device. In the adaptive noise cancelling topology (not shown), often called the "adaptive notch filter," a periodic reference is used in both inphase and quadrature (90 degrees phase-shifted) forms to cancel correlated components in the input. The LMS algorithm is used to produce the inphase and quadrature multiplicative weights necessary to cancel the reference's component from the input.

The realization of amplitude-control loops for the CCPLL interference canceller closely parallels the adaptive notch-filter approach. This is motivated by the convergence of the LMS solution to a minimum mean-square

error estimate, and the MAP receiver topology's requirement that the subtracted interference estimate be an MMSE one.

The implementation of the CCPLL interference canceller with amplitude control may also be in the form
5 so-called Vector-locked Loops (not shown). The resulting implementation is a so-called self-bootstrapping, continuous analog version of the discrete digital adaptive filter, with the reference inputs being provided by the VCO's in the capture effect PLL's. The system is called self-
10 bootstrapping because it can acquire the interfering signals by itself because of the FM capture effect. Experimental trials with the CCPLL system have reported that acquisition is a two-stage process, as might be expected. Software simulations have been reported,
15 confirming this finding. The first stage consists of a PLL capturing and locking to the strongest signal present. The second phase is marked by the convergence of the amplitude control weights. The two phases are then repeated by the second PLL, resulting in FM demodulation of both the weaker
20 and the stronger signal.

Because of error and delay in frequency tracking of the PLL's, as well as imperfect capture, the inputs to the amplitude estimators often contain some leakage from the undesired signal. This cross leakage affects the
25 convergence of the amplitude-control loops and results in further imperfections in cancellation. Because the CCPLL structure contains a reference for each signal present, it is tempting to think of recursively applying the adaptive noise canceller in an attempt to limit cross leakage. A
30 successful second application of the adaptive noise canceller, cross-coupling the VCO references themselves, would serve to decrease components of the undesired signal in each reference and, therefore, be potentially more

robust. Such a configuration would have an additional feedback path within each PLL structure as shown in Figure 11 which depicts a CCPLL with amplitude controls 115a, 115b and leakage controls 120a, 120b. However, before considering this "leakage-control" topology further, it is helpful to understand the effect of the leakage on the normal CCPLL canceller.

The effect of a residual, strong signal component in the VCO output of the weak signal tracking PLL 114b (referred to as PLL #2) on its amplitude-control loop 115b can be considered as leakage of signal into the reference input of an adaptive notch filter. This will cause some cancellation of the strong signal prior to the input of the strong signal tracking PLL 114a (PLL #1). The strong signal is denoted s_1 and the weak signal s_2 ; the transfer function for the weak signal to the VCO of PLL #2 is $H_{2,2}(f)$. Consider the two signals to be uncorrelated and stationary. Similarly, let the transfer function for the s_1 into the VCO of PLL #2 be $H_{1,2}(f)$, and let the transfer functions for s_1 and s_2 into the VCO output of PLL #1 be $H_{1,1}(f)$ and $H_{2,1}(f)$, respectively. Neglecting intermodulation terms, the power spectral density of the input to the amplitude-control loop #2, the VCO output of PLL #2, denoted x_2 , is approximately:

$$S_{x_2, x_2}(f) = S_{s_2, s_2}(f) |H_{2,2}(f)|^2 + S_{s_1, s_1}(f) |H_{1,2}(f)|^2.$$

The cross spectrum between the input to the system, $(s_1 + s_2)$, and the VCO output of PLL #2, x_2 , is:

$$S_{x_2, (s_1 + s_2)}(f) = S_{s_2, s_2}(f) H_{2,2}^*(f) + S_{s_1, s_1}(f) H_{1,2}^*(f)$$

where * denotes complex conjugation. Because of the feedback in the CCPLL structure, a causal solution must be achieved as opposed to a noncausal solution.

However, examination of the noncausal case will demonstrate the same behavior seen for the real causal system. It is interesting to note that the noncausal analysis leads to an exact expression for the ratio of strong signal power to weak signal power at the inputs of the PLL's (under stationary signal conditions) in an uncoupled system. The signal power ratio follows a power inversion law. For the uncoupled system, the unconstrained so-called Wiener solution for the power ratio at the input of PLL #1 is

$$SIR_{in_1} \left(\frac{s_1}{s_2} \right) = \frac{1}{SIR_{out_1} \left(\frac{s_1}{s_2} \right)}$$

and, at the input to PLL #2,

$$SIR_{in_2} \left(\frac{s_2}{s_1} \right) = \frac{1}{SIR_{out_2} \left(\frac{s_2}{s_1} \right)}$$

Heuristically, these expressions indicate that a canceller with desired signal leaking into its reference of the interference can do no better than to give an output signal-to-interference ratio equal to the reciprocal of the signal-to-interference ratio in the reference. Hence, strong desired-signal leakage into the interference reference will significantly limit the achievable cancellation.

By examining these two equations, we can perhaps gain some insight into how two identically designed PLL's would behave during acquisition. If both PLL's were to initially capture the same signal, say s_1 , each would decrease the other's $SIR_{in} \left(\frac{s_1}{s_2} \right)$ according to the power-inversion relations above. Both amplitude loops would be

headed for convergence at a point, making s_2 the stronger signal. FM capture effect provides gain preferentially to the stronger signal at the input of the PLL. One PLL input, say PLL #2, must cross the capture threshold for s_2 first, making its $SIR_{out}(\frac{s_1}{s_2})$ less than one. This would reverse the direction of convergence for the amplitude-control loop #2, which uses that PLL's output as a reference. Thus, $SIR_{in}(\frac{s_1}{s_2})$ for the PLL #1 increases, reinforcing its capture of s_1 . The FM capture effect further reinforces $SIR_{out}(\frac{s_1}{s_2})$, and provides the feedback necessary to complete the self-bootstrapping process.

The actual causal system acquires the signal in much the same way, according to approximate solutions to the coupled differential equations that define the system and signal-acquisition experiments reported by others.

The system shown in Fig. 11 adds "leakage-control loops" 120a and 120b to the VCO outputs of the two PLL's. These adaptive noise-cancelling loops attempt to better orthogonalize the two VCO references, increasing capture in the interference-cancelling reference. The use of a second pair of adaptive noise-cancelling loops would cause the SIR at the input of the PLL's to converge to the inverse square of the SIR produced by the PLL's, rather than just to the inverse of the SIR at the PLL output. The resulting system has decreased acquisition time.

The addition of the leakage-control loops 120a, 120b in Fig. 11 creates a stable, steady-state solution where both PLL's are tracking the stronger signal. This new stable state is equivalent to an uncoupled system. While forcing the VCO outputs to be uncorrelated with respect to each other, the leakage-control loops produce

the degenerate solution where both VCO outputs are uncorrelated to either signal. This, in turn, causes little or no cancellation of the strong signal at the input to either VCO, and strong-signal capture occurs in both PLL's, reinforcing the stable state. The leakage-control loops configuration is not used in the preferred embodiment.

If a CCPLL-based interference-cancelling receiver is to be extended to more than two signals, simply tracking the amplitude of each received signal is not necessarily optimal. This is because a minimum mean-square error estimate for the sum total interference is desired, and the MMSE estimate of a sum is not necessarily the sum of the individual MMSE estimates. As a result, interference amplitude-tracking loops are associated with each desired signal rather than with each interfering signal. The amplitudes of the interfering signal components to be subtracted from the input are estimated jointly for each desired signal in method paralleling multireference adaptive interference-cancellation techniques. If the VCO-produced references of the summed interferers are uncorrelated with each other, the estimation of the amplitudes of the summed interference will be the same as separately estimating the amplitudes of the component signals.

Amplitude estimation for only the inphase portion of each signal should suffice, in some instances, reducing the hardware of the CCPLL dramatically. Implementation of this system can be attempted either with a feedforward amplitude estimator using the form of the estimator shown in Fig. 12 of the drawings, or in a feedback loop form shown in Fig. 13 by using the gradient-descent form of the optimal solution.

The amplitude estimator in Fig. 12 adds a pair of lowpass filters 121a, 121b in a direct form, feedforward optimum amplitude estimator wherein the input and phase estimate are combined to produce an amplitude estimate of the interference, and where the amplitude estimator does not form a loop including the actual cancellation or subtractor operator.

The amplitude estimator in Fig. 13 also uses lowpass filters 121a, 121b, but in a gradient-descent form, feedback optimum amplitude estimator wherein the input interference is cancelled by the previous estimate of interference. In this regard, the residual is correlated against the phase estimate of interference to produce the next iteration or update of the amplitude interference estimate.

The direct form shown in Fig. 12 simplifies the dynamics of the system by replacing a feedback loop by a feedforward loop. In this case, the weight estimate is just the lowpass portion of the product of the original input signal (without any cancellation) and the inphase estimate. The lowpass filters 121a, 121b are to have a cutoff frequency high enough to pass the amplitude information, which is typically near the lower band edge of the modulating information. The estimator is a coherent AM receiver with the PLL's coherent phase reference as its coherent carrier reference. The coherent AM receiver thus tracks the amplitude of the FM signal just the same as if it were modulated on a constant-frequency carrier.

The gradient-descent solution has the form:

$$\vec{w}_{k+1} = \vec{w}_k - \frac{\beta}{2} \frac{\partial}{\partial \vec{w}} (E[e^2]) \big|_{\vec{w} = \vec{w}_k}.$$

Here $E[e^2]$ denotes the expected value over the ensemble of noise. For the special case given here, the negative of the gradient is just $2E[x_1(v-w_1x_1)] = E[x_1e]$. Hence, the gradient-descent solution in Fig. 13 is just the inphase part of the amplitude-tracking loop given in the iterative CCPLL. However, the gradient-descent implementation form requires integrators 122a, 122b which the direct form does not require.

Each of the implementations shown in Figs. 12 and 13 offers a reduction in both hardware and dynamical complexity. They do not contain a limiter prior to each PLL's input. However, for the steady-state case where the interfering signal is mostly cancelled, the limiter has almost no effect and the equations presented here are valid. The defining equations are:

$$\begin{aligned}\dot{x}_{1I} &= \beta_1[A_1\cos(\psi_1-\phi_1) + A_2\cos(\psi_2-\phi_1) - x_{1I}] * g_{1I} \\ \dot{x}_{1Q} &= \beta_1[A_1\sin(\psi_1-\phi_1) + A_2\sin(\psi_2-\phi_1) - x_{1Q}] * g_{1Q} \\ \dot{x}_{2I} &= \beta_2[A_1\cos(\psi_1-\phi_2) + A_2\cos(\psi_2-\phi_2) - x_{2I}] * g_{2I} \\ \dot{x}_{2Q} &= \beta_2[A_1\sin(\psi_1-\phi_2) + A_2\sin(\psi_2-\phi_2) - x_{2Q}] * g_{2Q} \\ \dot{\phi}_1 &= \alpha_1[A_1\sin(\psi_1-\phi_1) + A_2\sin(\psi_2-\phi_1) - x_{2I}\sin(\phi_2-\phi_1) - x_{2Q}\cos(\phi_2-\phi_1)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2[A_1\sin(\psi_1-\phi_2) + A_2\sin(\psi_2-\phi_2) - x_{1I}\sin(\phi_1-\phi_2) - x_{1Q}\cos(\phi_1-\phi_2)] * h_{L2}.\end{aligned}$$

Here the input is of the form:

$$v(t) = A_1\cos(\omega_c t + \psi_1) + A_2\cos(\omega_c t + \psi_2).$$

The first four equations specify the amplitude-control loops, while the last two specify the phase-locked loops. All the variables are real, with the amplitude estimates being represented by $x_{(1,2)I,Q}$ and the phase estimates by $\phi_{1,2}$. The symbol $*$ denotes convolution. The loop filters for PLL's 1 and 2 are, respectively, h_{L1} and h_{L2} , and the lowpass filters in the amplitude-control loops are g_{1I} , g_{1Q} , g_{2I} , and g_{2Q} . The PLL gains are α_1 and α_2 , and the amplitude-control loop gains are β_1 and β_2 .

The addition of limiters would have a significant effect during signal acquisitions by enhancing the strong and suppressing the weak signals. Since the amplitude-control loops are completely preceding the limiter, the limiter does not affect their operation except through its effect on the phase estimates. As a result, in a system with bandpass limiters, the first four differential equations above are unchanged. However, since the limiter nonlinearly mixes the additive signal components before they reach the PLL, the operations of the PLL cannot be distributed among the additive signal components as before. This is because the limiter has already distorted the linear combination of these components before they get to the PLL. Without the limiter the PLL's multiplication of its input by the VCO's output ($-\sin\phi_1$ or $-\sin\phi_2$) could be arithmetically distributed among the additive input components to form the product terms whose lowpass portions result in the sines and cosines of phase differences seen above.

With the limiter in place, however, this distribution and its corresponding trigonometric reduction and lowpass filtering to sines and cosines of phase differences cannot take place. The three operations, limiter, bandpass filter, and VCO multiplier are noncommutative, and therefore, the combination prevents further reduction of the phase-tracking differential equations. With the limiter in place, the equations therefore take on the direct form of the implementation:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1[\sin(\phi_1)BPLIM(A_1\cos\psi_1 + A_2\cos\psi_2 - x_{2I}\cos\phi_2 - x_{2Q}\cos\phi_2)]*h_{L1} \\ \dot{\phi}_2 &= \alpha_1[\sin(\phi_2)BPLIM(A_1\cos\psi_1 + A_2\cos\psi_2 - x_{1I}\cos\phi_1 - x_{1Q}\cos\phi_1)]*h_{L2},\end{aligned}$$

where BPLIM(y) denotes the bandpass limiter's operation on y.

The gradient-descent form of the reduced solution, derived earlier and shown in Fig. 13, follows by noting that its only difference from the system defined by the six equations above is that the quadrature amplitude estimates are set identically zero. The equations are therefore obtained by eliminating the terms multiplied by x_{1Q} and x_{2Q} and eliminating the equations for x_{1Q} and x_{2Q} . Since there is now only one amplitude estimate per signal, the I, Q subscripts will be dropped. As a result, the reduced gradient-descent solution is governed by the simplified equations given below (again for negligible noise):

$$\begin{aligned}\dot{x}_1 &= \beta_1[A_1\cos(\psi_1-\phi_1) + A_2\cos(\psi_2-\phi_1) - x_1] * g_1 \\ \dot{x}_2 &= \beta_2[A_1\cos(\psi_1-\phi_2) + A_2\cos(\psi_2-\phi_2) - x_2] * g_2 \\ \dot{\phi}_1 &= \alpha_1[A_1\sin(\psi_1-\phi_1) + A_2\sin(\psi_2-\phi_1) - x_2\sin(\phi_2-\phi_1)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2[A_1\sin(\psi_1-\phi_2) + A_2\sin(\psi_2-\phi_2) - x_1\sin(\phi_1-\phi_2)] * h_{L2}.\end{aligned}$$

The use of the direct feedforward form shown Fig. 12 of the optimal MMSE amplitude estimate simplifies the system even more, leaving just the two nonlinear equations for the phase-locked loops. In the direct feedforward form, amplitude estimates x_1 and x_2 are computed directly as follows:

$$\begin{aligned}x_1 &= [(A_1\cos\psi_1 + A_2\cos\psi_2)\cos\phi_1] * g_1 \\ x_2 &= [(A_1\cos\psi_1 + A_2\cos\psi_2)\cos\phi_2] * g_2\end{aligned}$$

where g_1 and g_2 are lowpass filters as before. Eliminating the double frequency terms gives:

$$\begin{aligned}x_1 &= [A_1\cos(\psi_1-\phi_1) + A_2\cos(\psi_2-\phi_1)] * g_1 \\ x_2 &= [A_1\cos(\psi_1-\phi_2) + A_2\cos(\psi_2-\phi_2)] * g_2.\end{aligned}$$

Substituting these expressions into the phase-locked equations for ϕ_1 and ϕ_2 results in the two differential equations that define the feedforward system:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1[A_1\sin(\psi_1-\phi_1) + A_2\sin(\psi_2-\phi_1) - [(A_1\cos(\psi_1-\phi_2) + A_2\cos(\psi_2-\phi_2))*g_2]\sin(\phi_2-\phi_1)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2[A_1\sin(\psi_1-\phi_2) + A_2\sin(\psi_2-\phi_2) - [(A_1\cos(\psi_1-\phi_1) + A_2\cos(\psi_2-\phi_1))*g_1]\sin(\phi_1-\phi_2)] * h_{L2}.\end{aligned}$$

It is instructive to consider a perturbation of the feedforward system from a steady-state solution. Let the input amplitudes A_1 and A_2 be constant, and let the driving input phase functions be:

$$\psi_1 = \omega_1 t \quad \psi_2 = \omega_2 t + \epsilon.$$

Since the phase-locked loops under consideration are second order or better ($h_{L1,2}$ each has at least one pole), the loops will track the input phase rate (instantaneous frequency) as well as the input phase. We will consider a perturbation to the phase variable and consider the average phase rates tracked by the loops in order to remain in lock. In real systems this approach is further justified by the time constants associated with the loop lowpass filter, which are noted for the ability to "flywheel" the loop through relatively short noise bursts. Therefore, let the loop initial conditions be:

$$\phi_1 = \omega_1 t + \delta \quad \phi_2 = \omega_2 t + \gamma.$$

Then the differential equations are:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1[A_1\sin(-\delta) + A_2\sin(\Delta\omega t + \epsilon - \delta) - [(A_1\cos(-\Delta\omega t - \gamma) + A_2\cos(\epsilon - \gamma))*g_2]\sin(\Delta\omega t + \gamma - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2[A_1\sin(-\Delta\omega t - \gamma) + A_2\sin(\epsilon - \gamma) - [(A_1\cos(-\delta) + A_2\cos(\Delta\omega t + \epsilon - \delta))*g_1]\sin(-\Delta\omega t - \gamma + \delta)] * h_{L2}\end{aligned}$$

where $\Delta\omega = \omega_2 - \omega_1$. Let g_1 and g_2 be narrowband filters so that they cut off below frequency $\Delta\omega$, as they would be for tracking amplitude functions whose bandwidths are less than those of the modulating information on an FM signal. The differential equations then simplify to:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1[A_1\sin(-\delta) + A_2\sin(\Delta\omega t + \epsilon - \delta) - A_2\cos(\epsilon - \gamma)\sin(\Delta\omega t + \gamma - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2[A_1\sin(-\Delta\omega t - \gamma) + A_2\sin(\epsilon - \gamma) - A_1\cos(-\delta)\sin(-\Delta\omega t - \gamma + \delta)] * h_{L2}.\end{aligned}$$

If $\Delta\omega$ is outside the pass band of the lowpass loop filters h_{L1} and h_{L2} , the system displays the dynamics of ordinary, first-order phase locked loops without any coupling. More importantly, the system behaves as if the interference were not present:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1 A_1 \sin(-\delta) \\ \dot{\phi}_2 &= \alpha_2 A_2 \sin(\epsilon - \gamma).\end{aligned}$$

For small perturbations, ϵ , δ , and γ , the system can be linearized to:

$$\begin{aligned}\dot{\phi}_1 &= -\delta \alpha_1 A_1 \\ \dot{\phi}_2 &= -(\gamma - \epsilon) \alpha_2 A_2\end{aligned}$$

which contains a sink at $(\delta = 0, \gamma = \epsilon)$. Since δ and $\gamma - \epsilon$ are the instantaneous phase errors of ϕ_1 and ϕ_2 , respectively, the response of the linearized system will be to correct the phase-error perturbation with an exponential decay according to the equations:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1 A_1 (\psi_1 - \phi_1) \\ \dot{\phi}_2 &= \alpha_2 A_2 (\psi_2 - \phi_2).\end{aligned}$$

Further examination of the two feedforward equations leads one to consider what happens if the phase tracks coincide ($\phi_1 = \phi_2$), especially in the case where both PLL's might momentarily acquire the same signal. In this case, it would appear that the system acts as if there were no cross-coupling at all, since the term that is due to the cross-coupling is multiplied by $\sin(\phi_1 - \phi_2)$. Note that for a system with bandpass limiters, the nonlinear action of the limiter prevents the trigonometric reduction that results in the factor of $\sin(\phi_1 - \phi_2)$. This term comes from the multiplication of the phase-locked loop's VCO with the subtracted estimate of the interference. In a system with a limiter, strong signal enhancement occurs prior to the

PLL input. If the interference estimate is relatively close to the interference in frequency and phase, the nonlinear enhancement is based on the strength of the interference minus the estimate relative to that of the desired signal. As a result, if the phase tracks of the interferers were only to coincide momentarily ($\phi_1 = \phi_2$), the interference and its estimate together are suppressed at the input to the PLL, facilitating desired-signal capture and preventing the possibility of nullifying the effect of cross-coupling.

However, this examination of this case for the system without limiters is of interest during signal acquisition, specifically for the event where both loops capture the same signal. Consider a small perturbation of this case where $\phi_1 = \phi_2$. Let the input be:

$$\psi_1 = \omega_1 t \quad \psi_2 = \omega_2 t + \epsilon.$$

Let the initial conditions be:

$$\phi_1 = \omega_1 t + \delta \quad \phi_2 = \omega_1 t + \gamma, \quad \gamma \neq \delta.$$

The differential equations are then:

$$\begin{aligned} \dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) + A_2 \sin(\Delta\omega t + \epsilon - \delta) - [A_1 \cos(-\gamma) + A_2 \cos(\Delta\omega t + \epsilon - \gamma)] g_2] \sin(\gamma - \delta) h_{L1} \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\gamma) + A_2 \sin(\Delta\omega t + \epsilon - \gamma) - [A_1 \cos(-\delta) + A_2 \cos(\Delta\omega t + \epsilon - \delta)] g_1] \sin(\delta - \gamma) h_{L2}. \end{aligned}$$

Assuming that $g_{1,2}$ and $h_{L1,2}$ remove the terms of frequency $\Delta\omega$, as before:

$$\begin{aligned} \dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) - A_1 \cos(-\gamma) \sin(\gamma - \delta)] \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\gamma) - A_1 \cos(-\delta) \sin(\delta - \gamma)]. \end{aligned}$$

By applying trigonometric identities and combining terms, the following equations are obtained:

$$\begin{aligned} \dot{\phi}_1 &= \frac{\alpha_1 A_1}{2} [\sin(-\delta) - \sin(2\gamma - \delta)] \\ \dot{\phi}_2 &= \frac{\alpha_2 A_1}{2} [\sin(-\gamma) - \sin(2\delta - \gamma)]. \end{aligned}$$

Let the two loops be identical so that $\alpha_1 = \alpha_2 = \alpha$, and define $G = A_1 \alpha$. For small perturbations, the system of equations can be linearized to:

$$\dot{\phi}_1 = -G \gamma \qquad \dot{\phi}_2 = -G \delta,$$

5 which by definition of ϕ_1 and ϕ_2 is the coupled system:

$$\dot{\phi}_1 = G (\psi_1 - \phi_2) \qquad \dot{\phi}_2 = G (\psi_1 - \phi_1).$$

At this point we can consider the dynamics of the system. There are two cases to consider, one in which γ and δ are of the same sign, and one where they are of different signs. When the two errors are of the same sign, the linearization indicates that both errors will decrease in absolute value. However, since the rate of change of ϕ_1 is proportional to the phase error of ϕ_2 (γ) rather than to its own phase error δ , and ϕ_2 is likewise cross-controlled, the decrease will not return to an equilibrium at $\delta = \gamma = 0$. Without loss of generality, let $|\delta| < |\gamma|$. Because the initial absolute value of the phase error in ϕ_1 is less than that in ϕ_2 , and the rate of correction of ϕ_1 is greater than that in ϕ_2 , ϕ_1 will reach the point of zero phase error while ϕ_2 still has nonzero phase error. At this point we have the following conditions for the rates of change and the phase errors:

$$\begin{array}{ll} \dot{\phi}_1 \neq 0 & \dot{\phi}_2 = 0 \\ \delta = 0 & \gamma \neq 0. \end{array}$$

Clearly, the phase error for ϕ_1 , δ , will cross over zero while the phase error for ϕ_2 , γ , is still on the original side of zero, providing us the perturbation case where δ and γ are of different signs.

When δ and γ are of different signs, the sign of ϕ_2 , γ , will increase with time, causing the rate of drift of both phase-locked loops to increase. This positive

feedback will continue to accelerate the drift and shortly invalidates the linearization of the differential equations. Since the coupled differential equations increase the absolute value of small phase errors when the phase errors are of opposite sign, the equilibrium at $\phi_1 = \phi_2$, i.e., capture of the same signal by each loop, is unstable.

For the moment, consider the linearized dynamics of this drift away from the state where both loops are locked on the same signal. Without loss of generality let $|\delta| < |\gamma|$. From the linearization above, since the rate of growth of $|\delta|$ is proportional to $|\gamma|$, and the rate of growth of $|\gamma|$ is proportional to $|\delta|$, the linearized system will tend to correct the imbalance in absolute value. Hence there are negative feedback dynamics in the growth of the absolute phase errors, tending to balance them so that $\gamma \approx -\delta$. These dynamics will continue to cause both $|\delta|$ and $|\gamma|$ to grow at the same increasing rate and value, until the linearization breaks down.

It has been assumed that the average frequency of the phase-locked loops was constant. As both absolute phase errors begin not only to increase monotonically, but also to accelerate in opposite directions, this condition is violated. The oscillator frequencies of the two phase-locked loops begin to separate, one decreasing and the other increasing. The first parts of the linearization to break down are the linear approximations for $\sin(2\gamma - \delta)$ and $\sin(2\delta - \gamma)$. Assuming that $|\gamma| \approx |\delta|$ near the breakdown and using a standard cutoff for the linearization of the sine function to be $\frac{\pi}{6}$, the linearization is good for absolute phase errors $< \frac{\pi}{18}$. For a loop with a gain (G) of about 30,000 (loop lock range = 30 kHz) the linearized solution holds until the tracking

frequency has drifted approximately 800 Hz. Since this presents a far wider frequency difference from the originally input than the typical amplitude-tracking filter (g_1 , g_2) bandwidths, these filters will begin to remove the
 5 originally tracked signal's component from the amplitude estimate.

Returning to the original equations for this perturbation and filtering out both components from the amplitude-control loops, the equations governing in this
 10 regime are:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) + A_2 \sin(\Delta\omega t + \epsilon - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\gamma) + A_2 \sin(\Delta\omega t + \epsilon - \gamma)] * h_{L2}.\end{aligned}$$

δ and γ are time-varying functions. These are also the equations governing the two phase-locked loops without any cross-coupling. At this point the loops begin to reacquire
 15 the signal as normal PLL's would since the entire assembly has lost lock on both the amplitude and the phase of either signal. There is one major difference, however, in that one of the loops' frequency estimates is biased toward the interfering signal, and one is biased away from it. The
 20 loop whose frequency is moving towards ω_2 will become less effective in filtering out that signal's power, and as a result, will be significantly slower in reacquiring the first signal. The cancellation of the first signal at this loop's input will have time to take hold, allowing the
 25 possible acquisition of the second signal.

An examination of the feedforward canceller topology presented suggests that a major source of error in the amplitude estimates will be due to the VCO signal phase estimate's not being entirely uncorrelated with the
 30 undesired signal. For example, the stronger input signal

may corrupt the weaker signal tracking VCO estimate sufficiently to affect the amplitude estimate of the weaker signal. However, even with perfect capture by the PLL's, it is possible for the amplitude estimate to be corrupted by the undesired signal. This can easily occur when interfering FM signal inputs are momentarily close in instantaneous frequency, i.e., when their respective modulating signals are close in amplitude. When the instantaneous frequency difference falls within the passband of the amplitude-tracking lowpass filter, the amplitude estimate will be corrupted by the undesired signal. In fact, since nonideal lowpass filters are used, the undesired signal will have a corrupting effect, somewhat diminished, even when the instantaneous frequency difference is slightly outside the nominal passband of these filters. This unavoidable corruption of the amplitude estimate can be viewed as an undesired signal-leakage path into the interference-canceller reference, an effect that limits performance of the canceller discussed. It is, therefore, desirable to suppress any potential cross-leakage in generating the amplitude estimate.

One possible way of limiting such cross-leakage is to derive the amplitude estimate from as clean a source of the desired signal as it possible. The canceller topology already contains such a source during steady-state operation. The input to each limiter-PLL assembly has already had the estimate of the undesired signal subtracted from it, and it is a good source from which to produce the desired-signal amplitude estimate. In computing the amplitude estimate, one could multiply the VCO-produced signal estimate not by the overall canceller input signal $v(t)$, but rather by the input to its own limiter-PLL combination. The effect of leakage can be minimized in the canceller steady state by doing so. This

canceller topology is shown in Figure 14 and is referred to as the difference-amplitude-tracking topology.

In accordance with the invention, the difference-amplitude tracking technique, where the amplitude is tracked after the interference has been received, is a novel and particularly useful subsystem. Since this is used to estimate amplitude variation, either caused by channel distortion or deliberate variation, it is separable from the FM interference canceller as a whole. The difference amplitude tracking topology can be formed as Inphase-only and Inphase-and-Quadrature, and can be applied to any interference cancelling receiver, not only FM.

In the difference-amplitude tracking topology of Fig. 14, the signal S1 plus the residual of S2 (S2) is fed back to the S2 channel. In the previously discussed feedback topology of Fig. 13, the same information is instead fed back to the S1 channel and would reduce to zero in the steady state.

To examine this variation, we formally replace the terms for the amplitude estimates $([A_1 \cos(\psi_1 - \phi_2) + A_2 \cos(\psi_2 - \phi_2)] * g_2$ and $[A_1 \cos(\psi_1 - \phi_1) + A_2 \cos(\psi_2 - \phi_1)] * g_1$ by a_2 and a_1 , respectively, in the two defining differential equations for the feedforward system. Now, from the topology defined in the previous paragraph and shown in Fig. 14, the equations for a_1 and a_2 are:

$$\begin{aligned} a_1 &= [A_1 \cos(\psi_1 - \phi_1) + A_2 \cos(\psi_2 - \phi_1) - a_2 \cos(\phi_2 - \phi_1)] * g_1 \\ a_2 &= [A_1 \cos(\psi_1 - \phi_2) + A_2 \cos(\psi_2 - \phi_2) - a_1 \cos(\phi_1 - \phi_2)] * g_2 \end{aligned}$$

Note that in the desired steady state, $\phi_1 = \psi_1$, $\phi_2 = \psi_2$, $a_1 = A_1$, and $a_2 = A_2$, so that these equations become:

$$\begin{aligned} a_1 &\approx [A_1 \cos(\psi_1 - \phi_1)] * g_1 \\ a_2 &\approx [A_2 \cos(\psi_2 - \phi_2)] * g_2 \end{aligned}$$

Again, consider the perturbation on the desired state. As before, let the input amplitudes A_1 and A_2 be constant, and let the driving input phase functions be:

$$\psi_1 = \omega_1 t \quad \psi_2 = \omega_2 t + \varepsilon.$$

- 5 Let the loop initial conditions be a perturbation of the desired steady state:

$$\phi_1 = \omega_1 t + \delta \quad \phi_2 = \omega_2 t + \gamma.$$

Then the loop differential equations are:

$$\begin{aligned} \dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) + A_2 \sin(\Delta\omega t + \varepsilon - \delta) - a_2 \sin(\Delta\omega t + \gamma - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\Delta\omega t - \gamma) + A_2 \sin(\varepsilon - \gamma) - a_1 \sin(-\Delta\omega t - \gamma + \delta)] * h_{L2}, \end{aligned}$$

- 10 with a_1 and a_2 being:

$$\begin{aligned} a_1 &= [A_1 \cos(-\delta) + A_2 \cos(\Delta\omega t - \delta) - a_2 \cos(\Delta\omega t + \gamma - \delta)] * g_1 \\ a_2 &= [A_1 \cos(-\Delta\omega t - \gamma) + A_2 \cos(\gamma) - a_1 \cos(-\Delta\omega t + \delta - \gamma)] * g_2. \end{aligned}$$

With some trigonometric substitution to combine terms of the same frequency, the amplitude terms α_1 and α_2 become:

$$\begin{aligned} \alpha_1 &= [A_1 \cos(-\delta) + (A_2 - a_2 \cos(\gamma) + a_2 \sin(\gamma)) \cos(\Delta\omega t - \delta)] * g_1 \\ \alpha_2 &= [(A_1 - a_1 \cos(\delta) + a_1 \sin(\delta)) \cos(-\Delta\omega t - \gamma) + A_2 \cos(\gamma)] * g_2. \end{aligned}$$

- 15 As seen above, any leakage of the undesired signal through the amplitude lowpass filter will be significantly attenuated for small δ and γ . After filtering out the amplitude terms with frequency $\Delta\omega$, the perturbation is reduced to the same system referred to earlier
- 20 after the amplitude filtering:

$$\begin{aligned} \dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) + A_2 \sin(\Delta\omega t + \varepsilon - \delta) - A_2 \cos(\varepsilon - \gamma) \sin(\Delta\omega t + \gamma - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\Delta\omega t - \gamma) + A_2 \sin(\varepsilon - \gamma) - A_1 \cos(-\delta) \sin(-\Delta\omega t - \gamma + \delta)] * h_{L2}. \end{aligned}$$

- The advantage of the difference-amplitude-tracking topology over the regular feedforward topology is as expected, in there is extra attenuation of the undesired
- 25 signal in the amplitude estimate even when the instantaneous frequencies are close.

Consider now what happens to the difference-amplitude-tracking topology when the two loops start tracking the same signal. As before, let the driving phase be:

$$5 \quad \psi_1 = \omega_1 t \quad \psi_2 = \omega_2 t + \varepsilon,$$

and let the initial conditions be:

$$\phi_1 = \omega_1 t + \delta \quad \phi_2 = \omega_1 t + \gamma, \quad \gamma \neq \delta.$$

The equations for the amplitude estimates are then:

$$a_1 = [A_1 \cos(-\delta) + A_2 \cos(\Delta\omega + \varepsilon - \delta) - a_2 \cos(\delta - \gamma)] * g_1$$

$$a_2 = [A_1 \cos(-\gamma) + A_2 \cos(\Delta\omega + \varepsilon - \gamma) - a_1 \cos(\delta - \gamma)] * g_2$$

10 Filtering the $\Delta\omega$ frequency terms and letting δ and γ be small, these equations become:

$$a_1 \approx A_1 - a_2$$

$$a_2 \approx A_1 - a_1.$$

Since the two amplitude equations above are linearly dependent, the two amplitude estimates are underdetermined and, hence, they can vary widely in relation to the signal amplitude, A_1 , while still satisfying the defining equation. When no interfering signal is present, the amplitude estimates are free to wander, possibly producing unstable behavior. In a practical system this effect would be dealt with by disabling the canceller when no significant interference is present, as indicated by the residual power level after the desired signal has been removed.

25 The free wandering of the amplitude estimates in the undesired equilibrium indicates that the dynamics of the difference-amplitude canceller of Fig. 14 may be different from those of the previously examined feedforward canceller of Fig. 12. This difference can be easily checked in the differential equations by considering a_1 and $a_2 = A_1 - a_1$ fixed, as is justified by the much longer time

constants of the amplitude-estimation filters in comparison with the PLL's. The defining differential equations are then:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) + A_2 \sin(\Delta\omega t + \epsilon - \delta) - a_2 \sin(\gamma - \delta)] * h_{L1} \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\gamma) + A_2 \sin(\Delta\omega t + \epsilon - \gamma) - a_1 \sin(-\gamma + \delta)] * h_{L2}.\end{aligned}$$

- 5 Substituting $a_2 = A_1 - a_1$ and filtering out the $\Delta\omega$ frequency terms gives:

$$\begin{aligned}\dot{\phi}_1 &= \alpha_1 [A_1 \sin(-\delta) - (A_1 - a_1) \sin(\gamma - \delta)] \\ \dot{\phi}_2 &= \alpha_2 [A_1 \sin(-\gamma) + a_1 \sin(\gamma - \delta)],\end{aligned}$$

which, for small δ, γ linearizes to:

$$\begin{aligned}\dot{\phi}_1 &= -\alpha_1 [a_2 \gamma + a_1 \delta] \\ \dot{\phi}_2 &= -\alpha_2 [a_2 \gamma + a_1 \delta].\end{aligned}$$

By substituting for δ and $\gamma, \psi_1 - \phi_1$ and $\psi_1 - \phi_2$, respectively, we obtain:

$$\begin{aligned}\dot{\phi}_1 &= -\alpha_1 A_1 \left[\psi_1 - \frac{1}{A_1} (a_2 \phi_2 + a_1 \phi_1) \right] \\ \dot{\phi}_2 &= -\alpha_2 A_1 \left[\psi_1 - \frac{1}{A_1} (a_2 \phi_2 + a_1 \phi_1) \right].\end{aligned}$$

- 10 The behavior of this cross-coupled system is clearly dependent on the amplitude-estimate values. The equations above verify the intuitive solution that the PLL with an amplitude estimate close to A_1 will be properly corrected, while the PLL farther away will be cross-coupled in its
15 phase correction in a manner similar to the simpler feedforward system described earlier. It is also important to note that the behavior of the difference-amplitude system around the undesired equilibrium does serve to demonstrate not only that its dynamics tend to be more
20 complicated than the simpler feedforward canceller, but also that it tends to be more sensitive to initial

conditions. The initial conditions of the amplitude estimates greatly affect the perturbation dynamics.

From the foregoing, it will be apparent that Cross-coupled phase-locked loops (CCPLL)-based FM interference-cancelling receivers are suitable as an interference canceller for use in multiaccess communications. Based on an understanding of the operational principles behind the CCPLL interference canceller and on optimal estimation theory, new interference-cancelling systems are provided in accordance with the invention. Some of the new systems are relatively simple and provide only one amplitude-tracking loop. Others have improved steady-state operation, obtained by forming the estimate of a signal's amplitude from an interference-cancelled reference. For application to a given channel, the parameters of these receivers, i.e., the loop bandwidth, the loop filter transfer function, and the amplitude-estimation filters, can be optimized by a known technique, i.e., matched filtering. However, even without optimization, such interference cancellers demonstrate that reception of both of two co-channel FM interferers is possible with such cancellers.

Improvements to the basic CCPLL system based on the optimal MMSE amplitude estimator have been presented. A feedforward amplitude-estimation procedure results from direct solution of the MMSE amplitude equations. This simplifies the resulting system dynamics by reducing the number of defining differential equations from six to a maximum of four for both inphase and quadrature tracking. The tracking of the amplitude as a complex value is shown to provide little cancellation gain when the PLL generally tracks close in phase. The importance of the quadrature component thus depends on the reception environment and the PLL tracking parameters. The tracking loop for the

quadrature-amplitude component not only increases the hardware requirements but also increases the amplitude-tracker noise bandwidth. In some cases, the additional degree of freedom due to complex amplitude tracking increases weak-signal acquisition time. Consequently, forms of the feed-forward canceller that tracked only the inphase component of the amplitude were developed, further reducing the number of defining differential equations from four to two.

The reduction in differential equations made it possible to show that the system has a stable equilibrium at the desired steady-state of tracking both signals. Furthermore, it was shown that the equilibrium of both loops tracking the strong signal is unstable. The perturbation analysis revealed that the bandwidth of the amplitude-tracking loops, since it is generally less than the bandwidth of the PLL loop filter, was important relative to the instantaneous frequency difference of the two signals. The strong signal amplitude will corrupt the weak-signal amplitude estimate if a difference frequency term can pass the amplitude filters.

To minimize this source of corruption, a variation of the feedforward system was developed which develops the amplitude estimates from the input signals to the PLL-limiter assemblies, allowing the interference cancellation to reinforce the desired steady-state. This variation, the difference-amplitude-tracking topology of Fig. 14, possesses the same stable equilibrium as the inphase-tracking feedforward system, but with greater immunity to periods when the instantaneous frequency difference is within the bandpass of the amplitude-estimation filter. However, the difference-amplitude-tracking topology has more complicated dynamics than the simpler feedforward cancellers. Hence, successful separation and demodulation

of each of two co-channel interfering FM modulated signals is achieved by all four feedforward interference-canceller topologies (inphase and quadrature amplitude-tracking, inphase-only amplitude tracking, inphase and quadrature difference-amplitude-tracking, and in-phase-only difference-amplitude-tracking). Such interference cancellers may be specifically applied to the cellular telephone channel. The cellular telephone channel is an important example of a multiuser channel since it uses the natural geographic attenuation of signals to allow a limited number of frequency slots to be reused in cells separated in the service area. Rules for determining when a frequency may be reused are determined by tolerable interference levels. Hence by allowing an increase in the level of interference that can be tolerated, frequencies may be reused more often within a given fixed geographical area. This would increase the cellular system's spectral efficiency, allowing more communications to take place over the existing limited spectrum allocation. The multipath fading property of this channel is a major factor in interference-related call degradation. The problem to be addressed with interference cancellers in accordance with the present invention is defined, along with estimates of the potential for frequency reuse gains.

The frequency-allocation of the cellular telephone channel in the United States, like that of other radio communications channels, is the product of much administrative regulation, and as such, can be administratively defined. Since cellular voice transmission is full duplex as opposed to push-to-talk, the cellular channel is naturally divided into mobile-to-base and base-to-mobile frequency sets. Because of the requirement for having competing cellular systems available in each metropolitan area, the available bandwidth in each set is then halved

between two competing carriers. The individual channel spacing in the United States is set at 30 kHz, and voice transmission is FM with pre-emphasis and a maximum peak frequency of deviation of ± 12 kHz.

5 Propagation in the individual channels has been modeled as an inverse fourth-order law with Rayleigh fading and large scale average lognormal shadowing. Since the shadowing is highly geographically dependent and occurs in an average (long-term) signal level sense, it will not be
10 dealt with further here. While shadowing may provide some relief from interfering signals, the Rayleigh fading devastates the on-average capture effect. Fades of from 10 to 40 dB with durations on the order of one to ten milliseconds occur at rates from 1 to 100 per second. However,
15 it should be remembered that capture can be observed on the time scale of the FM channel bandwidth, or at least the tracking loop bandwidth for PLL's, a minimum of ± 3 kHz. As a result, demodulator outputs can and do actually show signal capture during the relatively long time between
20 fades.

 Use of the cellular telephone system in the United States has grown at an unexpectedly rapid rate. Judging from continuing sales of new units, the demand is not yet saturated, even though the costs of use and
25 subscriber equipment are still relatively high, with even the least expensive mobile phone units priced at over three hundred dollars. The technical challenge of serving such a large traffic volume over a limited frequency-allocation is the main one facing the growing cellular telephony
30 industry. To help towards a solution, new multiaccess modulation and coding schemes are being generated elsewhere. For them, implementation is still a long time off, partly because of the administrative process of obtaining

a frequency-allocation, and partly because of the substantial subscriber investment in the existing cellular system.

The cellular channel has 24 kHz allocated for voice transmission. Voice is syllabically companded (companded with a time constant on the order of 20 milliseconds), pre-emphasized and frequency-modulated. The peak frequency deviation is 12 kHz, maintained by amplitude clipping if necessary. The rms frequency deviation is 2 kHz, with f_{rms}^2 log normally distributed with standard deviation of about 5 dB.

The current FM cellular telephone receiver is a simple limiter-discriminator demodulator. While the limiter-discriminator based receiver is inexpensive and simple to produce, advances in analog integrated circuits continually make more complex receiver designs practical. Entire loop-based FM receivers integrated on a chip can be purchased by consumers for under five dollars. At this cost, the cellular telephone receiver, the main element responsible for call quality in a cellular phone costing several hundred dollars or more, can certainly afford to have some money spent on improved performance, especially to increase the traffic capacity.

Because of this rapidly varying environment, the FM signal must be strong enough to overcome the more frequent fading, and must be fairly robust in the presence of deeper fades. For data communications, the fading situation is a problem that can be solved by appropriate burst-error-correction codes.

Discarding amplitude is an information-lossy process. At worst, an amplitude track can at least provide

a figure of confidence for the frequency estimates taken at the same time to be used as side information in further filtering processes. At best, the amplitude tracks can be used to feed a received-signal environment model in an effort to separate the signal of interest. As an example, consider the case of an angle-modulated signal in additive Gaussian noise. Clearly, the angle information will be corrupted the most when the received amplitude is near its mean, and the received information will be the least corrupted when the amplitude is nearest its maximum and minimum, except in the case of origin encirclements in the phase plane. These can be more easily detected with amplitude information, since a time track of the amplitude will decrease at the same time as the instantaneous frequency is increasing. By using the amplitude information inherent in the received signal, a joint estimation (amplitude and phase) approach has been shown to produce reception with 5 dB better output signal-to-noise ratio than more conventional PLL-based receivers.

For high-traffic cellular systems, the interference power is usually significantly greater than the background noise power. Frequency allocation schemes must meet the requirement that the minimum mean signal-to-interference ratio be +18 dB, in the current scheme. At first glance this seems more than a little conservative, considering the FM capture effect. However, the basis for the SIR requirement is derived not from considering a signal of constant power, but from the maximum tolerable losses under multipath fading. For speech systems, tolerable variations in maximum acceptable probability of loss that is due to multipath fades vary from .1 as the highest acceptable value for speech down to 0.01 for a high-quality speech system. The problem for the interference canceller is then to reduce the level of interference

sufficiently so that track of the desired signal can be maintained through a fade without loss.

We apply the interference canceller in such a way as to make use of the difference in instantaneous frequency between uncorrelated, co-channel interfering FM signals. The application of interference cancellers, in accordance with the invention, is directed at receiving both the stronger and the weaker of two signals.

The interference canceller must be able to track signals with time-varying amplitudes. In an effort to track the Rayleigh fading envelopes of the signals with minimal delay, the amplitude-estimation filters are chosen to be first-order Butterworth filters with cutoff frequencies larger than the bandwidth of the amplitude modulation. In order to force a desired PLL to be the first to acquire a signal, the amplitude-estimation filter is made wider on one loop than on the other. The amplitude loop bandwidths are 400 Hz for the weak-signal estimator and 200 Hz for the strong-signal estimator.

The PLL's must be able to separate signals with the same general modulation characteristics, but with widely varying instantaneous power levels. Without the Rayleigh fading, the loop parameters would be proportional to the power ratio. The fading channel makes determination of the exact power ratio to be used difficult, but it is reasonable to design the loops for the mean power ratio. In a practical application the mean power ratio could be estimated or controlled in order to get the best response, but a fixed value can be chosen. Since the ultimate goal is to separate interferers with close mean power levels for frequency reuse in the same cell, the loops are designed for a mean power ratio close to one. The PLL's are

designed using the rules for 3 kHz modulation bandwidth at 12 kHz peak frequency deviation and a weak-to-strong amplitude ratio of 0.85 (1.4 dB SIR, strong to weak). Experimental results have been reported indicating that
5 performance is improved by matching the power level to the "weak" signal loop parameters. As a result, performance may suffer slightly at signal-to-interference ratios away from 0 dB.

One advantage of designing around a power ratio
10 close to one is that the bandwidths of both loops are kept small. This permits cleaner phase tracking of both signals in any case, since both PLL's now see a smaller instantaneous bandwidth. Both the fading, which shifts the signal phase as well as its amplitude, and the small loop band-
15 widths required for operation at a mean power ratio close to one predictably make the instantaneous loop phase errors larger. The error will degrade the performance of the inphase-only amplitude-tracking interference cancellers relative to the inphase and quadrature tracking forms,
20 which can correct for loop phase error in the amplitude estimates.

In addition to obtaining better performance from the inphase and quadrature amplitude-tracking forms, the difference-amplitude-tracking configuration performs
25 better than the feedforward configurations. This is because the fading channel requires amplitude tracking at a bandwidth close to the lower edge of the modulation bandpass. More interfering signal power can come through the amplitude filters, causing corruption of the amplitude
30 estimate. In this regard, the main advantage of the difference-amplitude configurations is the reduced corruption of amplitude estimates by the interference. In a system where the amplitude-tracking filters are wider,

this factor should become important in determining performance.

In the certified copy of U.S. Patent Application Serial No. 07/910,829 (from which priority is claimed) can be found an Appendix A, which is hereby incorporated herein by reference, which is a document setting forth in greater detail applications of frequency modulation interference cancellers to multiaccess communication systems prepared by and reflecting efforts by the inventor in the development of the present invention. While Appendix A is not considered necessary for a proper understanding to enable one of ordinary skill in the art to make and practice the invention, it is provided for added convenience and insight.

Referring now more particularly to Figs. 15 through 22 of the drawings, there is shown a presently preferred embodiment of the invention, an interference cancelling receiver system for analog FM transmission similar to that used for cellular telephony in the United States. By way of example, and for clarity, the preferred embodiment is shown as a receiver for two interfering signals. However, the number of interfering signals is a parameter in the design, i.e., the receiver design can be directly expanded to cover more than two interfering signals.

The preferred embodiment of the invention is a digital hardware implementation of the inphase and quadrature difference amplitude tracking receiver for two interfering FM signals and, as such, is similar to the embodiment of Fig. 14 previously described. The system operates on a sample clock frequency of 120 kHz, with inphase and quadrature sampling. The fundamental word-length in the digital system is 16 bits.

Referring now more particularly to Fig. 15 of the drawings, the interfering signals and noise enter the system through a conventional antenna (not shown), over line 100, and are conventionally downconverted and filtered to a convenient, standard intermediate frequency (IF) by subsystem 101. The IF is then converted to baseband inphase and quadrature (90 degree phase-shifted) (I&Q) 12-bit digital samples, oversampled at a 120 kHz sample rate by subsystem 102 in Fig. 15.

As shown in Fig. 16, the I&Q analog-to-digital (A/D) conversion subsystem 102 may be of conventional design, formed by splitting the IF on line 110 into two streams through a power divider 111 and mixing one stream at a mixer 112 with $\cos(2\pi f_{IF}t)$, a local oscillator at zero degrees phase, and the other stream at a mixer 113 with $-\sin(2\pi f_{IF}t)$, a local oscillator of 90 degrees phase, where f_{IF} is the IF frequency. These two signals are then low-pass filtered by a pair of filters 114, 115, respectively, to a 30 kHz bandwidth to produce the desired baseband I&Q signal streams. The baseband I&Q signals are then sampled and quantized by 12 bit A/D converters 116, 117, respectively. The 12 bit digital data streams then enter the interference cancelling FM demodulator subsystem 103, over lines 118, 119, as shown in Fig. 15.

The interference cancelling FM demodulator subsystem 103 is shown in great detail in Fig. 17. Digital I&Q samples are input to the FM demodulator subsystem 103, and estimates of the I&Q samples of each of the two interfering signals are subtracted from the input at 120a and 120b, respectively. The resulting difference signals have reduced interference levels, with one signal stronger on the "a" I&Q paths and the other signal stronger on the

"b" I&Q paths. Each path pair is input to a phase-locked loop (PLL) of conventional design, comprised of a phase detector 121, loop gain and filter 122, and a numerically controlled oscillator (NCO) 123. The PLL demodulates the FM signal (at 3a or 3b) and produces an amplitude-normalized estimate (at 4a or 4b) of the signal it is demodulating. This estimate, output from the PLL's NCO, enters an amplitude estimator 124 (at 5a or 5b) which constructs the interference estimate (at 6a or 6b) to be subtracted from the signal input of the other signal's demodulator at 120b or 120a. The I&Q samples of the difference signal first enter the phase detector 121 which produces a signal equal to the sine of the phase difference (at 2a or 2b) between the difference signal (at 1a or 1b) and the NCO's current signal estimate (at 4a or 4b).

Referring now to Fig. 18 of the drawings, the phase detector 122 may be of conventional design. It first produces an estimate of the instantaneous signal power by squaring each component (I&Q) at (1) and forming the sum I^2+Q^2 . The reciprocal square root of this value ($1/\sqrt{I^2+Q^2}$) is then looked up in a table at (2), and multiplied by the I&Q components to normalize the instantaneous magnitude at (3), producing the signals I1 and Q1. The NCO outputs, which are produced by the NCO already in normalized form, are input as Inphase Input #2 (I2) and Quadrature Input #2 (Q2). I1 and Q2, and Q1 and I2 are multiplied to form the products $I1*Q2$ and $Q1*I2$ at (5). These results are then differenced at (6) to produce an output $Q1*I2 - I1*Q2$. Note that if we write the difference signal as $A*\cos(X1)$ (Inphase) and $A*\sin(X1)$ (Quadrature), and the NCO as $\cos(X2)$ and $\sin(Y2)$, then $(1/\sqrt{I^2+Q^2}) = 1/|A|$, $I1 = \cos(X1)$ and $Q1 = \sin(X1)$. The output is then $\sin(X1)*\cos(X2) - \cos(X1)*\sin(X2) = \sin(X1-X2)$ which is approximately $X1-X2$

when this value, the phase difference, is small. It is this value which is output from the phase detector 221 (at 2a or 2b) in Fig. 17.

5 The phase difference output (at 2a or 2b) in Fig. 17 is input to the loop filter and gain 222, shown detailed in Fig. 19, at (1). Loop gain, G , is applied by a multiplier 228 whose output is sent to both an integrator 229 at (2) with gain K , and a summer 232 to produce an integral-plus-proportional-type output at 10 (3). This output represents the demodulated FM signal and it is both sent back to the NCO 223 (Fig. 17) to increment the phase value in the NCO, and, as described earlier, out of the FM demodulator (at 3a or 3b in Fig. 17) as the demodulated output.

15 The NCO 223 accumulates the phase increment values input to it, with the resulting sum being an estimate of the signal phase at a given instant in time. It outputs I&Q values representing the cosine and sine of the accumulated phase, respectively. The I&Q output 20 pair (at 4a or 4b in Fig. 17) is used in the phase detector 221 of the PLL, as previously described.

In the interference-cancelling demodulator 103, the I&Q outputs from the NCO 223 are also input to an amplitude estimator 224 (at 5a or 5b in Fig. 17) 25 where estimates (at 6a or 6b in Fig. 17) of the I&Q input samples of each signal are produced to be subtracted from the opposing PLL's input (at 1b or 1a in Fig. 17). In the preferred embodiment, the Inphase and Quadrature Difference Amplitude Tracking form, this is 30 accomplished by complex correlation of the NCO outputs (at 4a or 4b in Fig. 17) with the difference, interference-cancelled signal (at 1a or 1b in Fig. 17).

The amplitude estimator 224 is shown in greater detail in Fig. 20 of the drawings. The sample estimator can be thought of in two stages. The first stage is the production of I&Q narrowband amplitude estimates by complex correlation. The second stage is the complex modulation of I&Q outputs from the NCO 223 (Fig. 17) by the amplitude estimates to produce the I&Q sample values.

The complex correlation is performed as a complex conjugate multiplication followed by lowpass filtering by lowpass filters 230. The I&Q difference samples (at 1a or 1b in Fig. 17) are input as 1I and 1Q in Fig. 20, and the I&Q NCO outputs (at 4a or 4b in Fig. 17) are input as 2I and 2Q in Fig. 20. The four cross products formed in the first multiplier section (1) in Fig. 20 are: $1I \cdot 2I$, $1Q \cdot 2Q$, $1Q \cdot 2I$, and $1I \cdot 2Q$. These are then added to form the I&Q samples of the complex conjugate product at (2), $3I = 1I \cdot 2I + 1Q \cdot 2Q$, and $3Q = 1Q \cdot 2I - 1I \cdot 2Q$. 3I and 3Q are then each lowpass filtered to form the I&Q amplitude estimates 4I and 4Q at (3).

The I&Q amplitude estimates 4I and 4Q are then complex multiplied with the NCO inputs 2I and 2Q to form the signal estimates. This is accomplished by forming the cross-products $4I \cdot 2I$, $4Q \cdot 2Q$, $4I \cdot 2Q$, and $4Q \cdot 2I$ at the second multiplier section (4), and forming the I&Q sums $5I = 4I \cdot 2I$, and $5Q = 4I \cdot 2Q + 4Q \cdot 2I$, at (5). These sums, which are the estimates of the separated, received signal I&Q samples, are then subtracted from the opposing signal PLL's input in the interference-cancelling demodulator 203 (at 1b or 1a in Fig. 17).

The lowpass filter 230 used in Fig. 20 follows a basic first-order recursive design. The coefficients are for a first-order digital Butterworth filter. The same basic filter design, with different bandwidths

(controlled by the filter coefficients) is used in all the digital lowpass filters in this embodiment of the invention.

Figure 21 of the drawings illustrated the lowpass filter 230 in greater detail. The signal samples to be filtered are input at (1). The input sample, and a one-sample delayed version of the input, are each multiplied by their respective coefficients, CI1 and CI2 respectively, and are summed at (2). This sum is then added to the product of the previous output and its coefficient, CO1 (3), producing the new output at (4). This new output is both output from the filter 230 and placed in the one-sample delay register to be used in the next sample's computation.

Note, that, for the lowpass filter 230 shown in more detail in Fig. 21, if any coefficient (CI1, CI2, or CO1) is 1.0, the multiply associated with that coefficient is removed from the filter. If the coefficient is an even power of 2, the multiplier becomes a shifting operation, e.g., a multiply by 4 is a logical left shift of 2 bits. Negative powers of two, e.g., 1/4, are logical right shifts. Negative coefficients can be handled as above, with the following addition changed to a subtraction.

The demodulated outputs of the Interference-Cancelling FM Demodulator 203 are passed to the Tone Detector and Channel Selection subsystem 204 in Fig. 15. Fig. 22 illustrates the latter subsystem 204 in greater detail. Both demodulated inputs (X1 and X2) are multiplied by the I&Q outputs of an NCO (YI and YQ at (2)), producing the I&Q product pairs (YI*X1, YQ*X1), and (YI*X2), YQ*X2). Each signal is passed through a lowpass filter 230, which must be narrower than the differences of the two signal marker frequencies. The

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filtered outputs at (3) are Z1I, Z1Q, and Z2I, Z2Q. These are then each squared and each I and Q pair are summed to form the signal powers as in the phase detector, i.e., $Z1I*Z1I + Z1Q*Z1Q$, and $Z2I*Z2I + Z2Q*Z2Q$. The power values are compared by a comparator 233 at (4), resulting in the selection (5) at multiplexor 234 of the demodulated output with the strongest component of the desired marker frequency for output at (6).

The resulting selected output is then conventionally converted via a digital to analog converter 205 and lowpass filtered through a conventional analog filter 206 to the desired signal bandwidth at (5) in Fig. 15, to provide the desired demodulated and filtered output over line 207.

Circuit Parameters for the system of Figs. 15 through 22 are as follows:

Digital Word Length = 16 bits

Sampling Rate = 120 kHz, Inphase and Quadrature Samples

Amplitude Estimator Filters:

First order Butterworth filters

transformed by bilinear transform method

Estimator #1 (Inphase & Quad)

Lowpass Filter #1: Bandwidth = 400 Hz

Filter coefficients: $CI1=1.$, $CI2=1.$,

$CO1=(1.-\tan(\pi/300))/(1.+\tan(\pi/300))$

Estimator #2 (Inphase & Quad)

Lowpass Filter #2: Bandwidth = 200 Hz

Filter coefficients: $CI1=1.$, $CI2=1.$,

$CO1=(1.-\tan(\pi/600))/(1.+\tan(\pi/600))$

Phase-Locked Loop (PLL) Parameters:

Both are second order phase-locked loops

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with integral-plus-proportional-type loop filters (1+K/s type)

Loop Filter Update (both PLL's, once each clock cycle)

5 ACC = in + ACC; out = in + K×ACC;
 Loop filter integrator gains:
 Loop #1 (a) : $K_1 = 10000/120000 = 1/12$
 Loop #2 (b) : $K_2 = 13400/120000 = 67/600$

10 Note: alternative analog embodiment has
 K's multiplied by 120000

Total Loop Gain =
 (Loop Gain) × (phasedetector gain) × (NCO gain)

15 In this case the phase detector gain
 is 1 unit/rad,
 Loop Gain = G1 (PLL #1 (a)) or G2 (PLL #2 (b)).
 The values of G1 and G2 depend on the
 NCO's chosen.

20

Total Loop Gain (PLL #1, (a)) =
 $G_1 \times (\text{NCO gain}) = 20000\pi/60000 = \pi/3 = 1.04720$
 Total Loop Gain (PLL #2, (b)) =
 $G_2 \times (\text{NCO gain}) = 13400\pi/60000 = 67\pi/300 = 0.70162$

25 If the NCO input is in cycles, as is
 common, then the NCO gain is 2π
 radians/unit. Resulting in:
 $G_1 = 1/6 = 0.166667$, and
 $G_2 = 67/600 = 0.111667$

Note: alternative analog embodiment
has gains multiplied by 120000

Demodulated Output Filtering (Conventional Analog
Design):

5 Bandwidth = 4 kHz

Tone-based signal Identification

Lowpass Filter #3: Bandwidth = 200 Hz

Filter coefficients: $CI1=1.$, $CI2=1.$,
 $CO1=(1.-\tan(\pi/600))/(1.+\tan(\pi/600))$

10 Marker frequencies should be between 4 kHz
and 5 kHz, separated by at least 400 Hz.

A digital design engineer of ordinary skill and
familiar with the state of the art in digital signal
processing and digital hardware as taught in common
15 textbooks, e.g., M. Morris Mano, Digital Logic and Computer
Design, Englewood Cliffs, NJ: Prentice-Hall, 1979; P.
Horowitz and W. Hill, The Art of Electronics, New York:
Cambridge University Press, 1980; and L. R. Rabiner and B.
Gold, Theory and Application of Digital Signal Processing,
20 Englewood Cliffs, NJ: Prentice-Hall, 1975 (especially Ch.
9), will be able to readily make and practice the invention
from the aforescribed description.

It is important to point out that the informa-
tion-theoretic motivations for interference cancellation
25 are not dependent upon modulation type. While the cancel-
lers described are primarily for analog and digital FM
systems, similar cancellers could potentially be designed
for coded digital modulation schemes. The redundancy and
regularity of such schemes would make the interference
30 cancellation even easier to implement, and more effective
as well. For systems that currently mitigate interference

through coding techniques, i.e., CDMA systems, the addition of interference cancellation at the receiver input, fed by the decoded output, will allow the resulting system to tolerate still more interference in the channel. The various embodiments of the invention presented here are simply manifestations of the discovery that treating structured interferences as completely random noise without trying to exploit the discoverable structure is wasteful of the information represented by the discoverable structure.

The new and improved multiaccess communication system of the present invention is unique in that it harnesses interference to increase system capacity available for plural access by single or multiple users. The improved system provides receivers which are independent of any power hierarchy in that desired signal extraction does not depend on relative position of the desired signal in the power hierarchy. The system can detect even the weakest signal in the family of overlapping signals, and the relative power hierarchy need not be stable or preserved. Hence, the system of the present invention satisfies a long existing need in the communications industry for an accurate, reliable, easy to use, relatively simple and economical system for exploiting the frequency spectrum to provide increased system capacity available to multiple users.

It will be apparent from the foregoing that, while particular forms of the invention have been illustrated and described, various modifications can be made without departing from the spirit and scope of the invention. Accordingly, it is not intended that the invention be limited, except as by the appended claims.

CLAIMS

1. A communication system for extracting a selected signal, irrespective of relative signal strengths, from a plurality of such signals having some relationship to the selected signal, comprising:

5 first means for selecting a desired signal from the plurality of such related signals;

second means for estimating the magnitude of at least one electrical characteristic in the remaining unselected signals in said plurality of signals; and

10 third means, responsive to said second means, for substantially cancelling said remaining unselected signals to facilitate reception of said desired signal, whereby any of such signals in said plurality of signals may be designated as said desired signal.

2. A system as set forth in claim 1, wherein said desired signal is demodulated by said system for reception.

3. A system as set forth in claim 2, wherein all of said plurality of signals are demodulated by said system.

4. A system as set forth in claim 1, wherein at least a portion of said system is analog.

5. A system as set forth in claim 1, wherein at least a portion of said system is digital.

6. A system as set forth in claim 1, wherein at least a portion of said system includes a feedforward electrical tracking network.

7. A system as set forth in claim 1,
wherein at least a portion of said system
includes an inphase tracking electrical network.

8. A system as set forth in claim 1,
wherein at least a portion of said system
includes a quadrature electrical tracking network.

9. A system as set forth in claim 1,
wherein at least a portion of said system
includes both inphase and quadrature amplitude tracking.

10. A system as set forth in claim 1,
wherein at least a portion of said system
includes a difference-amplitude tracking network.

11. A system as set forth in claim 1,
wherein at least a portion of said system
includes both inphase and quadrature difference-
amplitude tracking.

12. An apparatus for use in a multiaccess
system for extracting a selected signal from a larger
family of such signals having some relationship to the
selected signal and overlapping predictable electrical
characteristics, the apparatus comprising:

first means for selecting a desired signal
from the family of such related signals;

second means for estimating the magnitude of
at least one predetermined electrical characteristic in
the remaining unselected signals in said family of
signals; and

third means, responsive to said second means,
for substantially canceling said remaining unselected
signals to facilitate reception of said desired signal,
whereby any of such signals in said family may be

designated as said desired signal to increase system capacity.

13. An apparatus as set forth in claim 12, wherein said desired signal is demodulated by said system for reception.

14. An apparatus as set forth in claims 12 or 13, wherein all of said signals are jointly demodulated by said system.

15. An apparatus as set forth in claim 12, wherein said system is analog.

16. An apparatus as set forth in claim 12, wherein said system is digital.

17. An apparatus as set forth in claim 12, wherein said system includes a feedforward electrical interference cancelling network.

18. An apparatus as set forth in claim 12, wherein said system includes an inphase amplitude-tracking feedforward interference canceller.

19. An apparatus as set forth in claim 12, wherein said system includes a difference-amplitude-tracking feedforward interference canceller.

20. An apparatus as set forth in claim 12, wherein said system includes an inphase and quadrature amplitude-tracking feedforward interference canceller.

21. A combination as set forth in claim 12,

wherein at least a portion of said system includes an inphase and quadrature difference-amplitude-tracking feedforward interference cancelling network.

22. A combination as set forth in any of claims 1-21, wherein at least a portion of said system includes a cross-coupled phase locked loop electrical network.

23. An apparatus for use in a communication system for access by multiple users to a plurality of available related signals which normally interfere with each other, the apparatus comprising:

- 5 first means for selecting a desired signal from a plurality of interfering signals;
- second means for estimating predetermined parameters in the remaining interfering signals; and
- 10 third means, responsive to said second means, for substantially cancelling the remaining interference signals to receive said desired signal, whereby any of said interfering signals may be designated as said desired signal to increase system capacity.

24. An apparatus as set forth in claim 23, wherein said system includes a cross-coupled phase locked loop electrical interference cancelling network.

25. A communication method for providing access to a single signal from within a plurality of mixed, normally interfering signals, comprising the steps of:

- 5 selecting a desired signal from a plurality of signals;
- estimating predetermined parameters in the remaining undesired signals; and
- 10 cancelling said undesired signals to receive said desired signal, whereby any of such signals may be

designated as said desired signal to increase spectral efficiency.

26. A method of increasing capacity in a communication system, comprising the steps of:

selecting a desired signal from a plurality of such signals providing interference;

5 demodulating said interference;

estimating said interference; and

utilizing the estimation of said interference to cancel said interference and thereby isolate said desired signal, whereby any of said plurality of signals
10 may be designated as said desired signal.

27. A method for extracting a selected signal from a larger family of such signals having some relationship to the selected signal and overlapping predictable electrical characteristics, comprising the
5 steps of:

selecting a desired signal from the family of such related signals;

estimating the magnitude of at least one predetermined electrical characteristic in the remaining
10 unselected signals in said family of signals; and

cancelling, in response to said estimation, said remaining unselected signals to facilitate reception of said desired signal, whereby any of such signals in said family may be designated as said desired
15 signal to increase capacity.

28. A method as set forth in claim 27, wherein said desired signal is demodulated for reception.

29. A method as set forth in claim 27, wherein all of said signals are jointly demodulated.

30. A method as set forth in claim 27,
wherein said estimating and cancelling steps
include inphase quadrature difference-amplitude-tracking
to cancel interference.

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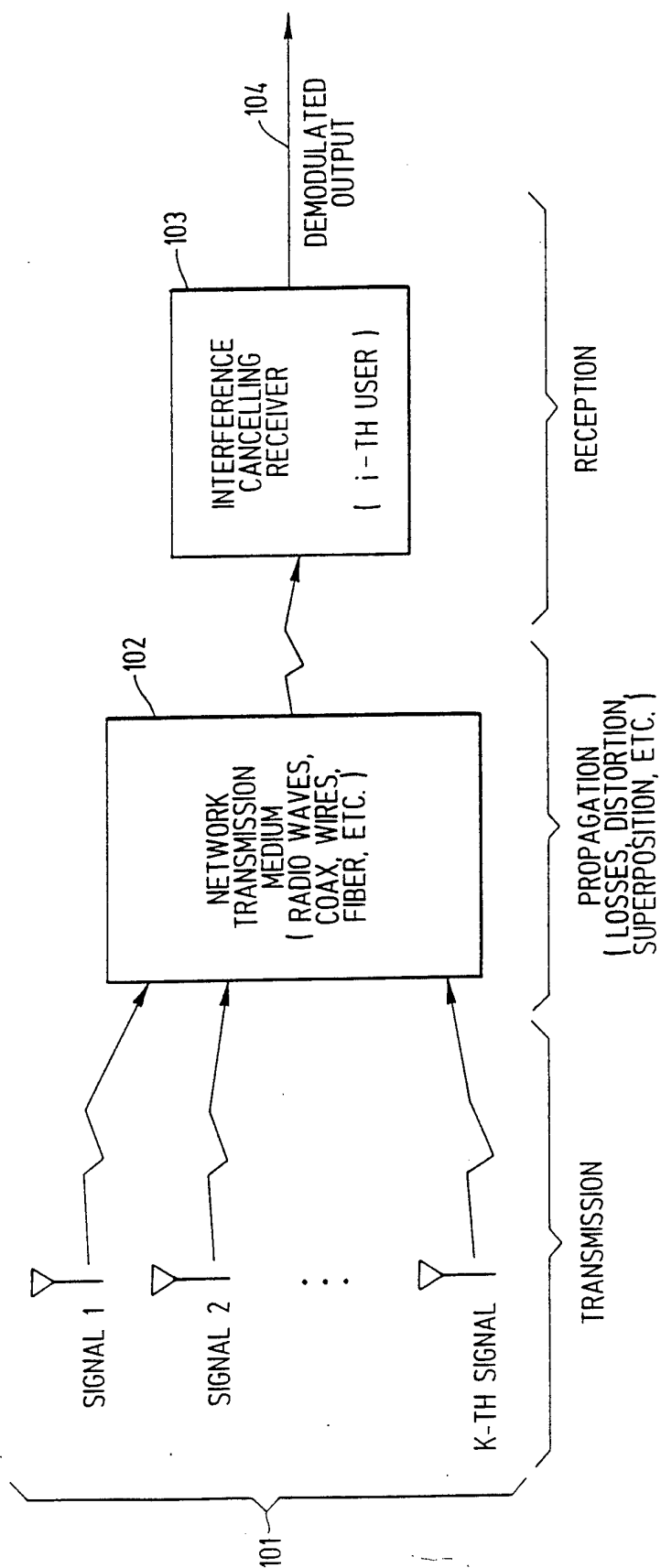


FIG. 1

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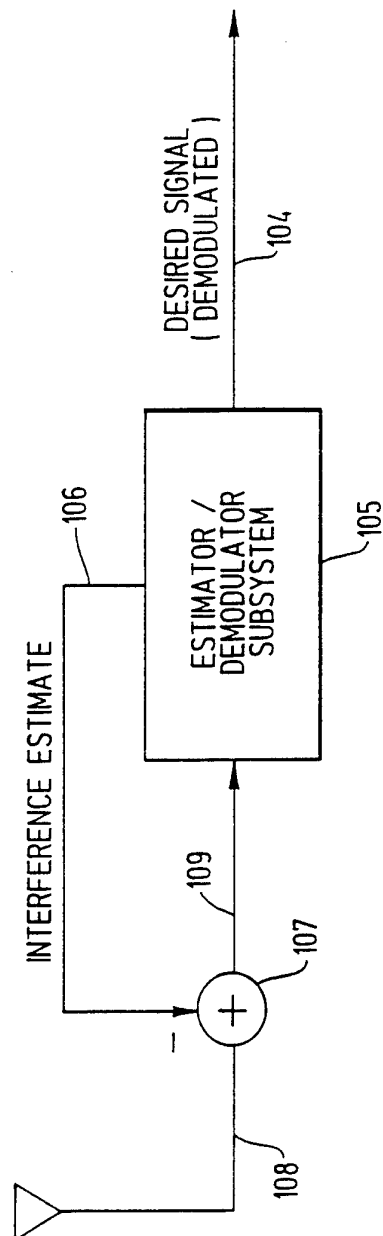


FIG. 2

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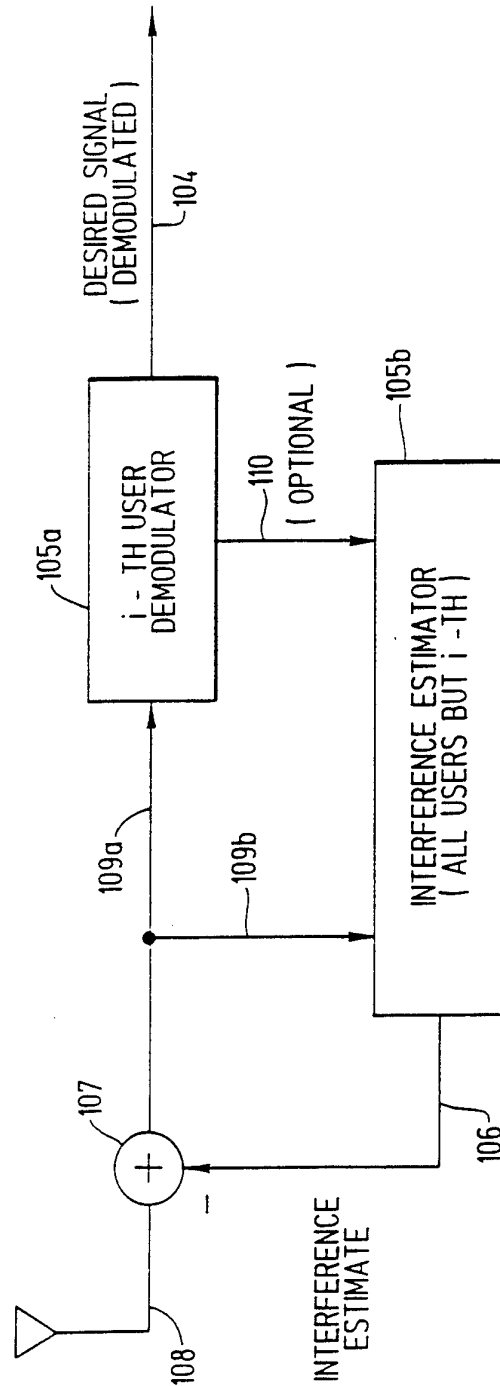


FIG. 3

4/22

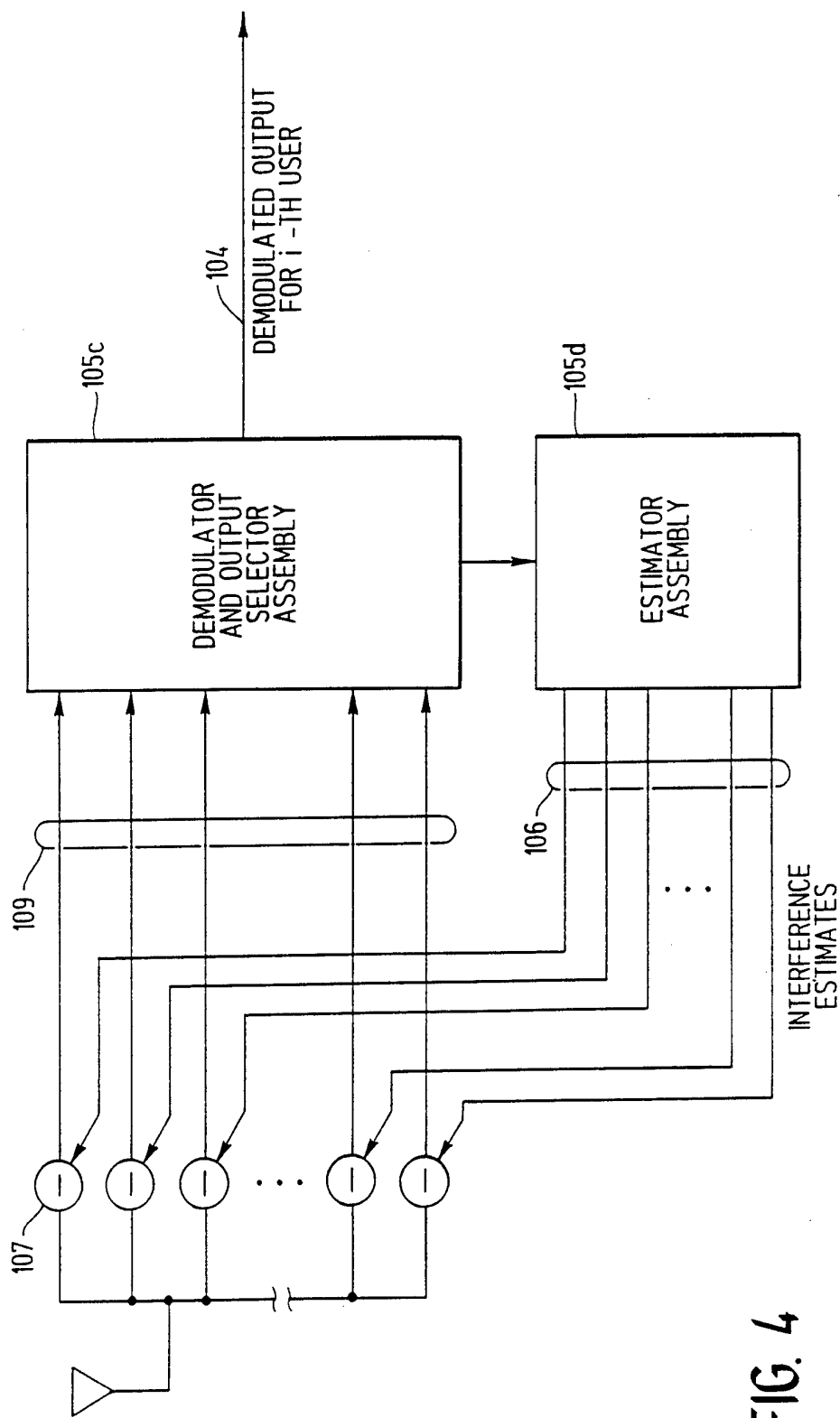


FIG. 4

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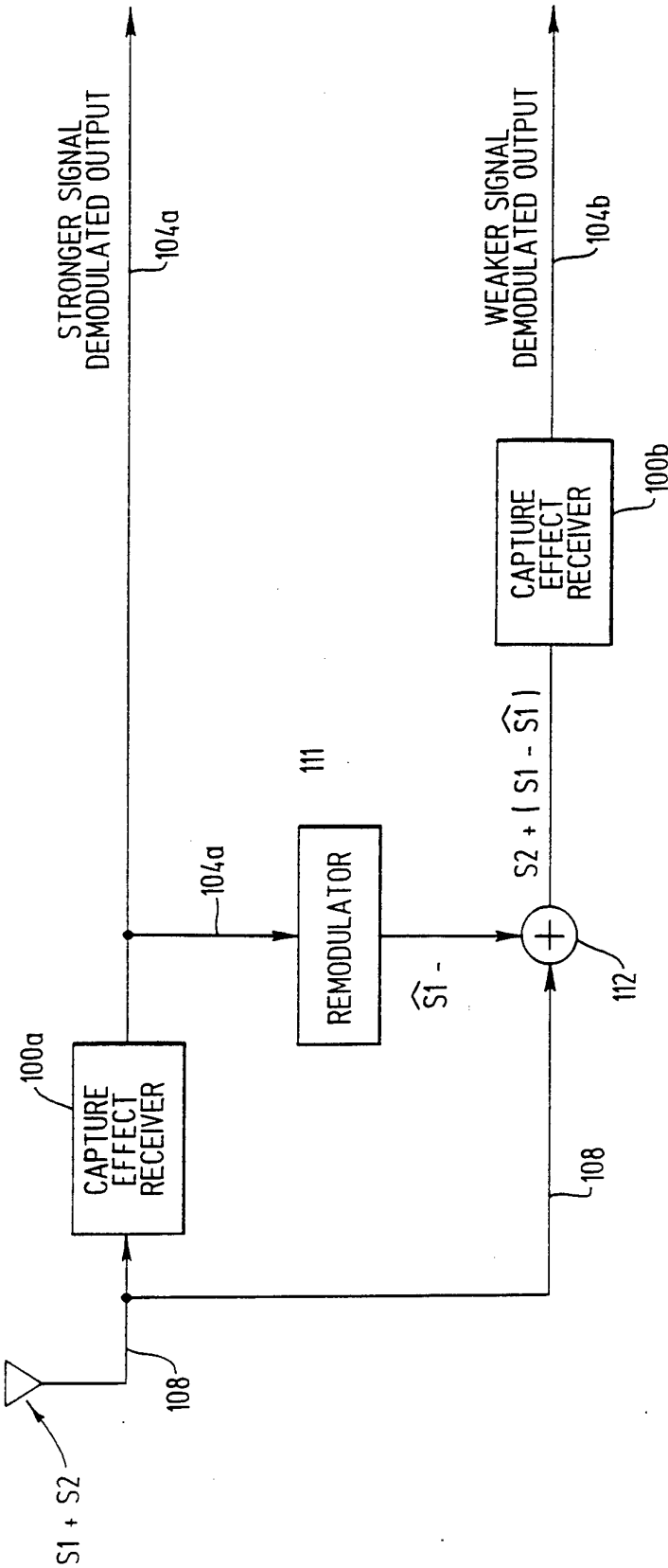


FIG. 5

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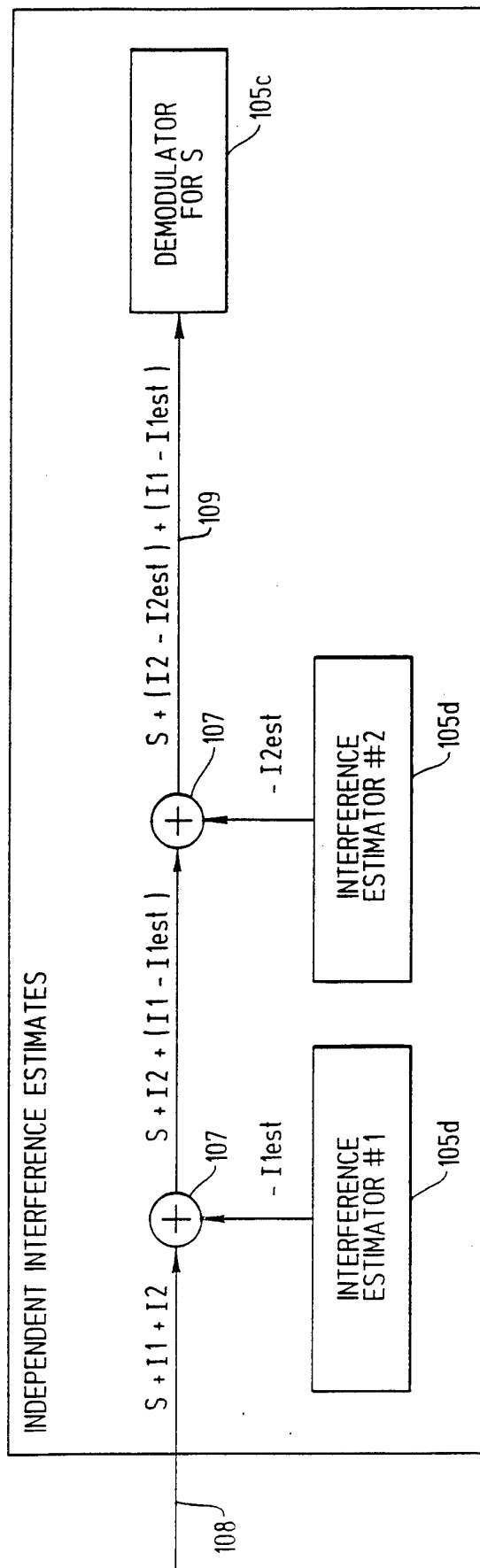


FIG. 6

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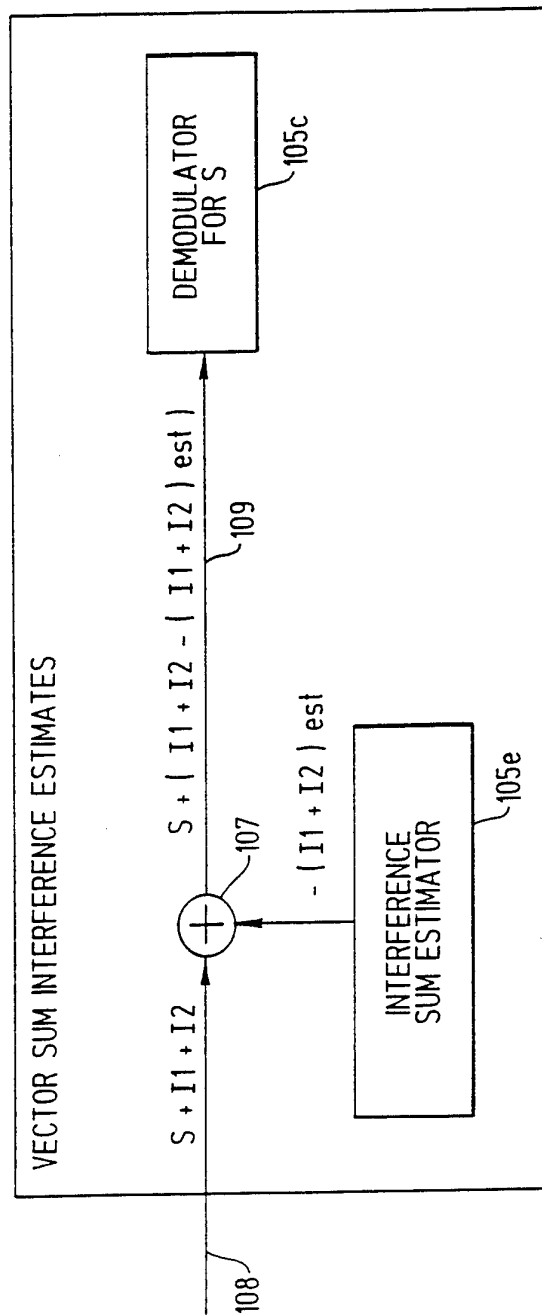


FIG. 7

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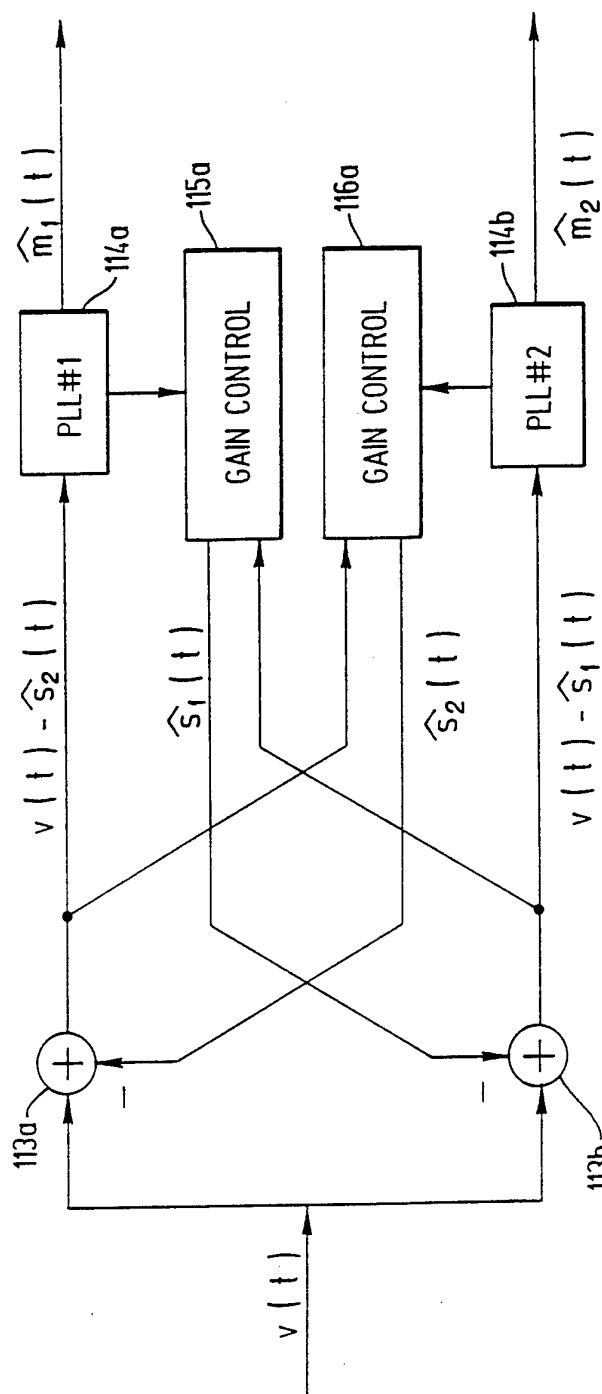


FIG. 8

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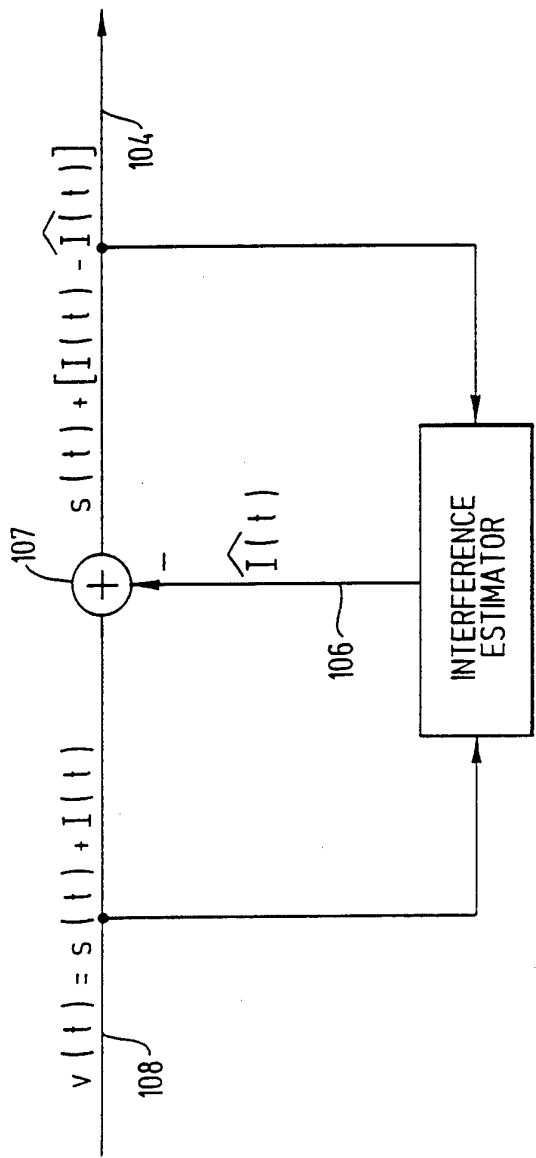


FIG. 9

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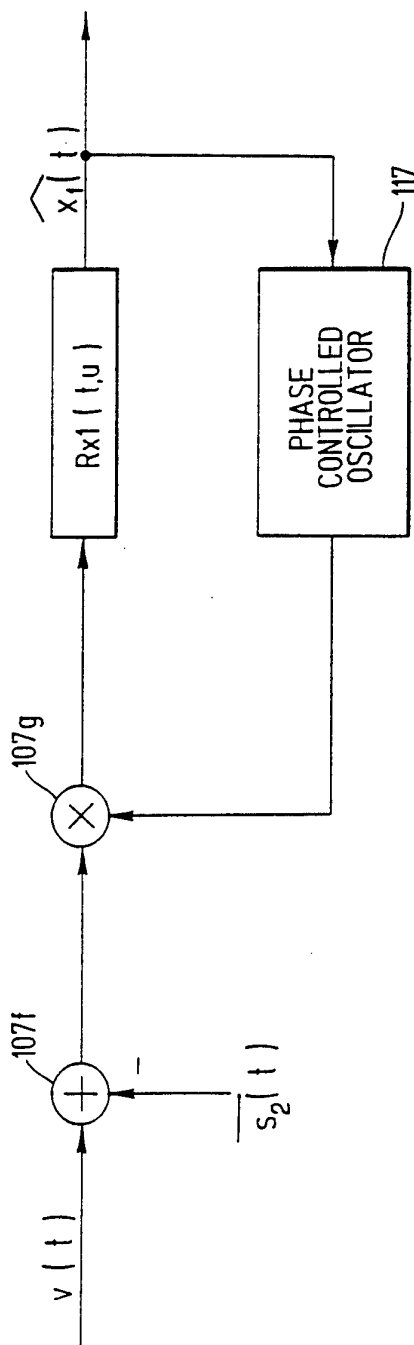


FIG. 10

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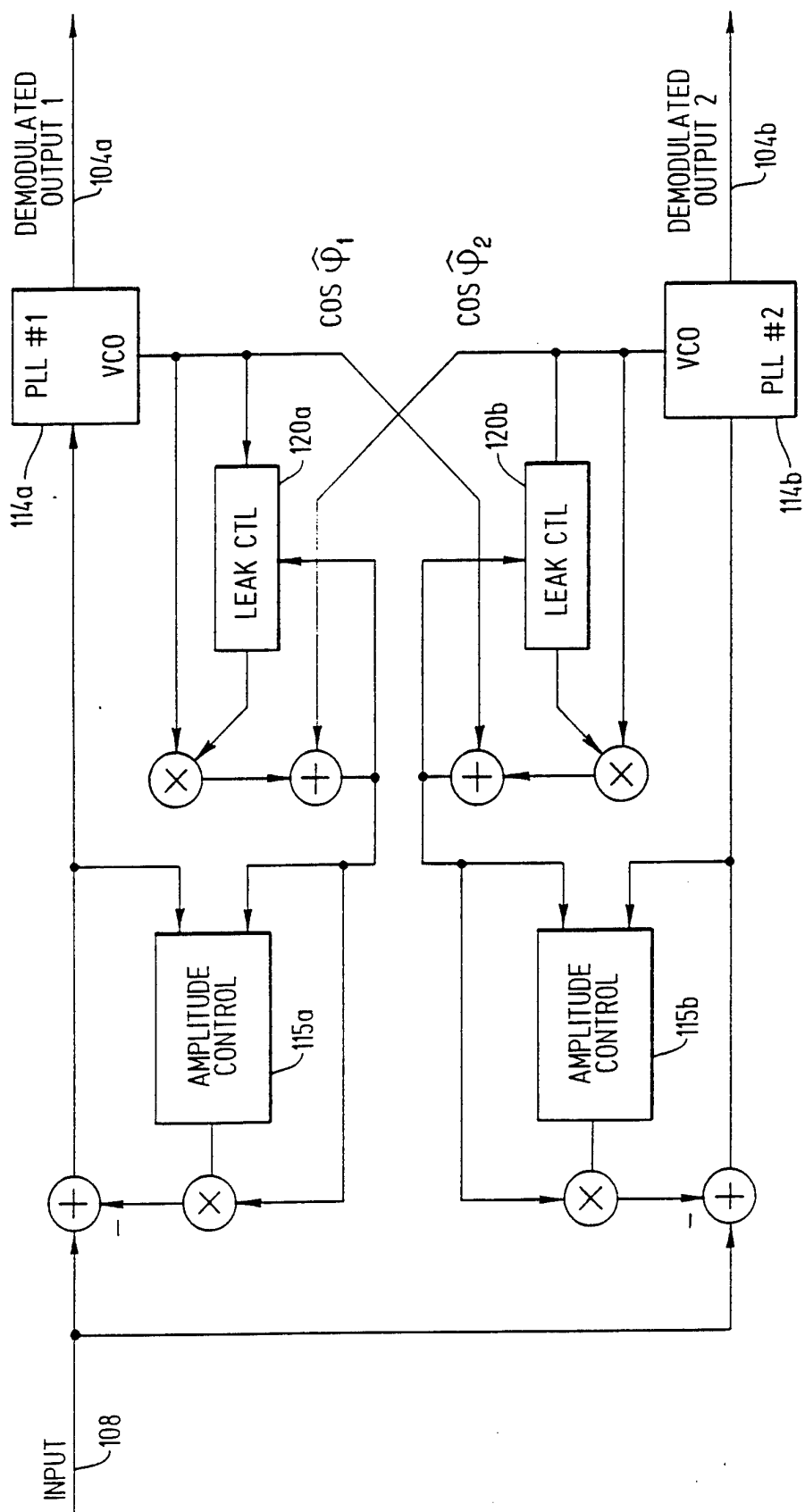


FIG. 11

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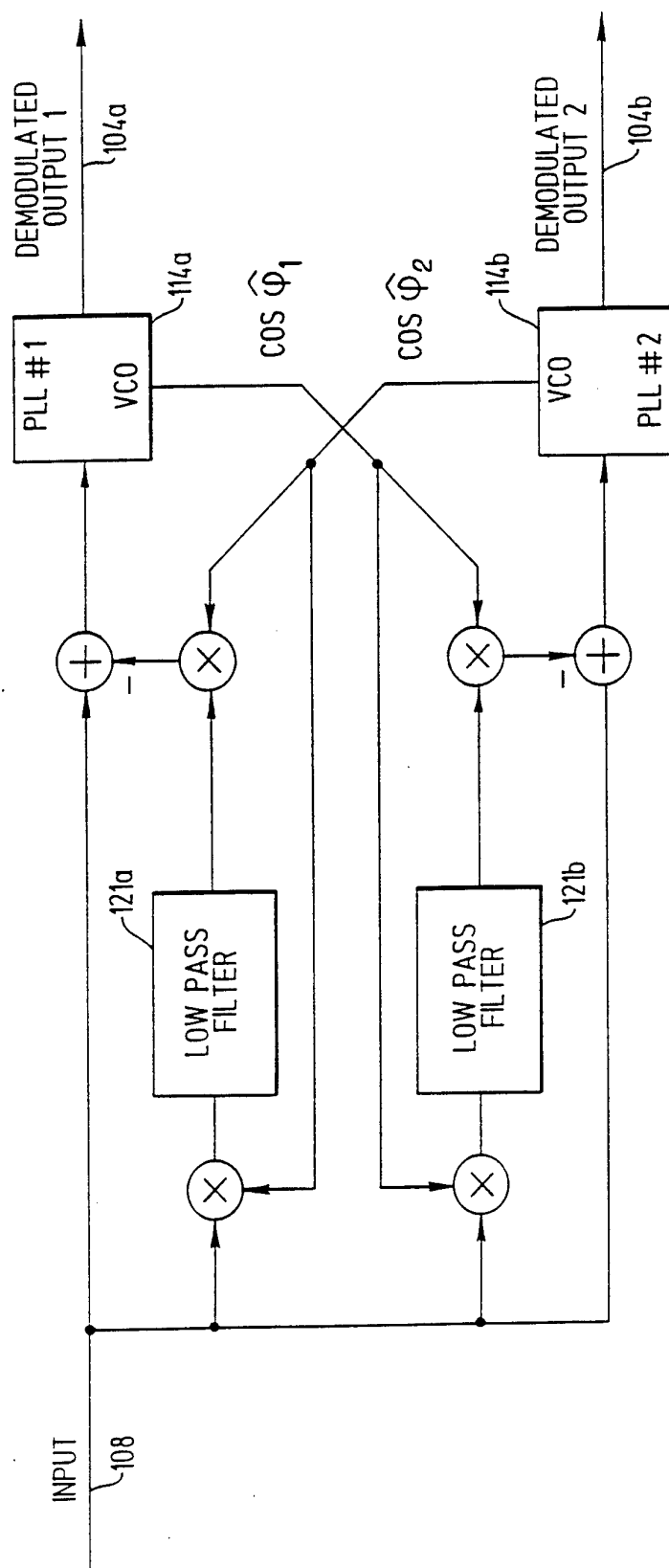


FIG. 12

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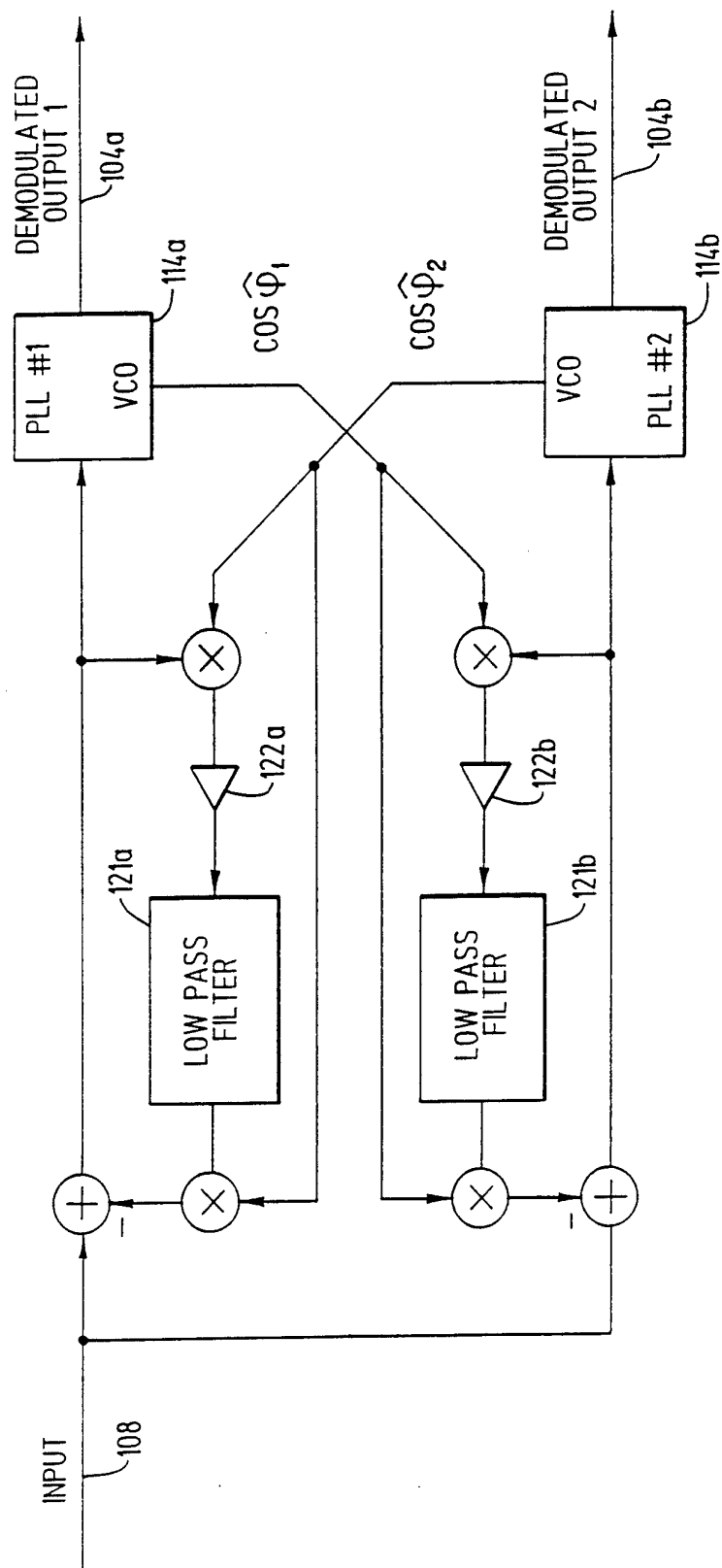


FIG. 13

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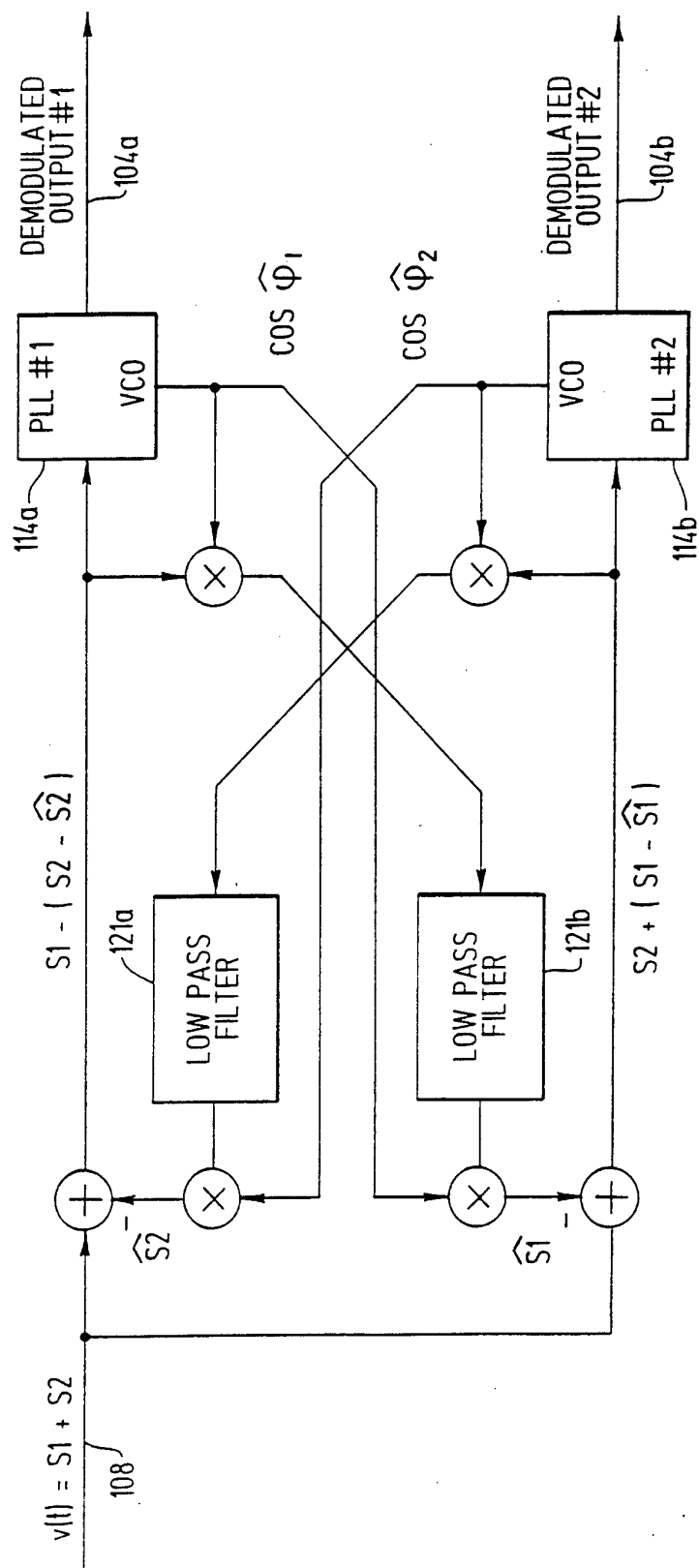


FIG. 14

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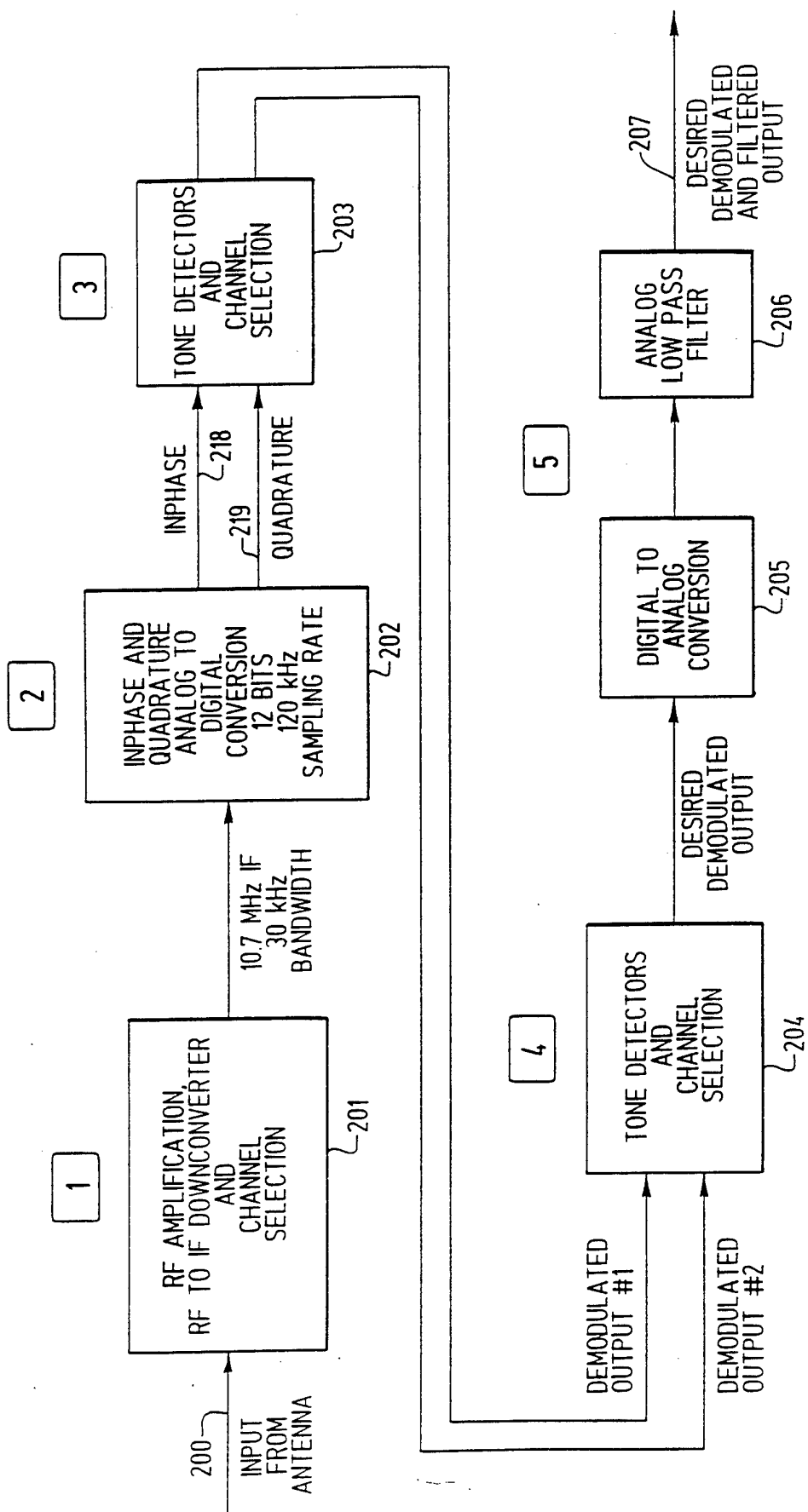


FIG. 15

17/22

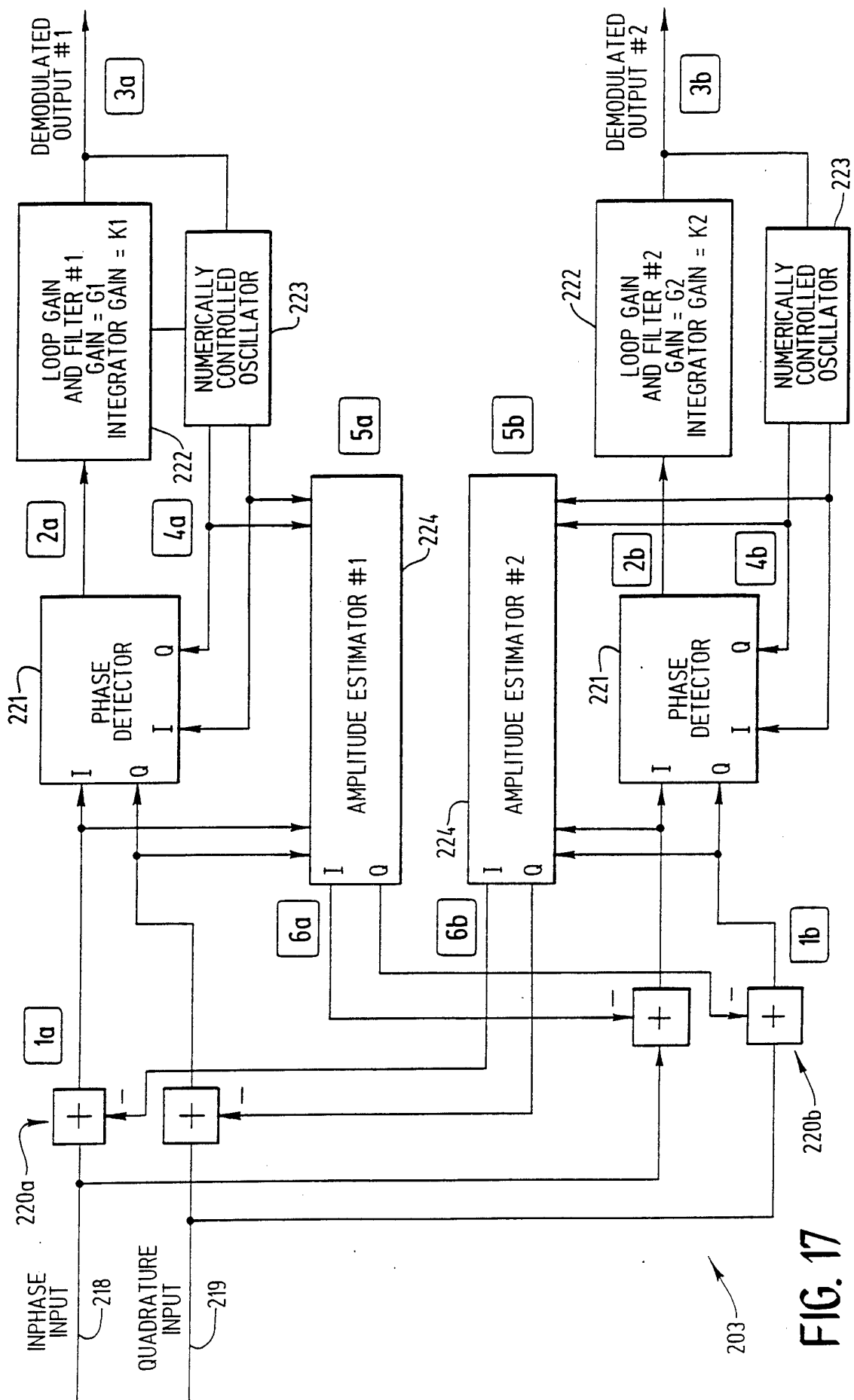


FIG. 17

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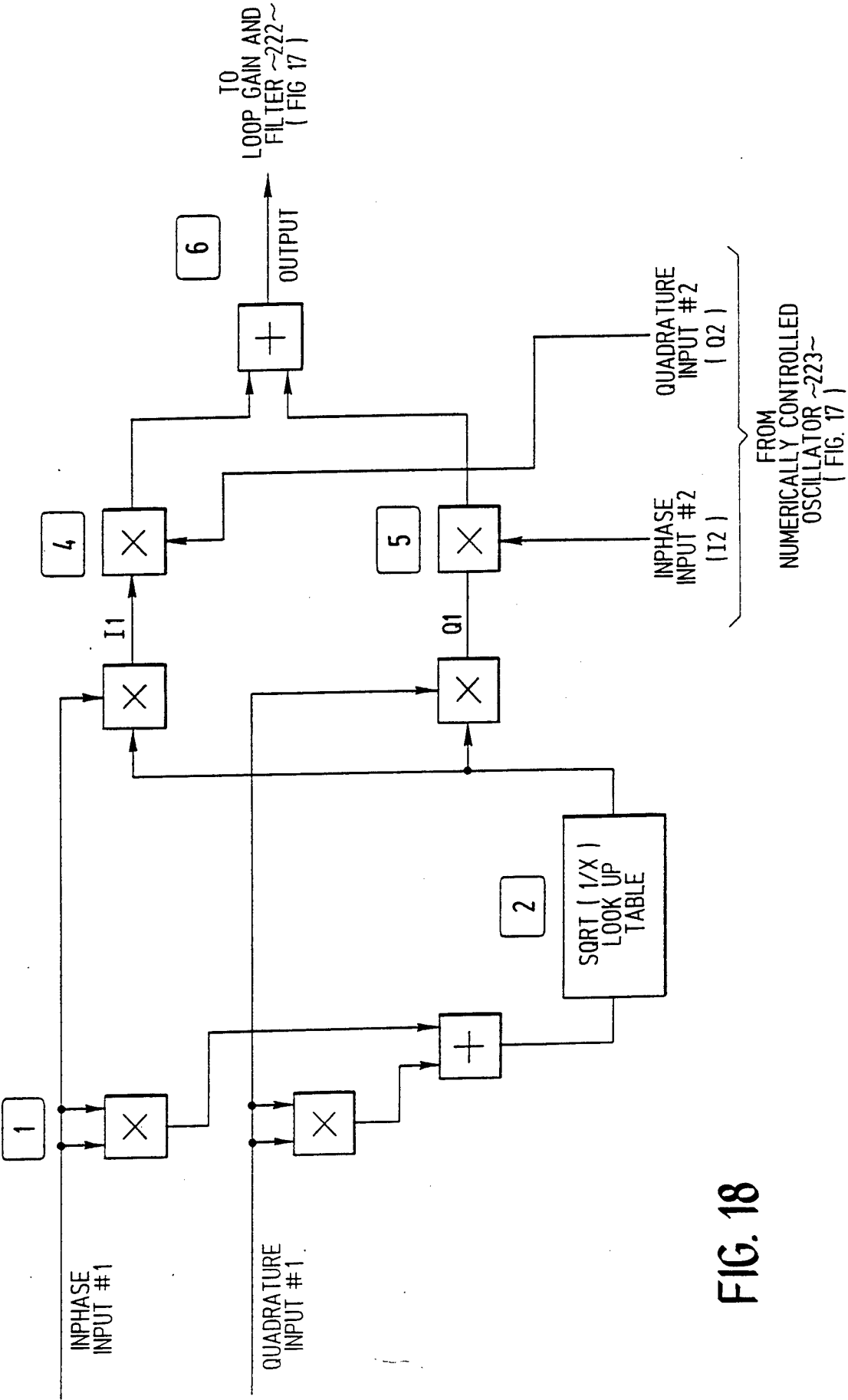


FIG. 18

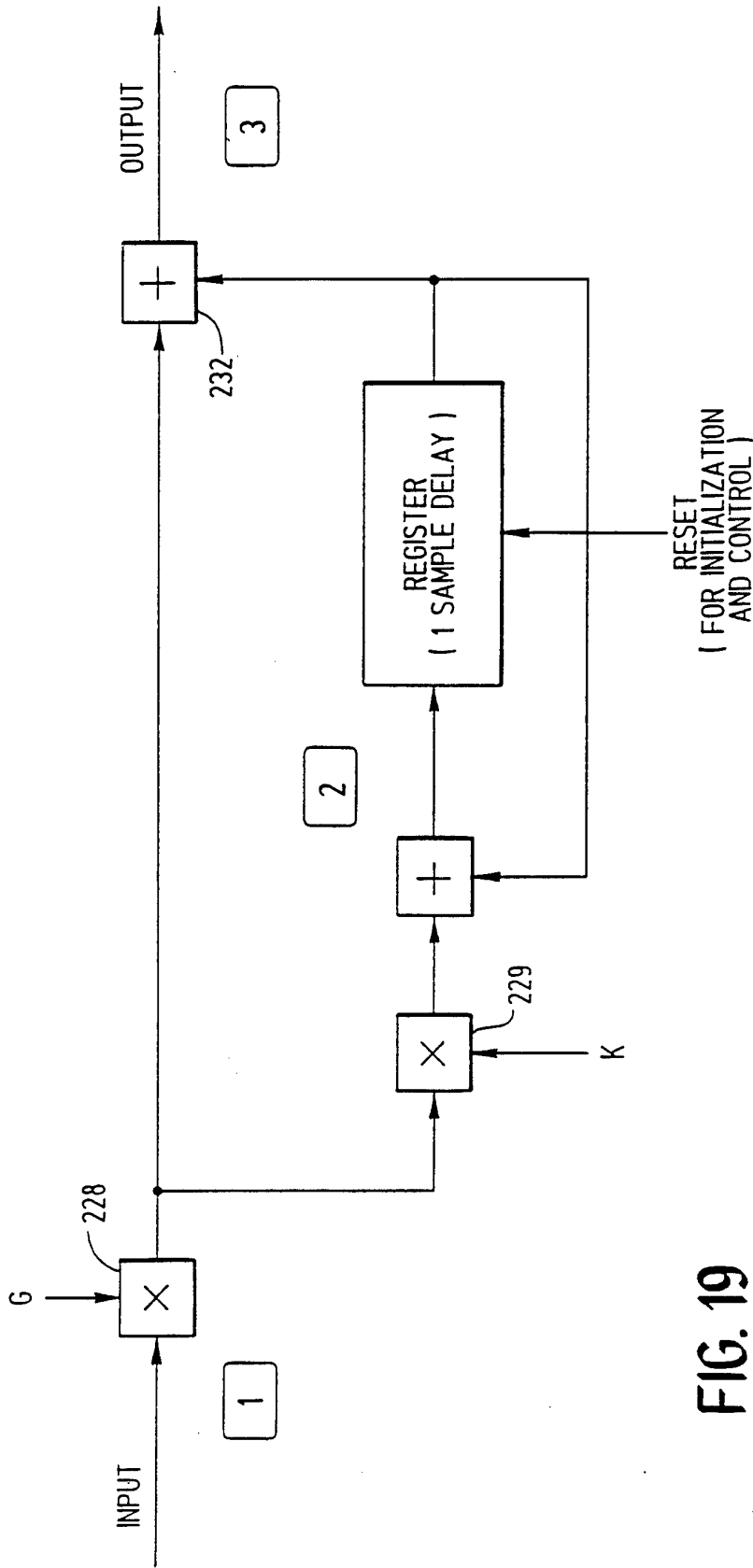


FIG. 19

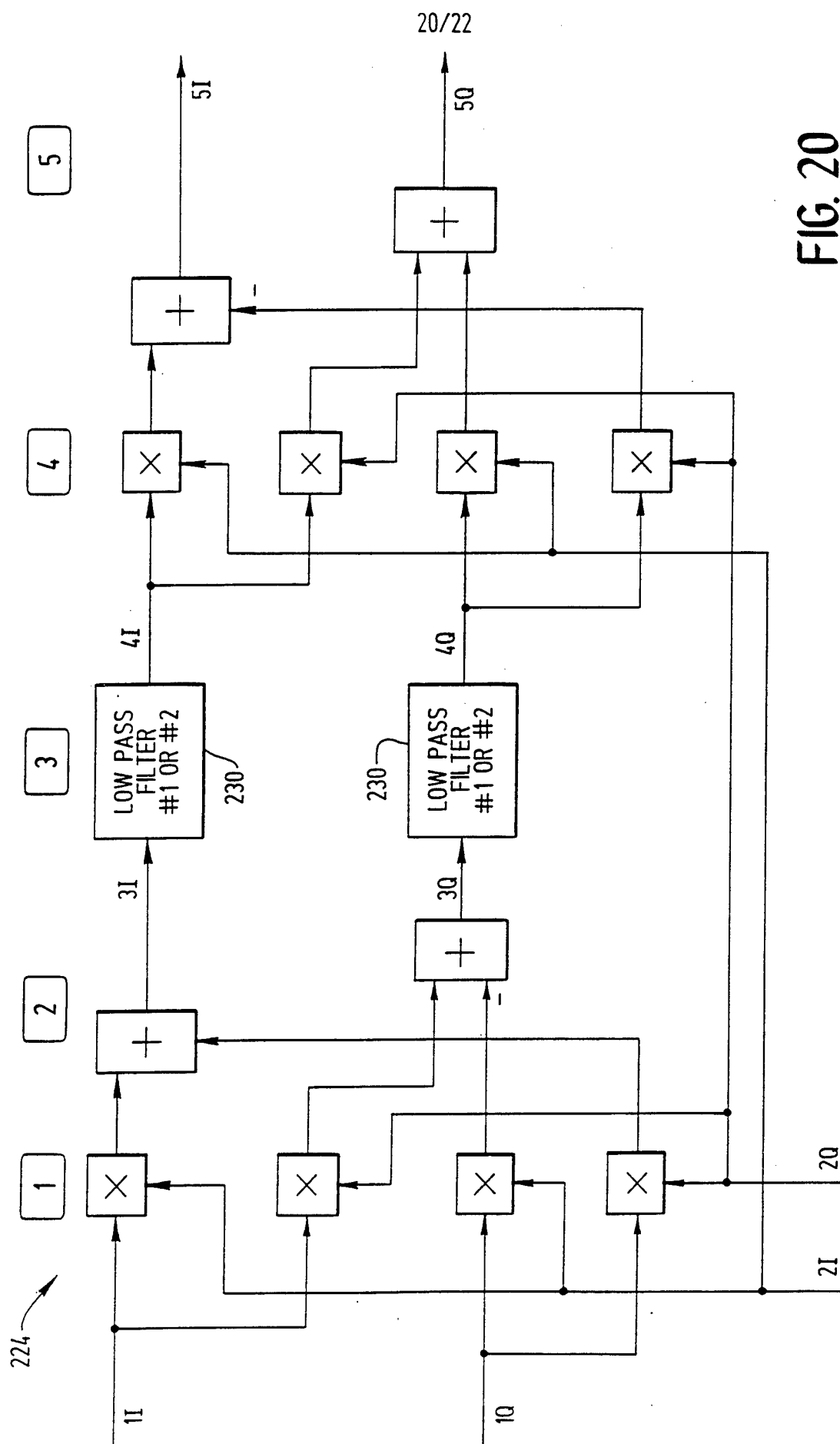


FIG. 20

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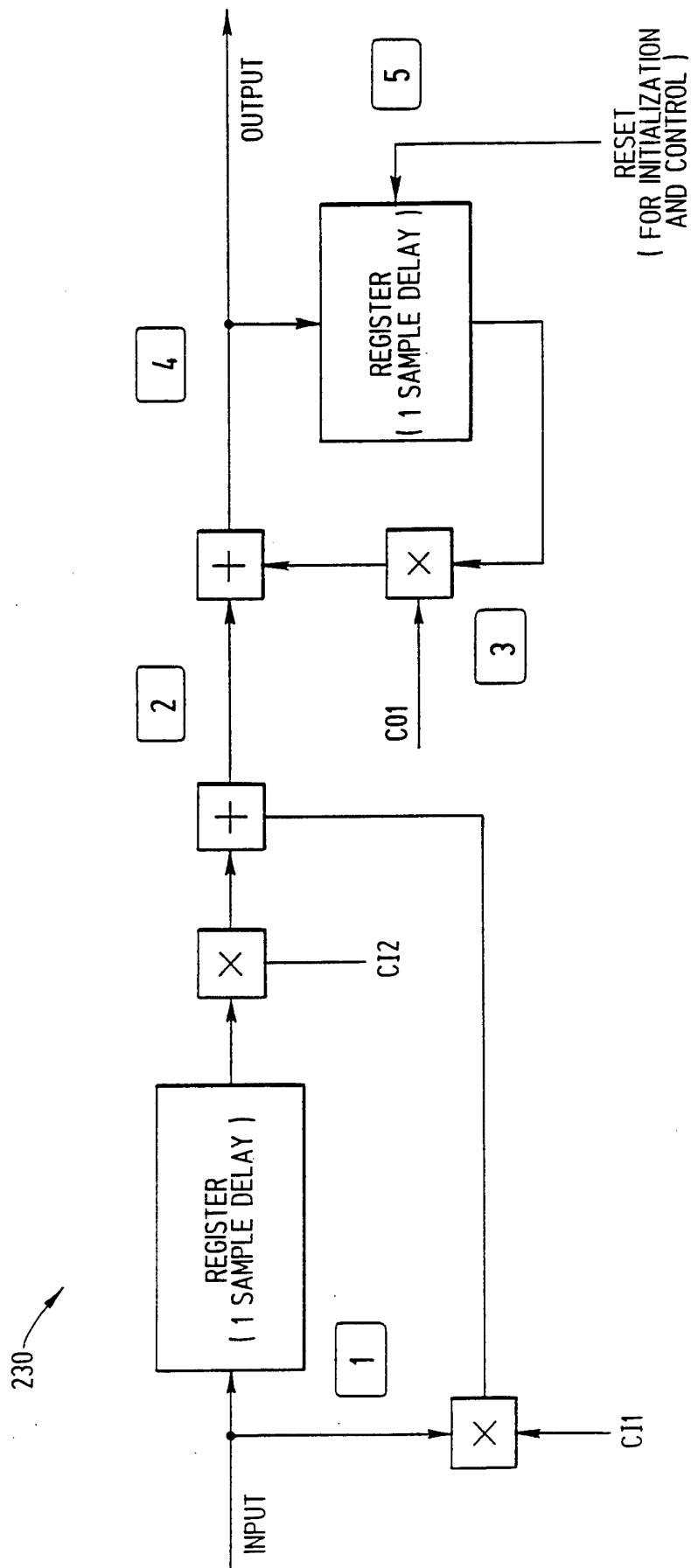


FIG. 21

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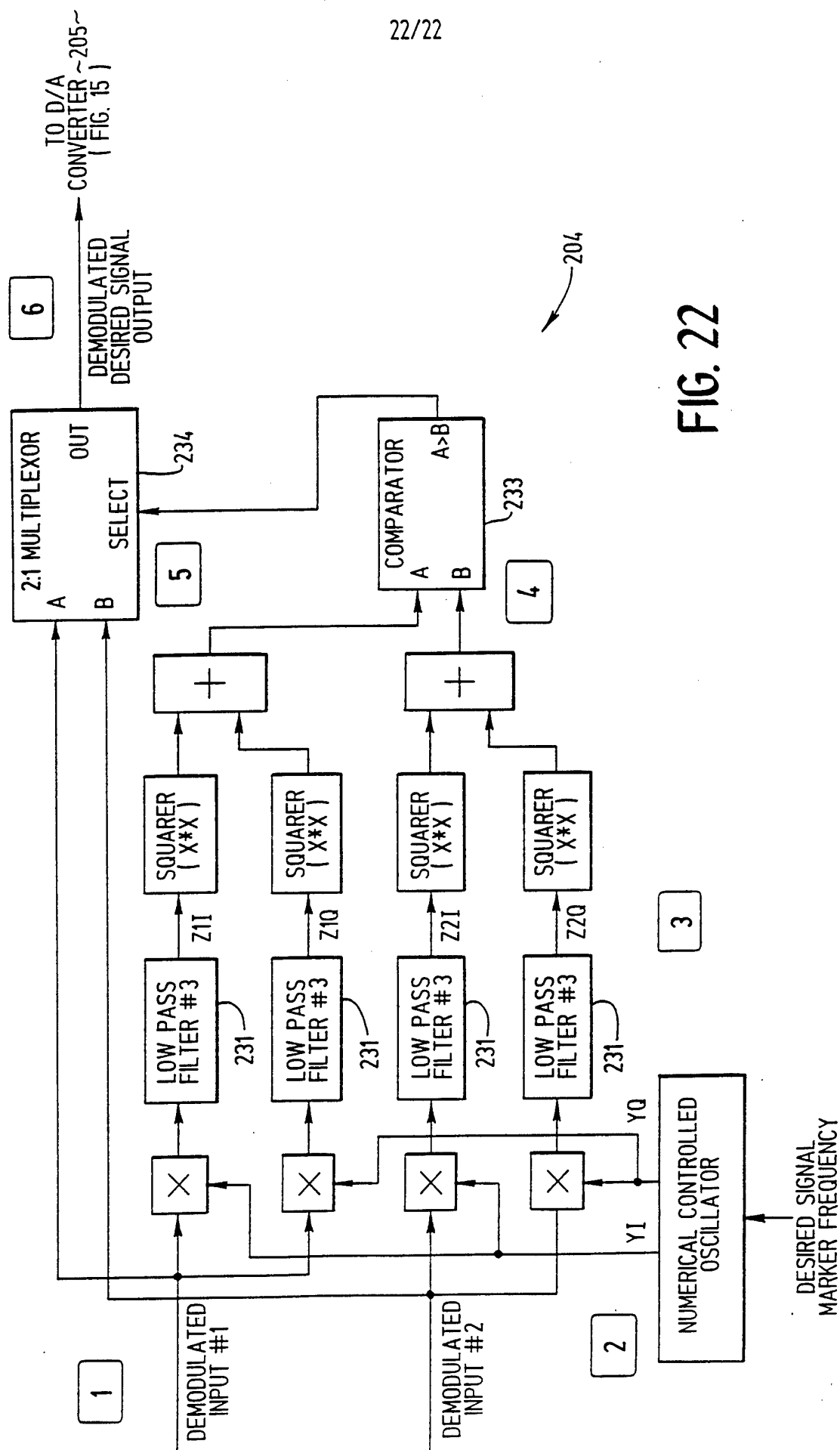


FIG. 22

INTERNATIONAL SEARCH REPORT

 Internat Application No
 PCT/US 93/06125

 A. CLASSIFICATION OF SUBJECT MATTER
 IPC 5 H04B1/12

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 5 H04B H04Q H03D H03L

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 practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US,A,3 873 931 (BASSE ET AL.) 25 March 1975 see abstract see column 2, line 27 - column 4, line 5 see sole figure --- -/--	1-4, 12-15, 22-29



Further documents are listed in the continuation of box C.



Patent family members are listed in annex.

* Special categories of cited documents :

- "A" document defining the general state of the art which is not considered to be of particular relevance
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- "O" document referring to an oral disclosure, use, exhibition or other means
- "P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search

10 November 1993

Date of mailing of the international search report

26. 11. 93

Name and mailing address of the ISA

 European Patent Office, P.B. 5818 Patentlaan 2
 NL - 2280 HV Rijswijk
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 Fax: (+31-70) 340-3016

Authorized officer

LYDON, M

INTERNATIONAL SEARCH REPORT

Internat Application No
PCT/US 93/06125

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	MILCOM 86 - PROCEEDINGS OF 1986 IEEE MILITARY COMMUNICATIONS CONFERENCE 5-9 October 1986, Monterey, California (US NEW YORK (US)	1-4, 12-15, 22-29
A	pages 47.1.1 - 47.1.5 S.SAY & F.CASSARA: "EXPERIMENTAL RESULTS ON THE CROSS-COUPLED PHASE-LOCKED LOOP INTERFERENCE CANCELLER WITH CLOSED LOOP AMPLITUDE CONTROL" see page 47.1.1, left column, line 11 - page 47.1.2, left column, line 7 see page 47.1.2, left column, line 45 - line 60 see figures 1,2,4 ---	7-11,30
A	TELECOMMUNICATIONS AND RADIO ENGINEERING vol. 35/36, no. 11 , November 1981 , WASHINGTON US pages 86 - 92 M.A.BYKHOVSKIY 'SINGLE-CHANNEL RADIO INTERFERENCE COMPENSATORS IN COMMUNICATION SYSTEMS' see page 86, line 1 - page 88, line 3 see page 89, line 45 - page 90, line 3 see page 91, line 22 - line 30 see figures 1,2 ---	1-4, 12-15, 22-29
A	PROCEEDINGS OF THE TWENTY-SECOND ASILOMAR CONFERENCE ON SIGNALS, SYSTEMS & COMPUTERS Pacific Grove, California (US) 31 October - 2 November 1988 NEW YORK (US) pages 270-273 E.H.SATORIUS ET AL.: "SUPPRESSION OF NARROWBAND INTERFERENCE VIA SINGLE CHANNEL ADAPTIVE PREPROCESSING" see abstract see page 270, left column, line 1 - right column, line 24 see page 271, right column, line 7 - line 28 see figure 2 -----	1,12, 22-27

INTERNATIONAL SEARCH REPORT

Information on patent family members

Internatic	Application No
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PCT/US 93/06125

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US-A-3873931	25-03-75	NONE	