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(71) Applicant: MOTOROLA INC. [US/US]; 1303 East Algonquin Road, Schaumburg, IL 60196 (US).

(72) Inventor: LING, Fuyun; 4190 Mumford Drive, Hoffman Estates, IL 60195 (US).

(74) Agents: PARMELEE, Steven, G. et al.; Motorola Inc., Intellectual Property Dept./KAB, 1303 East Algonquin Road, Schaumburg, IL 60196 (US).

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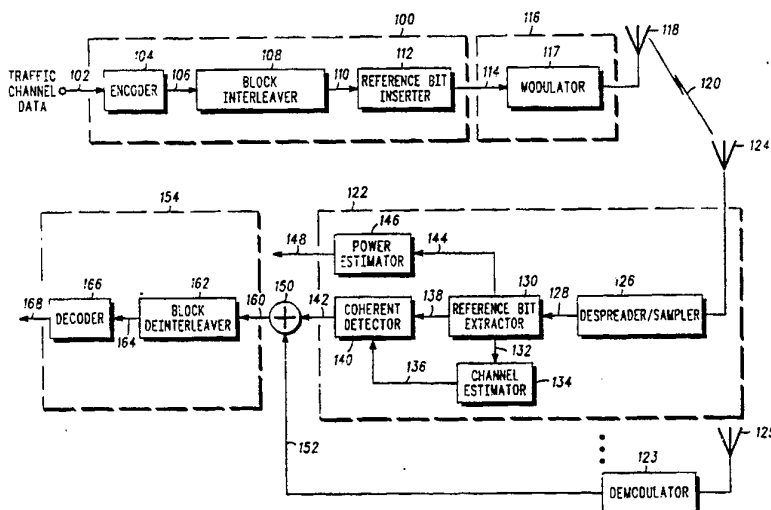
(57) Abstract

A method and apparatus is provided for encoding and decoding to facilitate coherent communication. In encoding, reference symbols are inserted into a stream of input data symbols (110) to form a reference coded stream of input data symbols (114). Subsequently, the reference coded stream of input data symbols are prepared for transmission over a communication channel by spreading the reference coded stream of input data symbols with a spreading code prior to transmission over the communication channel. In decoding, a received communication signal (120) is despread with a spreading code to derive a stream of reference samples (132) and a stream of data samples (138). The channel response is estimated by utilizing the stream of reference samples (132). Finally, an estimated data symbol is detected from the stream of data samples (138) by utilizing the estimated channel response.

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METHOD AND APPARATUS FOR COHERENT COMMUNICATION IN A SPREAD-SPECTRUM COMMUNICATION SYSTEM

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Field of the Invention

The present invention relates to communication systems which employ spread-spectrum signals and, more particularly, to a method
10   and apparatus for coherent communication in a spread-spectrum communication system.

Background of the Invention

15   Communication systems take many forms. In general, the purpose of a communication system is to transmit information-bearing signals from a source, located at one point, to a user destination, located at another point some distance away. A communication system generally consists of three basic components: transmitter, channel, and
20   receiver. The transmitter has the function of processing the message signal into a form suitable for transmission over the channel. This processing of the message signal is referred to as modulation. The function of the channel is to provide a physical connection between the transmitter output and the receiver input. The function of the receiver is
25   to process the received signal so as to produce an estimate of the

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original message signal. This processing of the received signal is referred to as demodulation.

One type of communication system is a multiple access spread-spectrum system. In a spread-spectrum system, a modulation technique

5   is utilized in which a transmitted signal is spread over a wide frequency band within the communication channel. The frequency band is much wider than the minimum bandwidth required to transmit the information being sent. A voice signal, for example, can be sent with amplitude modulation (AM) in a bandwidth only twice that of the information itself.

10   Other forms of modulation, such as low deviation frequency modulation (FM) or single sideband AM , also permit information to be transmitted in a bandwidth comparable to the bandwidth of the information itself. However, in a spread-spectrum system, the modulation of a signal to be transmitted often includes taking a baseband signal (e.g., a voice

15   channel) with a bandwidth of only a few kilohertz, and distributing the signal to be transmitted over a frequency band that may be many megahertz wide. This is accomplished by modulating the signal to be transmitted with the information to be sent and with a wideband encoding signal.

20   Three general types of spread-spectrum communication techniques exist, including direct sequence modulation, frequency and/or time hopping modulation, and chirp modulation. In direct sequence modulation, a carrier signal is modulated by a digital code sequence whose bit rate is much higher than the information signal

25   bandwidth.

Information (i.e. the message signal consisting of voice and/or data) can be embedded in the direct sequence spread-spectrum signal by several methods. One method is to add the information to the spreading code before it is used for spreading modulation. It will be

30   noted that the information being sent must be in a digital form prior to adding it to the spreading code, because the combination of the spreading code and the information typically a binary code involves modulo-2 addition. Alternatively, the information or message signal may be used to modulate a carrier before spreading it.

35   These direct sequence spread-spectrum communication systems can readily be designed as multiple access communication systems. For example, a spread-spectrum system may be designed as a direct

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sequence code division multiple access (DS-CDMA) system. In a DS-CDMA system, communication between two communication units is accomplished by spreading each transmitted signal over the frequency band of the communication channel with a unique user spreading code.

5   As a result, transmitted signals are in the same frequency band of the communication channel and are separated only by unique user spreading codes. These unique user spreading codes preferably are orthogonal to one another such that the cross-correlation between the spreading codes is approximately zero.

10   Particular transmitted signals can be retrieved from the communication channel by despreading a signal representative of the sum of signals in the communication channel with a user spreading code related to the particular transmitted signal which is to be retrieved from the communication channel. Further, when the user spreading

15   codes are orthogonal to one another, the received signal can be correlated with a particular user spreading code such that only the desired user signal related to the particular spreading code is enhanced while the other signals for all of the other users are not enhanced.

It will be appreciated by those skilled in the art that several

20   different spreading codes exist which can be used to separate data signals from one another in a DS-CDMA communication system. These spreading codes include but are not limited to pseudonoise (PN) codes and Walsh codes. A Walsh code corresponds to a single row or column of the Hadamard matrix.

25   Further it will be appreciated by those skilled in the art that spreading codes can be used to channel code data signals. The data signals are channel coded to improve performance of the communication system by enabling transmitted signals to better withstand the effects of various channel impairments, such as noise,

30   fading, and jamming. Typically, channel coding reduces the probability of bit error, and/or reduces the required signal to noise ratio usually expressed as error bits per noise density (i.e., E_b/N_0 which is defined as the ratio of energy per information-bit to noise-spectral density), to recover the signal at the cost of expending more bandwidth than would

35   otherwise be necessary to transmit the data signal. For example, Walsh codes can be used to channel code a data signal prior to modulation of

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the data signal for subsequent transmission. Similarly PN spreading codes can be used to channel code a data signal.

However, channel coding alone may not provide the required signal to noise ratio for some communication system designs which require the system to be able to handle a particular number of simultaneous communications (all having a minimum signal to noise ratio). This design constraint may be satisfied, in some instances, by designing the communication system to coherently detect transmitted signals rather than using non-coherent reception techniques. It will be appreciated by those skilled in the art that a coherent receiver requires less signal to noise ratio (in E_b/N_0) than that required by a non-coherent receiver having the same bit error rate (i.e., a particular design constraint denoting an acceptable interference level). Roughly speaking, there is a three deciBel (dB) difference between them for the Rayleigh fading channel. The advantage of the coherent receiver is more significant when diversity reception is used, because there is no combining loss for an optimal coherent receiver while there is always a combining loss for noncoherent receiver.

One such method for facilitating coherent detection of transmitted signals is to use a pilot signal. For example, in a cellular communication system the forward channel, or down-link, (i.e., from base station to mobile unit) may be coherently detected, if the base station transmits a pilot signal. Subsequently, all the mobile units use the pilot channel signal to estimate the channel phase and magnitude parameters. However, for the reverse channel, or up-link, (i.e., from mobile to base station), using such a common pilot signal is not feasible. As a result, those of ordinary skill in the art often assume that only non-coherent detection techniques are suitable for up-link communication.

As a result, many recent publications have focused on optimizing non-coherent reception in DS-CDMA systems. See for example the following articles.

- A. Salmasi and K.S. Gilhousen, "On The System Design Aspects of Code Division Multiple Access (CDMA) Applied to Digital Cellular And Personal Communications Networks," *Proc. of VTC'91*, pp. 57-62, 1991.

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- •   F. Ling and D. Falconer, "Orthogonal/Convolutional Coding
        for Reverse Channel CDMA Communication," *Proc. of
        VTC'92,* pp. 63-66, May, 1992, Denver, CO.

5   •   L. F. Chang and N. R. Sollenberger, "Comparison of Two
          Interleaving Techniques for CDMA Radio Communication
          Systems," *Proc. of VTC'92,* pp. 275-278, May, 1992,
          Denver, CO.

10 •   Y. J. Liu, "Soft Decision Decoding for a Bit-Interleaved
          Convolutionally Encoded Code Division Multiple Access
          System over Rayleigh Fading Channel," *Proc. of PIMRC'92,*
          pp. 128-132, Oct. 1992.

15   Each of these articles show that a substantial difference in performance
        exists when different coding, modulation, detection and interleaving
        techniques are used for up-link communication in cellular
        communication systems.
                In the A. Salmasi and K.S. Gilhousen article, a DS-CDMA
20   communication system is described which uses bit-by-bit interleaving
        within convolutional and orthogonal coding scheme to optimize non-
        coherent reception in DS-CDMA communication systems.
                In the F. Ling and D. Falconer article as well as the L. F. Chang
        and N. R. Sollenberger article, an up-link DS-CDMA system that
25   employs Walsh coding (i.e., orthogonal coding), non-coherent detection
        and using orthogonal symbol (i.e., word-by-word) interleaving instead of
        bit-by-bit interleaving was disclosed. The L. F. Chang and N. R.
        Sollenberger article shows that a word-by-word interleaved
        convolutional and orthogonal coding scheme requires about 1 to 1.4 dB
30   less E_b/N_0 than the similar bit-by-bit interleaving scheme described by in
        the A. Salmasi and K.S. Gilhousen article when the communication
        system which employs either scheme also utilizes power control of
        mobile communication unit which move at different speeds (e.g., move
        at rate ranging from 0 to 100 kilometers per hour). While the word-by-
35   word interleaving convolutional/orthogonal coding scheme has better
        performance than the bit-by-bit one, it has less implicit diversity than the

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latter. Moreover, it is still a non-coherent communication system and the combing loss can not be avoided.

Finally, the Y. J. Liu article describes a more sophisticated detection technique in which the performance of the up-link DS-CDMA communication system with Walsh coding and bit-level interleaving can be improved with a 4-port diversity combining without changing the interleaving method.

However, even in view of the above-described improvements for non-coherent communication systems, a need still exists for a communication system which employs coherent detection techniques.

Summary of the Invention

A method and apparatus is provided for encoding and decoding to facilitate coherent communication. In encoding, reference symbols are inserted into a stream of input data symbols to form a reference coded stream of input data symbols. Subsequently, the reference coded stream of input data symbols are prepared for transmission over a communication channel by spreading the reference coded stream of input data symbols with a spreading code prior to transmission over the communication channel. In decoding, a received communication signal is despread with a spreading code to derive a stream of reference samples and a stream of data samples. The channel response is estimated by utilizing the stream of reference samples. Finally, an estimated data symbol is detected from the stream of data samples by utilizing the estimated channel response.

Brief Description of the Drawings

FIG. 1 is a block diagram showing a preferred embodiment communication system in accordance with the present invention.

FIG. 2 is a block diagram showing a preferred embodiment communication channel frame structure for use in the preferred embodiment communication system shown in FIG. 1.

FIG. 3 is a block diagram showing a preferred embodiment channel estimator for use in the preferred embodiment communication system shown in FIG. 1.

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Detailed Description

In the course of the following discussion, a new approach for up-
5   link DS-CDMA communication is presented. This new approach
employs coherent-detection with reference-symbol based channel
estimation. It will be appreciated by those skilled in the art that other
types of communication systems (e.g., personal communication
systems, trunked systems, satellite communication systems, data
10   networks, and the like) may also be adapted and/or designed to use the
principles described herein. It will be shown that a substantial gain in
E_b/N_0 can be obtained relative to non-coherent detection techniques by
applying such a coherent detection method to up-link DS-CDMA
communication. In particular, simulation results have shown that the
15   required E_b/N_0 by using this new scheme is about 2.5 dB lower than
non-coherent detection of Walsh coding with bit-by-bit interleaving or
1.3 dB lower than non-coherent detection with Walsh symbol (i.e., word-
by-word) interleaving over the entire range of practical mobile
communication unit speeds (i.e., speeds of 0 to 100 kilometers per
20   hour). The analysis of this new scheme is given in the frequency-
domain. This frequency-domain analysis results in a simple formula that
characterizes the performance loss of such a scheme relative to perfect
coherent detection.

In order to perform effective coherent detection, it is necessary to
25   obtain an accurate channel estimate. There are basically two types of
channel estimation methods: data-based and reference-based. Data-
based channel estimation may be implemented as decision-directed or
non-decision-directed. For DS-CDMA up-link communication, the
channel estimator must operate at low signal-to-noise ratios and the
30   fading is relatively fast. As a result, the decision-directed approach is
not appropriate due to the high decision error rate. On the other hand, a
non-decision-directed method, such as the one described in the article
by A. J. Viterbi and A. M. Viterbi, "Nonlinear Estimation of PSK-
Modulated Carrier Phase with Application to Burst Digital
35   Transmission," *IEEE Trans. on Info. Theory*, Vol. IT-29, No. 4, pp. 543-
551, Jul. 1983, always has a phase ambiguity, e.g., 180° ambiguity for
binary phase shift keying (BPSK) or 90° ambiguity for quadrature phase

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shift keying (QPSK), in the channel estimate. As a consequence, it is necessary to use differential coding to eliminate its effect. However, as will be appreciated by those skilled in the art, in communication systems having a differential coded signal transmitted over Rayleigh fading

5   channels, even with coherent detection, still need over 3 dB higher E_b/N_0 than non-differentially coded phase shift keying (PSK) signaling.

One way to solve the decision error and phase ambiguity problem is to use reference symbols for channel estimation. Reference-symbol-based channel estimation is described as follows. Reference

10   symbols known to the receiver are inserted into a sequence of information bearing data symbols, which may be coded symbols. At the receiver, the received signal samples corresponding to the reference symbols are used to generate a channel estimate. Because the reference symbols are known to the receiver, there are no decision

15   errors and the resulting channel estimate does not have a phase ambiguity. As a result, a robust communication system with non-differentially coded signaling is provided.

The inserted reference symbols can be organized in blocks or uniformly distributed. For a flat fading channel, it is desirable to insert

20   reference symbols periodically and uniformly in the data stream. For a DS-CDMA up-link with a RAKE receiver for front end processing, we can treat the output of each RAKE "finger" as being a flat faded signal. Thus, the preferred embodiment communication system will uniformly insert one reference symbol for every M coded data symbols.

25   The basic operation of RAKE receivers are described in an article by R. Price and P.E. Green, Jr., "A Communication Technique for Multipath Channels," *Proceedings of the IRE*, March 1958, pages 555-570. Briefly, a RAKE receiver performs a continuous, detailed measurement of the multipath characteristic of a received signal. This

30   knowledge is then exploited to combat the selective fading by detecting the echo signals individually, using a correlation method, and algebraically combining those echo signals into a single detected signal. The intersymbol interference is attenuated by varying the time delay or phase between the various detected echo signals prior to their

35   algebraic combination.

Referring now to FIG. 1, a system for coherent communication in a spread-spectrum communication system is shown. In the encoding

portion **100** of the communication system, traffic channel data bits **102** are input to an encoder **104** at a particular bit rate (e.g., 9.6 kilobit/second). The input traffic channel data bits can include either voice converted to data by a vocoder, pure data, or a combination of the

5   two types of data. Encoder **104** encodes the input data bits **102** into data symbols at a fixed encoding rate (1/r) with an encoding algorithm which facilitates subsequent maximum likelihood decoding of the data symbols into data bits (e.g. convolutional or block coding algorithms). For example, encoder **104** encodes input data bits **102** (e.g., 192 input

10   data bits that were received at a rate of 9.6 kilobits/second) at a fixed encoding rate of one data bit to three data symbols (i.e., 1/3) such that the encoder **102** outputs data symbols **106** (e.g., 576 data symbols output at a 28.8 kilosymbols/second rate).

       The data symbols **106** are then input into an interleaver **108**.

15   Interleaver **108** organizes the data symbols **106** into blocks (i.e., frames) and block interleaves the input data symbols **106** at the symbol level. In the interleaver **108**, the data symbols are individually input into a matrix which defines a predetermined size block of data symbols. The data symbols are input into locations within the matrix so that the matrix

20   is filled in a column by column manner. The data symbols are individually output from locations within the matrix so that the matrix is emptied in a row by row manner. Typically, the matrix is a square matrix having a number of rows equal to the number of columns; however, other matrix forms can be chosen to increase the output interleaving

25   distance between the consecutively input non-interleaved data symbols. The interleaved data symbols **110** are output by the interleaver **108** at the same data symbol rate that they were input (e.g., 28.8 kilosymbols/second). The predetermined size of the block of data symbols defined by the matrix is derived from the maximum number of

30   data symbols which can be transmitted at a coded bit rate within a predetermined length transmission block. For example, if data symbols **106** are output from the encoder **104** at a 28.8 kilosymbols/second rate, and if the predetermined length of the transmission block is 20 milliseconds, then the predetermined size of the block of data symbols is

35   28.8 kilosymbols/second times 20 milliseconds which equals 576 data symbols which defines a 18 by 32 matrix.

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The interleaved data symbols 110 are then input to a reference bit inserter 112 which inserts L known reference bits for every M interleaved data symbols 110. To simplify the following discussion, it will be assumed that L=1 and M=6. In addition, it will be assumed that

5   each inserted reference bit is a zero bit. However, it will be appreciated by those skilled in the art that L and M could be any other value without departing from the scope and spirit of the present invention. In addition, the reference bits could be any known sequence such as all one bits or several one bits followed by several zero bits without departing from the

10   scope and spirit of the present invention. When L=1 and M=6, reference bit inserter 112 outputs 672 reference-coded bits 114 for each block (i.e., frame) such that a reference bit is inserted between each group of six data symbols. An example of a transmitted block (i.e., frame) of reference-coded data symbols 114 consisting of 42 bits is shown in FIG.

15   2 (where each d represents a data symbol and each r represents a reference bit).

The reference-coded data symbols 114 is output from encoding portion 100 of the communication system and input to a transmitting portion 116 of the communication system. The data symbols 114 are

20   prepared for transmission over a communication channel by a modulator 117. Subsequently, the modulated signal is provided to an antenna 118 for transmission over the communication channel 120.

The modulator 117 preferably prepares the data symbols 114 for direct sequence code divided spread-spectrum transmission by deriving

25   a sequence of fixed length codes from the reference-coded data symbols 114 in a spreading process. For example, each of the data symbols within the stream of reference-coded data symbols 114 may be spread to a unique nine bit length code such that a group of six data symbols is represented by a single 54 bit length code. In addition, each

30   reference bit within the stream of reference-coded data symbols 114 may select a ten bit length code. The codes representing the group of six data symbols and an associated reference bit preferably are combined to form a single 64 bit length code. As a result of this spreading process, the modulator 117 which received the reference-

35   coded data symbols 114 at a fixed rate (e.g., 28.8 kilosymbols/second) now has a spread sequence of 64 bit length codes having a higher fixed symbol rate (e.g., 307.2 kilosymbols/second). It will be appreciated by

those skilled in the art that the reference bits and data symbols within
the stream of reference-coded data symbols **114** may be spread
according to numerous other algorithms into a sequence of larger length
codes without departing from the scope and spirit of the present

5   invention.

The spread sequence is further prepared for direct sequence
code divided spread-spectrum transmission by further spreading the
spread sequence with a long spreading code (e.g. PN code). The
spreading code is a user specific sequence of symbols or unique user

10   code which is output at a fixed chip rate (e.g., 1.228 Megachips/second).
In addition to providing an identification as to which user sent the
encoded traffic channel data bits **102** over the communication channel
120, the unique user code enhances the security of the communication
in the communication channel by scrambling the encoded traffic

15   channel data bits **102**. In addition, the user code spread encoded data
bits (i.e. data symbols) are used to bi-phase modulate a sinusoid by
driving the power level controls of the sinusoid. The sinusoid output
signal is bandpass filtered, translated to an RF frequency, amplified,
filtered and radiated by an antenna **118** to complete transmission of the

20   traffic channel data bits **102** in a communication channel **120** with
BPSK modulation.

A receiving portion **122** of the communication system receives
the transmitted spread-spectrum signal from over the communication
channel **120** through antenna **124**. The received signal is sampled

25   into data samples by despreader and sampler **126**. Subsequently, the
data samples **142** are output to the decoding portion **154** of the
communication system.

The despreader and sampler **126** preferably BPSK samples the
received spread-spectrum signal by filtering, demodulating, translating

30   from the RF frequencies, and sampling at a predetermined rate (e.g.,
1.2288 Megasamples/second). Subsequently, the BPSK sampled
signal is despread by correlating the received sampled signals with the
long spreading code. The resulting despread sampled signal **128** is
sampled at a predetermined rate and output to a reference bit extractor

35   **130** (e.g., 307.2 kilosamples/second so that a sequence of four samples
of the received spread-spectrum signal is despread and/or represented
by a single data sample).

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   The reference bit extractor **130** preferably extracts the reference bits **132** from the despread sampled signal **128** and outputs the reference bits **132** to a channel estimator **134**. The remaining data samples **138** from the despread sampled signal **128** are output to a

5 coherent detector **140** for later coherent detection of data samples **142**.

   The channel estimator **134** correlates the extracted reference bits **132** with a known reference sequence of data samples to obtain unbiased, but noisy, channel estimates. In order to obtained a better channel estimate **136**, these noisy estimates may be passed through a

10 low-pass filter, which may be fixed or adaptive, to remove the high frequency noise components. The resulting channel estimates **136** are relatively noise-free and can be used for coherent detection. It should be noted that the low pass filtering only gives us a channel estimate for every $(M+1)T$, where M is the number of data symbols between each

15 reference bit inserted by reference bit inserter **112** (e.g., $M=6$) and where T is the time interval of each data sample. To perform coherent detection of transmitted data symbols, we need to have a channel estimate for every T. When $(M+1)T$ is short relative to the channel variation time constant, a simple but effective method to get a channel

20 estimate for every T is to perform linear interpolation between two channel estimates separated by $(M+1)T$. however, as will be appreciated by those skilled in the art more sophisticated interpolation techniques may be used if necessary.

   In the preferred embodiment coherent communication system,

25 power control may also be used to enhance overall system performance. The power control algorithms may be very similar to the algorithms used in non-coherent communication systems. The preferred embodiment power control algorithm preferably includes estimating received power every 1.25 ms (i.e., each block or frame), or

30 every 12 information bits, i.e., every 36 encoded bits or 42 total received signal samples. The power estimate may be calculated with several different techniques. One technique is to compute a channel estimate with a power estimator **146** by using simply the six reference signal samples (i.e., reference bits **144** from reference bit extractor **130**)in a

35 42 bit length block. The square of the magnitude of the channel estimate is then output by the power estimator **146** as the power estimate **148**.

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　　　　After channel estimates **136** are generated, the rest of the
receiver is conventional. The coherent detector **140** multiplies the
remaining data samples **138** from the despread sampled signal **128** by
the conjugate of the channel estimates **136** to generate coherently
5　　detected samples **142**.

　　　　As will be appreciated by those skilled in the art, multiple
receiving portions **122** through **123** and antennae **124** through **125**,
respectively, to achieve space diversity. The Nth receiver portion would
operate in substantially the same manner to retrieve data samples from
10　　the received spread-spectrum signal in communication channel **120** as
the above described receiving portion **122**. The outputs **142** through
152 of the N receiving portions preferably are input to a multiplier **150**
which diversity combines the input data samples into a composite
stream of coherently detected data samples **160**.

15　　　　The individual data samples **160** which form soft decision data
are then input into a decoding portion **154** including a deinterleaver
162 which deinterleaves the input soft decision data **160** at the
individual data level. In the deinterleaver **162**, the soft decision data
160 are individually input into a matrix which defines a predetermined
20　　size block of soft decision data. The soft decision data are input into
locations within the matrix so that the matrix is filled in a row by row
manner. The deinterleaved soft decision data **164** are individually
output from locations within the matrix so that the matrix is emptied in a
column by column manner. The deinterleaved soft decision data **164**
25　　are output by the deinterleaver **162** at the same rate that they were
input (e.g., 28.8 kilometrics/second).

　　　　The predetermined size of the block of soft decision data defined
by the matrix is derived from the maximum rate of sampling data
samples from the spread-spectrum signal received within the
30　　predetermined length transmission block.

　　　　The deinterleaved soft decision data **164**, are input to a decoder
166 which uses maximum likelihood decoding techniques to generate
estimated traffic channel data bits **168**. The maximum likelihood
decoding techniques may be augmented by using an algorithm which is
35　　substantially similar to a Viterbi decoding algorithm. The decoder **166**
uses a group of the individual soft decision data **164** to form a set of soft
decision transition metrics for use at each particular time state of the

maximum likelihood sequence estimation decoder **166**. The number of soft decision data **164** in the group used to form each set of soft decision transition metrics corresponds to the number of data symbols **106** at the output of the convolutional encoder **104** generated from
5   each input data bit **102**. The number of soft decision transition metrics in each set is equal to two raised to the power of the number of soft decision data **164** in each group. For example, when a 1/3 convolutional encoder is used in the transmitter, three data symbols **106** are generated from each input data bit **102**. Thus, decoder **166**
10   uses groups of three individual soft decision data **164** to form eight soft decision transition metrics for use at each time state in the maximum likelihood sequence estimation decoder **166**. The estimated data bits **168** are generated at a rate related to the rate that the soft decision data **164** are input to the decoder **166** and the fixed rate used to originally
15   encode the input data bits **102** (e.g., if the soft decision data are input at 28.8 kilometrics/second and the original encoding rate was 1/3 then estimated data bits **168** are output at a rate of 9600 bits/second).

Thus, a communication system for coherently encoding and decoding has been described above with reference to FIG. 1. In
20   summary, the communication system includes a first portion which encodes input data bits into data symbols, interleaves the data symbols in a symbol by symbol manner, inserts reference bits into the interleaved symbols, modulates and transmits the reference-coded data symbols over a communication channel. The communication system further
25   includes a second portion which receives and demodulates a signal from over the communication channel, estimates parameters of the communication channel, coherently demodulates data samples within received signal, deinterleaves the coherently detected data samples which are used as soft decision data within each received transmission
30   block, subsequently generates soft decision transition metrics from the deinterleaved individual soft decision data, and subsequently generates estimated data bits from the soft decision metrics by using maximum likelihood decoding techniques.

To more clearly described the reasoning behind this method of
35   using reference bits to coherently detect data samples and the operation of the channel estimator **134** as well as to facilitate further discussion, let us establish the following mathematical model.

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It will be appreciated by those skilled in the art that a fading channel can be modeled as a complex valued function of time t, denoted by h(t). The time interval of the convolutionally encoded-reference bits is denoted by T. The received signal after demodulation

5   and despreading is then sampled at every T. Assuming that one reference bit is inserted for every 6 encoded bits, the samples corresponding to the reference bit appears at $nT_r = 7nT$, n = ..., -1, 0, 1, ..., where $T_r = 7T$ is defined. Then, the noisy estimates may be written as:

10

$$\hat{h}(nT_r) = h(nT_r) + z(nT_r) \qquad \text{(eq. 1)}$$

where $z(nT_r)$ is the sampled additive noise. A better estimate than that shown in (eq. 1) can be obtained such that:

15

$$\bar{h}(nT_r) = \frac{1}{2N+1} \sum_{k=-N}^{N} \hat{h}(nT_r + kT_r) \qquad \text{(eq. 2)}.$$

For example N = 3 may be chosen. By doing so, an estimate is obtained for every reference bit interval T_r using 7 of the noisy estimates.

20   To obtain the channel estimate at kT for k ≠ 7n which are needed for generating coherent detection information, interpolation techniques can be used. The easiest interpolation method is to use linear interpolation. For example, assuming that k = 7n + i where 1 ≤ i ≤ 6, results in (eq. 2) being rewritten as:

25

$$\bar{h}(kT) = \frac{1}{7}\left[(7-i)\times\bar{h}(7nT) + i\times\bar{h}(7nT+7T)\right]$$

$$= \frac{1}{7}\left\{(7-i)\times\bar{h}(nT_r) + i\times\bar{h}[(n+1)T_r]\right\} \qquad \text{(eq. 3)}.$$

Other more sophisticated interpolation techniques can be used to further improve the estimation. However, when the channel fading is

30   slow relative to the sampling rate $1/T_r$ of the reference signal, the linear interpolation method described above is sufficient.

It will be appreciated by those skilled in the art that the sampled channel response $h(n) \equiv h(nT_r)$ can be modeled as a slowly time-

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varying random process with a power spectrum $\Phi(f)$, and $\Phi(f) = 0$ for $f <$ $-f_d$ and $f > f_d$ where f_d is the Doppler frequency. If f_d s not known or may change with time, the best estimate $\bar{h}(n)$ which can obtained is to pass the noisy estimate $\hat{h}(kT_r)$ through an ideal low-pass linear-phase filter

5   which rejects the noise components with frequency $|f| > f_{d,max}$. The optimal filter would have a frequency response, i.e. the Fourier transform of w_k denoted as:

$$f(\omega) \begin{cases} 1 & -f_{d,max} \leq f \leq f_{d,max} \\ 0 & \text{elsewhere} \end{cases} \qquad \text{(eq. 4)}.$$

10

One such filter is finite impulse response (FIR) filter which has an output which can be written as:

$$\bar{h}(n) = \bar{h}(nT_r) = \sum_{k=-N}^{N} w_k \hat{h}(nT_r + kT_r) \qquad \text{(eq. 5)}$$

15

where w_k is the FIR filter coefficients, or weights, for generating the estimate. It is easy to see that the $(2N+1)$ sample average method described above is a specially case of this weighted sum method, if $w_k = 1/(2N+1)$. By selecting w_k according to the criterion described

20   above, a better estimate of $h(n)$ can be obtained, although delay will be introduced. FIG. 3 illustrates how (eq. 5) may be implemented to generate a channel estimate using weighed average of noise channel estimated based on the reference signal.

It will be appreciated by those skilled in the art that the

25   performance of an ideal coherent receiver over a Rayleigh fading channel is well known. Therefore, the following discussion will analyze the performance loss of the reference-based channel estimation method relative to the optimal coherent receiver over such a channel.

The loss is mainly caused by two factors. First, because of the

30   insertion of non-information bearing reference bits, the energy per information bit (E_b) is effectively reduced (i.e., the data rate is reduced), when the total transmitted power remains the same. Second, in the presence of noise, there are errors in the channel estimates. The total loss is the combined result of these two factors.

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The loss in E_b due to reference insertion, denoted by ζ_i, can be simply expressed as:

$$\zeta_i = (M+1)/M = 1 + 1/M \qquad \text{(eq. 6)}$$

5

where $1/M$ is the insertion rate. For example, for $M = 6$, a loss in E_b of $10\log_{10}(7/6) \approx 0.67$ dB occurs.

In order to analyze the performance loss due to estimation error, a channel and signaling model must first be established for the DS-

10   CDMA up-link. A DS-CDMA up-link can be viewed as multiple flat fading channels. For each of the flat fading channels, the received signal after despreading can be expressed as:

$$r(k) = h(k)a(k) + z(k) \qquad \text{(eq. 7)}$$

15

where $r(k)$ is the received sample at kT, $a(k)$ is the corresponding transmitted data or reference symbol, $h(k)$ is a low-pass random complex variable characterizing the fading channel, and $z(k)$ is the additive noise or interference, which is approximately white and

20   Gaussian. The average signal to noise ratio is equal to $E[|a(k)|^2]E[|h(k)|^2]/\sigma_z^2$, where σ_z^2 is the noise variance. According to Jakes' channel model (as described in W. C. Jakes, Ed., *Microwave Mobile Communications*, John Wiley, New York, 1974), $h(k)$ is unbiased and Gaussian distributed. Its power spectrum, called Doppler spectrum,

25   is given by:

$$H(f) = \frac{3}{2\pi f_d}\left[1 - \left(\frac{f}{f_d}\right)^2\right]^{-1/2} \qquad \text{(eq. 8)}$$

where f_d is the Doppler frequency, which is a function of the carrier

30   frequency and the mobile communication unit speed. It can be further assumed that BPSK signaling is used. It will be appreciated by those skilled in the art that the following assumptions also are applicable to quadrature phase shift keying signaling. Given that $|a(k)| = 1$, then it may be assumed that the reference symbols are inserted at $(M+1)kT$, so

35   that $a((M+1)k)$ and $r((M+1)k)$ are the reference symbols and the

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corresponding received samples, respectively. By multiplying r((M+1)k) with the conjugate of the reference symbol $a^*((M+1)k)$, the resulting channel estimate is:

5

$$\hat{h}((M+1)k) = h((M+1)k) + a^*((M+1)k)z((M+1)k). \qquad \text{(eq. 9)}.$$

In the channel estimate given by (eq. 9), $h((M+1)k)$, is a low-pass random variable and the second term is the channel estimation error. When $1/(M+1)T > 2f_d$, the channel estimate at $(M+1)kT$ can be further

10   improved, i.e., the variance of the channel estimation error can be reduced. It is known from Wiener filtering theory that the optimal unbiased channel estimate, in the sense of maximizing the ratio of signal energy to the variance of estimation error, can be obtained by passing $\hat{h}((M+1)k)$ through a linear-phase filter whose magnitude

15   response is equal to the square-root of the quotient of H(f) divided by the noise spectrum. The optimal channel estimator based on the inserted reference symbols is indeed a linear phase matched filter, which is matched to the channel Doppler spectrum divided by the noise spectrum.

20   Practically, it is difficult to implement such an optimal estimator, because the Doppler and noise spectra are usually not known and will change with time. A sub-optimum and realistic solution is to use a fixed, linear-phase, low-pass filter, whose cut-off frequency is greater than or equal to the maximum possible Doppler frequency.

25   By filtering $\hat{h}((M+1)k)$ using an ideal low-pass linear phase filter, whose cut-off frequency is $f_{cut-off}$, with zero group delay, or equivalently a fixed group delay, and unit magnitude in its passband, the output at the filter can be expressed as:

30

$$\bar{h}((M+1)k) = h((M+1)k) + \bar{z}((M+1)k) \qquad \text{(eq. 10)}.$$

It can be shown that the variance of $\bar{z}((M+1)k)$, the residual estimation error, is equal to $\sigma_z^2 \times (2f_{cut-off} / f_r)$, where $f_r = 1/(M+1)T$ is the reference symbol insertion frequency. More precisely, after filtering, the variance

35   of the channel estimation error is reduced by a factor of $2f_{cut-off}/f_r$. Subsequently, by using linear interpolation, the channel estimates at kT,

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for k ≠ (M+1)k can be generated according to (eq. 3). From (eq. 7) and (eq. 10) the coherently detected samples can be written as:

$$\hat{a}(n) = r(k)\tilde{h}^*(k)$$
$$= |h(k)|^2 a(k) + \tilde{z}^*(k)h(k)a(k) + z(k)h^*(k) + z(k)\tilde{z}^*(k) \qquad \text{(eq. 11)}$$

5

If it is further assumed that $\tilde{z}^*(k)h(k)a(k)$ and $z(k)h^*(k)$ are independent and the high order error term $z(k)\tilde{z}^*(k)$ is negligible, then the total noise variance at the output of the detector is equal to $|h(k)|^2 (\sigma_z^2 + \sigma_{\tilde{z}}^2)$. By comparing this result with the optimum coherent receiver for max-ratio

10   combining, which has an noise variance of $|h(k)|^2 \sigma_z^2$ at the detector output, it may be concluded that the loss due to estimation error is approximately equal to:

$$\zeta_{est} \approx 1 + \sigma_{\tilde{z}}^2/\sigma_z^2 = 1 + (2f_{cut-off}/f_r) \qquad \text{(eq. 12).}$$

15

By combining (eq. 6) and (eq. 12), the total performance loss may be expressed as:

$$\zeta_{total} \approx \left[1 + (2f_{cut-off}/f_r)\right] \times (1 + 1/M) \qquad \text{(eq. 13).}$$

20

The system performance may be optimized by selecting a proper M to minimize ζ_{total}.

For an r = 1/3 convolutionally coded system with information bit rate of 9600 bits per second, the coded bit rate is equal to 28800 bits per

25   second. Letting M = 6, then f_r = 4800 Hertz (Hz). For an $f_{cut-off}$ = 300 Hz, the total loss is (1+1/8)×(1+1/6) = 63/48 or about 1.14 dB. Since an optimal coherent receiver requires an E_b/N_0 over 3 dB less than that required by a non-coherent receiver, a gain of over 2 dB can be expected.

30   When diversity combining of signals from multiple RAKE fingers and/or multiple antennas is used, the difference in performance between the coherent receiver and non-coherent receiver could be even larger because the combining loss in a non-coherent receiver does not exist for a coherent receiver. Although this additional

35   advantage may be partially canceled out due to the fact that, when

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diversity combining is used, the communication system is likely to be
operated at lower signal-to-noise ratios for the individual signals to be
combined than without diversity. As a result, the second order term in
(eq. 11) can not be ignored.

5   In the above analysis. it has been assumed that an ideal low-
pass filter is used for channel estimation. Realization of such an ideal
filter will require an infinite delay. A practical filter must have a transition
band to have a finite delay. Thus, the allowable maximum Doppler
frequency should be less than the effective bandwidth of the filter. For

10   example, a filter may be designed which has a transition band from 200
Hertz to 400 Hertz. The resulting filter will have a delay less than five
millisecond while keeping the effective bandwidth equal to 300 Hertz.
By using such a filter, there is no additional loss if the mobile
communication unit speed is less than 220 kilometers per hour.

15   Furthermore in the above analysis. it was assumed that the data
and reference symbols are transmitted continuously. As a result, the
reference signals are available every T_r time interval and these
reference signals can be used for channel estimation by means of low-
pass filtering. In some cases, such as when variable rate speech

20   transmission is used, it is desirable to transmit data in short and
discontinuous blocks. In such a case, the referenced symbols are also
transmitted discontinuously and the low pass filtering method described
above should be modified to be applicable as described below.

First, it will be appreciated by those skilled in the art that when the
25   duration of a data block is short relative to the time constant of channel
variation, the channel response may be assumed to vary linearly during
the time span of a block. Thus, the channel response h(kT) may be
expressed as:

30          $$h(kT) = \alpha + \beta k \qquad \text{(eq. 14)}$$

where α and β are two complex constants to be estimated. These two
constants may be determined by using the received reference samples
through linear best-fitting based on the least squares (LS) principle.
35   The details of such estimation methods are illustrated by the following
example.

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If a transmitted data block (i.e., frame) is assumed to consists of 36 data symbols and with 6 inserted reference symbols (see FIG. 2). As a result, 42 symbols are transmitted per data block. This data block may be separated by time intervals in which no data are transmitted. Thus,

5   when such a short block is received, we only have 6 reference samples during the time interval of interest.

In this example, the channel estimation is performed by using only the received reference samples. The received samples may be denoted by r(k), k = 0, ..., 41, which can be written as:

10

$$r(k) = h(kT)a(k) + z(k) \qquad (eq. 15)$$

where a(k) is the kth transmitted symbol, which can be either a data symbol (not known to the receiver) or a reference symbol (known to the

15   receiver), and z(k) is the additive noise at k. In this example, for a reference sample r(k), k=7i+3, i = 0, 1, 2, 3, 4, and 5, since a(k) is known, a noisy channel estimate can be obtained as:

$$\hat{h}(kT) = r(k)a^*(k) \qquad (eq. 16)$$

20

where $a^*(k)$ represents the complex conjugate of a(k).   By minimizing the LS error between h(kT) given by (eq. 14) and $\hat{h}(kT)$, the LS estimates of α and β may satisfy the following:

25     $$\alpha N + \beta \sum_k k = \sum_k \hat{h}(k) \qquad (eq. 17)$$

$$\alpha \sum_k k + \beta \sum_k k^2 = \sum_k k\hat{h}(k) \qquad (eq. 18)$$

where N is the number of elements in the summation and $\hat{h}(kT)$ is given by (eq. 16) and the summation index k takes the value k = 3+7i, i = 0, 1,

30   ..., 5. The solution of α and β is given by:

$$\alpha = d\left[\sum_k k^2 \times \sum_k \hat{h}(k) - N \sum_k k\hat{h}(k) \right] \qquad (eq. 19)$$

$$\beta = d\left[N \sum_k \hat{h}(k) - \sum_k k \times \sum_k \hat{h}(k) \right] \qquad (eq. 20)$$

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where,

$$d = \sqrt{\left[N \sum_k k^2 - \left(\sum_k k \right)^2 \right]}$$ (eq. 21).

5

Thus, the estimates channel responses at time kT, k = 0, 1, ..., 41 can be computed according to (eq. 14) using the estimated α and β.

Although the invention has been described and illustrated with a certain degree of particularity, it is understood that the present

10   disclosure of embodiments has been made by way of example only and that numerous changes in the arrangement and combination of parts as well as steps may be resorted to by those skilled in the art without departing from the spirit and scope of the invention as claimed. For example, the modulator, antennas and demodulator portions of the

15   preferred embodiment communication system as described were directed to CDMA spread-spectrum signals transmitted over a radio communication channel. However, as will be understood by those skilled in the art, the encoding and decoding techniques described and claimed herein can also be adapted for use in other types of

20   transmission systems like those based on time division multiple access (TDMA) and frequency division multiple access (FDMA). In addition the communication channel could alternatively be an electronic data bus, wireline, optical fiber link, satellite link, or any other type of communication channel.

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Claims

What is claimed is:

5   1.   A communication unit comprising:
          (a)   reference means for inserting reference symbols into a
                 stream of input data symbols to form a reference coded
                 stream of input data symbols; and
          (b)   spreading means, operatively coupled to the reference
10   means, for preparing the reference coded stream of input
                 data symbols for transmission over a communication
                 channel by spreading the reference coded stream of input
                 data symbols with a spreading code prior to transmission
                 over the communication channel.

15

     2.   The communication unit of claim 1 wherein the spreading means
           comprises first spreading means for spreading the reference
           symbols at a first rate of spreading and second spreading means
           for spreading the stream of input data symbols at a second rate of
20   spreading.

     3.   A communication unit comprising:
          (a)   a demodulator adapted for despreading a received
                 communication signal with a spreading code to derive a
25   stream of reference samples and a stream of data
                 samples;
          (b)   a channel estimator, operatively coupled to the
                 demodulator, and adapted for estimating the channel
                 response by utilizing the stream of reference samples; and
30   (c)   a detector, operatively coupled to the demodulator and the
                 channel estimator, and adapted for generating an
                 estimated data symbol from the stream of data samples by
                 utilizing the estimated channel response.

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4.   The communication unit of claim 3 wherein the demodulator
      comprises a first despreader adapted for despreading the
      received communication signal with one of a first despreading
      code and a first rate of despreading, respectively, to derive the
5   stream of reference samples, and a second despreader adapted
      for despreading the received communication signal with one of a
      second despreading code and a second rate of despreading,
      respectively, to derive the stream of data samples.

10   5.   The communication unit of claim 3 wherein the channel estimator
      comprises one of the group consisting of:
            (a)   means for estimating the channel response by low-pass
                   filtering the stream of reference samples;
            (b)   means for estimating the channel response as a linear
15   function of sampling time; and
            (c)   filtering means for generating an estimate of the channel
                   response associated with each reference sample by low-
                   pass filtering the stream of reference samples, and
                   interpolating means, operatively coupled to the filtering
20   means, for generating an estimated channel response
                   between at least two of the reference sample estimated
                   channel responses.

      6.   The communication unit of claim 3 wherein the detector
25   comprises one of the group consisting of:
            (a)   means for generating the estimated data symbol from the
                   stream of data samples by correlating the estimated
                   channel response with the stream of data samples;
            (b)   means for generating an estimated data bit by utilizing
30   maximum likelihood decoding techniques to derive the
                   estimated data bit from the estimated data symbol.
            (c)   means for generating an estimated data bit by utilizing a
                   Viterbi maximum likelihood decoding algorithm to derive
                   the estimated data bit from the estimated data symbol.

35

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7.   A method of communication comprising:

(a)   inserting reference symbols into a stream of input data symbols to form a reference coded stream of input data symbols;

5   (b)   preparing the reference coded stream of input data symbols for transmission over a communication channel by spreading the reference coded stream of input data symbols with a spreading code prior to transmission over the communication channel; and

10   (c)   transmitting the spread reference coded stream of input data symbols over the communication channel.

8.   The method of claim 7 wherein the step of inserting comprises inserting reference symbols into a stream of input data symbols

15   according to an insertion algorithm to form a reference coded stream of input data symbols, the insertion algorithm comprising inserting reference symbols at a rate greater than twice of channel variation frequency of a communication channel over which the spread reference coded stream of input data symbols is

20   to be transmitted.

9.   A method of processing a received communication signal comprising:

(a)   despreading the received communication signal with a

25   spreading code to derive a stream of reference samples and a stream of data samples;

(b)   estimating the channel response by utilizing the stream of reference samples; and

(c)   generating an estimated data symbol from the stream of

30   data samples by utilizing the estimated channel response.

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10. The method of claim 39 wherein the estimating step comprises:

(a) generating an estimate of the channel response associated with each reference sample by low-pass filtering the stream of reference samples; and

5

(b) generating an estimated channel response between at least two of the reference sample estimated channel responses.
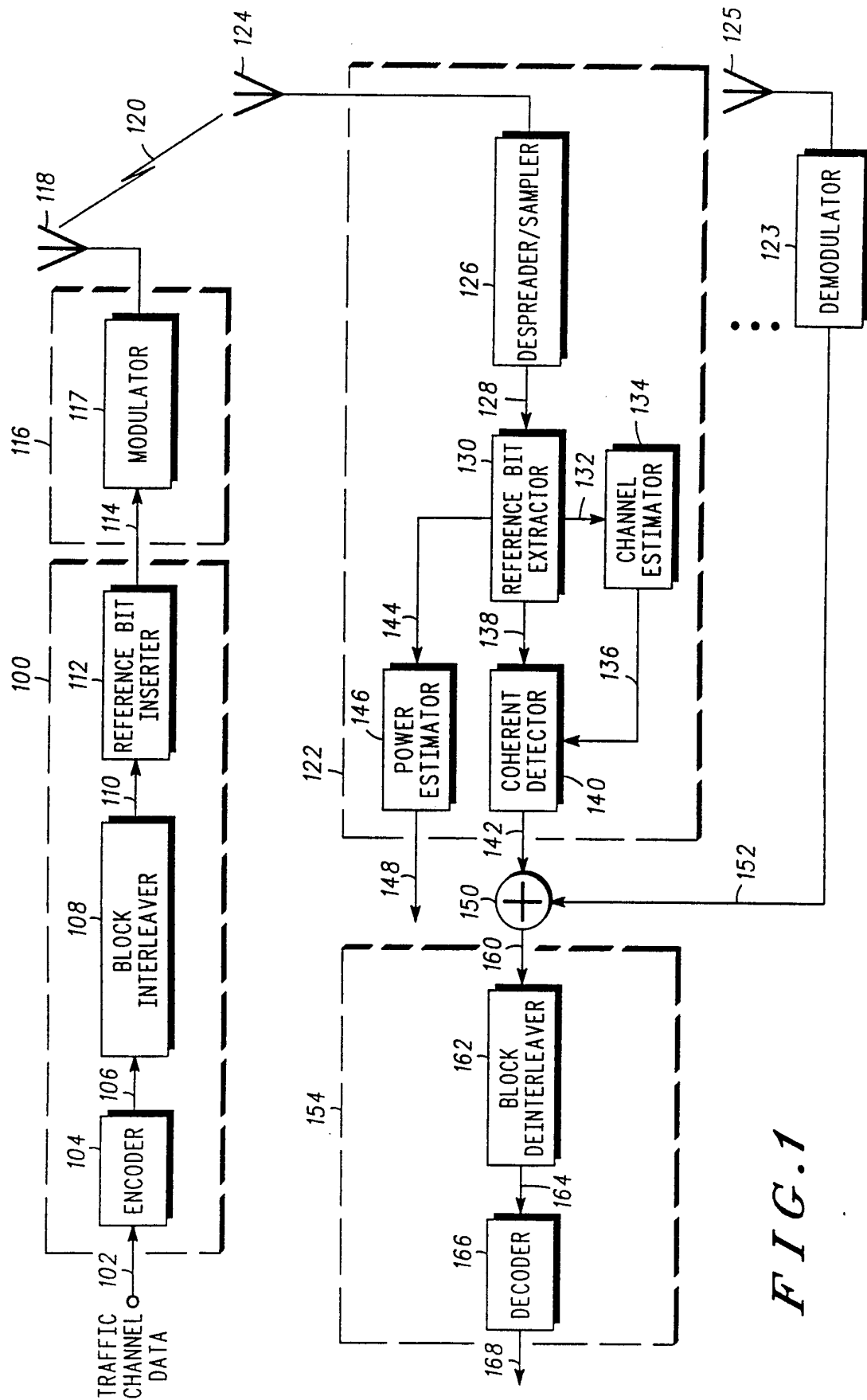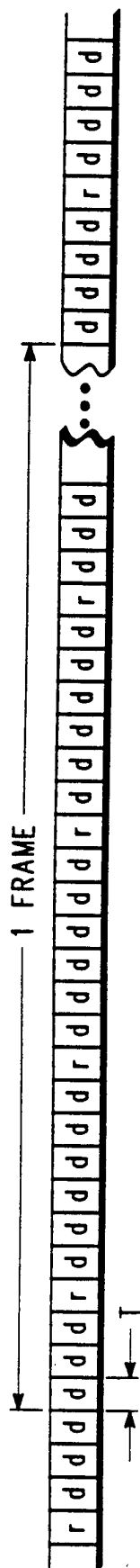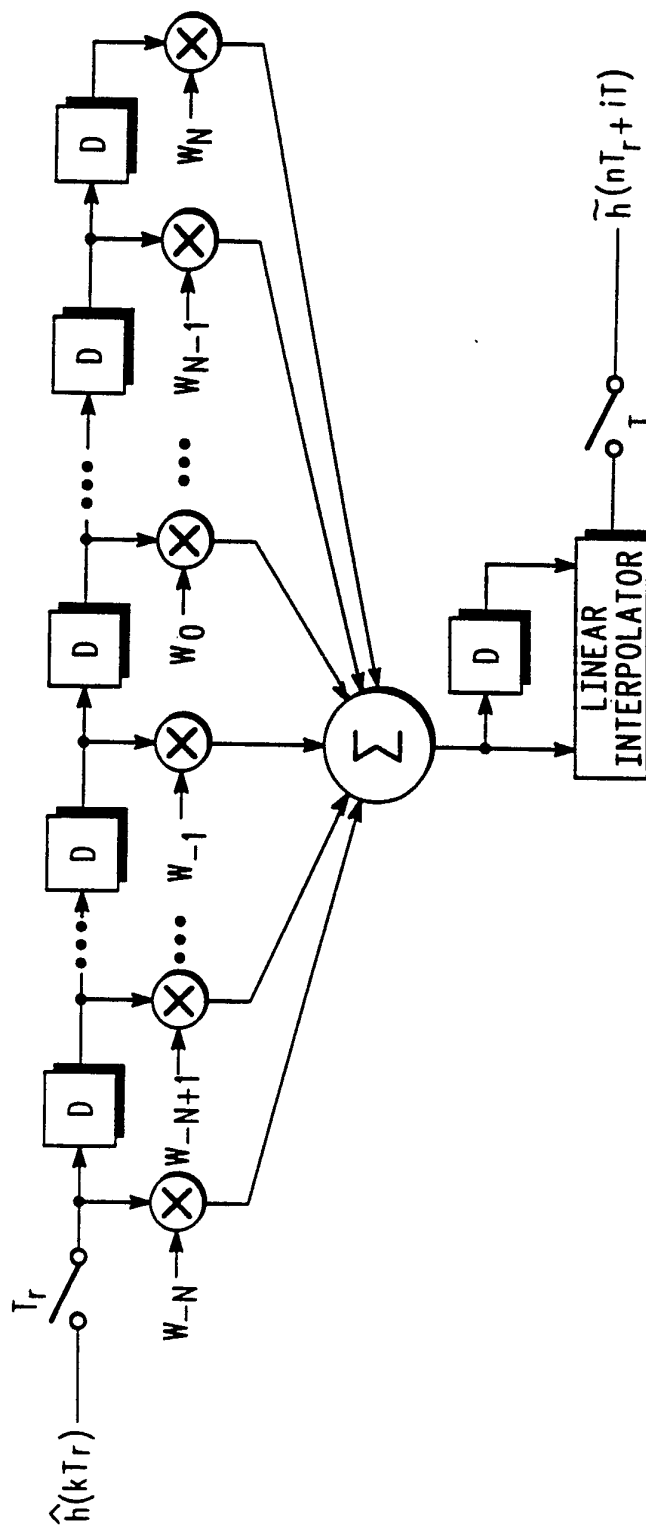
F I G . 1

F I G . 2

F I G . 3

A. CLASSIFICATION OF SUBJECT MATTER

IPC(5)   :H04K 1/00
US CL   :375/1; 380/42, 33, 34

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)

U.S. : 375/1; 380/42, 33, 34

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

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

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US, A, 5,181,225 (NEESER ET AL.) 19 January 1993	1-10
A	US, A, 4,559,633 (KAN ET AL.) 17 December 1985	1-10
A	US, A, 4,730,340 (FRAZIER, JR.) 08 March 1988	1-10
A, P	US, A, 5,214,669 (ZAREMBOWITCH) 25 May 1993	1-10
A	US, A, 5,029,184 (ANDREN ET AL.) 02 July 1991	1-10

☐ Further documents are listed in the continuation of Box C.   ☐ See patent family annex.

*	Special categories of cited documents:	"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
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Date of the actual completion of the international search	Date of mailing of the international search report
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Facsimile No. (703) 305-3230	Telephone No. (703) 308-0463

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