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COHERENT MODULATION OF CPM SIGNALS

Field of the Invention

The present invention relates generally to the field of digital communications. More particularly, the present invention relates to a method and system for coherently modulating and demodulating a digital communication signal.

Background of the Invention

Conventional continuous phase modulation (CPM) schemes inherently employ differential encoding to implement the continuous phase property of the modulation scheme. Such a scheme is described, for example, in T. Aulin, J.B. Anderson and C.E.W. Sundberg, *Digital Phase Modulation, Applications of Communications Theory*, Plenum Press, New York, New York, 1986. CPM signals are typically demodulated using a coherent demodulator and then differentially detected to obtain the data symbols. One drawback of the differential detection process is that an error in each demodulated symbol output by the demodulator results in errors in two output data symbols. This error rate is double the error rate achieved by coherent demodulation of a coherently modulated signal.

On a Gaussian channel, this doubling of the symbol error rate does not cause a significant loss in performance. However, over a fading channel, the bit error rate (BER) varies linearly with the signal to noise ratio (SNR); thus, the doubling of the symbol error rate implies a 3 dB performance loss. To avoid this loss, a coherently encoded CPM scheme is desirable.

For most applications requiring transmission of digital data over nonlinear and/or fading channels, constant-envelope digital modulation is desirable. Because of their simplicity, constant-envelope modulation



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schemes such as phase-shift keying (PSK) and frequency-shift keying (FSK) have often been used. The continuing need for better spectral utilization has led to the search for more bandwidth-efficient modulation schemes. It is known that the bandwidth of constant-envelope digital modulation schemes can be reduced by smoothing the variations of the information-carrying phase using an analog filter. Such a modulation scheme is called continuous phase modulation (CPM).

10 In addition to improved spectral efficiency, CPM schemes provide a coding gain over PSK schemes. The coding gain is due to the memory that is introduced by the phase-shaping analog filter, which can be used by a trellis decoder to decode the transmitted data symbols.

15 A coherently modulated CPM system is described, for example, in Rimoldi, "A Decomposition Approach to CPM", *IEEE Transactions on Information Theory*, 34(2): 260-270 (1988). Rimoldi discloses a CPM system including a linear, time-invariant continuous phase encoder and a linear, time-invariant memoryless modulator. The time-invariant tilted phase trellis disclosed by Rimoldi results in a memoryless modulator of reduced complexity. The continuous phase encoder and the decoder are also simpler than for prior MSK systems, and the bit error probability for practical signal-to-noise ratios (SNRs) is reduced by a factor of two. However, the Rimoldi system teaches a complex implementation, and further does not teach how to convert a conventional CPM scheme into a constant envelope modulation scheme in which the data symbols are coherently encoded.

30 CPM schemes using Nyquist-3 pulse shaping have been suggested for use in mobile communications systems in, for example, Sayar and Pasupathy, "Nyquist 3 Pulse Shaping in Continuous Phase Modulation", *IEEE Transactions on Communications*, COM-35(1): 57-67 (1987). Sayar et al.

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disclose how Nyquist-3 pulse shaping in CPM schemes provides generally improved spectral qualities. However, Sayar et al. do not disclose how Nyquist-3 pulse shaping can be practically implemented for coherently modulated data symbols. Further, Sayar et al. do not recognize the performance limitations inherent in practical applications of Nyquist-3 pulse shaping with conventional receiver filters.

It would be desirable to adapt a continuous phase modulation scheme to perform coherent modulation and coherent encoding of data symbols to maintain a low error rate over a fading channel.

It would further be desirable for such a CPM signal to be demodulated using a simple demodulator.

15 Summary of the Invention

The present invention overcomes the above-noted problems, and provides other advantages, by providing a method for coherently modulating CPM signals. According to exemplary embodiments of the present invention, digital information in the form of data symbols is encoded; the encoded information is passed through a coherent CPM modulator. The receiver samples the transmitted information signal at symbol intervals or other predetermined intervals, and demodulates and decodes the sampled information.

Brief Description of the Drawings

A more complete understanding of the features, operation, and advantages of the present invention can be obtained by reading the following Detailed Description of the Preferred Embodiments in conjunction with the accompanying drawings in which like reference indicia indicate like elements, and in which:

FIG. 1 is a block diagram of an exemplary



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communication system according to an embodiment of the present invention;

FIGs. 2A-B are flow charts showing alternative exemplary communication methods according to the present invention;

FIGs. 3A-B are a plot of an exemplary phase pulse for use in the system and method of the present invention and a graphical comparison of the spectra of full-response and Nyquist-3 CPM schemes, respectively;

FIGs. 4A-B are a scatter plot and eye diagram, respectively, of received symbols using a narrowband IF filter; and

FIG. 5 is a scatter plot of received symbols using a compensation filter according to an embodiment of the present invention.

Detailed Description of the Preferred Embodiments

A CPM signal is characterized by a frequency pulse $g(t)$, or by a corresponding phase pulse $f(t)$. The relationship between the two is as follows:

$$f(t) = \int_{-\infty}^t g(u) du \quad (1)$$

Typically, $g(t)$ is time-limited to some interval $(0, LT)$. In addition, the normalization on $f(t)$ is that

$$f(\infty) = \frac{1}{2} \quad (2)$$

In terms of a sequence of data symbols a_k , and a modulation index h , a transmitted CPM signal is given by

$$s(t) = e^{j2\pi h \int_{-\infty}^t \sum_{k=-\infty}^{\infty} a_k g(t-kT) dt} \quad (3)$$

(4)

In terms of $f(t)$, equation (3) can be rewritten as
The constant envelope property of a CPM scheme is evident from the above equation. It will be appreciated that



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$$s(t) = e^{j2\pi h \sum_{k=-\infty}^{\infty} a_k f(t-kT)}$$

equations (3) and (4) employ inherent differential decoding in CPM schemes.

The frequency pulse $g(t)$ is non-zero only over the interval $(0, LT)$, where T is the symbol period. This
5 implies that

$$f(t) = \begin{cases} 0, & t \leq 0 \\ 0.5 & t \geq LT. \end{cases} \quad (5)$$

Using the above information, equation (4) can be rewritten for the transmitted CPM signal as follows. For $mT \leq t \leq (m+1)T$,

$$s(t) = e^{j2\pi h \left(\sum_{k=-\infty}^{m-L} a_k \frac{1}{2} \cdot \sum_{k=1}^L a_{m-k+1} f(t-kT) \right)} \quad (6)$$

10 where

$$\tau = t - mT. \quad (7)$$

For example, in a 4-level CPM scheme with modulation index 0.25, the symbol sequence $a_k \in \{-3, -1, 1, 3\}$, and the
15 encoded symbol sequence $\hat{a}_k$ can be defined such that $\hat{a}_k \in \{0, 1, 2, 3\}$. The relationship between symbols a_k and encoded symbols $\hat{a}_k$ can be defined as:

$$a_k = 2\hat{a}_k - 3. \quad (8)$$

Thus, the CPM signal in this example is given by

$$s(t) = e^{j\frac{\pi}{2} \left(\sum_{k=-\infty}^{m-L} (2\hat{a}_k - 3) \frac{1}{2} \cdot \sum_{k=1}^L (2\hat{a}_{m-k+1} - 3) f(t-kT) \right)} \quad (9)$$

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$$= e^{j\frac{\pi}{2} \left(\sum_{k=-\infty}^{m-L} \hat{a}_k \cdot \sum_{k=1}^L 2\hat{a}_{m-k+1} f(t-kT) \right)} e^{j\frac{\pi}{2} \left(\sum_{k=-\infty}^{m-L} \frac{-3}{2} \cdot \sum_{k=1}^L (-3) f(t-kT) \right)} \quad (10)$$

From equation (10), it is evident that the modulated signal can be separated into a



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data independent term $s_i(t)$ given by

$$s_i(t) = e^{j\frac{\pi}{2} \left(\sum_{k=-\infty}^{n-L} \frac{-3}{2} \cdot \sum_{k=1}^L (-3) A_r \cdot LT - kT \right)}$$
(11)

and a data-dependent term $s_d(t)$ given by

$$s_d(t) = e^{j\frac{\pi}{2} \left(\sum_{k=-\infty}^{n-L} 4 \cdot \sum_{k=1}^L 2^{d_{n-L-k}} A_r \cdot LT - kT \right)}$$
(12)

- 5 It will be appreciated that the first summation in the expression for $s_d(t)$ in equation (12) is only unique modulo 4, by virtue of the complex exponentiation operation.

Now, consider a second symbol sequence b_k , which is defined by

$$b_k = \left(\sum_{i=-\infty}^k \hat{a}_i \right) \bmod 4$$
(13)

- 10 where mod N means the remainder divided by N. It will be appreciated that N may be represented by 2^q in a general case.

In this example, since

$$b_k \in \{0, 1, 2, 3\}$$
(14)

- 15 and

$$\hat{a}_k = (b_k - b_{k-1}) \bmod 4$$
(15)

then, in terms of the second symbol sequence b_k , the expression for $s_d(t)$ can be rewritten as

$$s_d(t) = e^{j\frac{\pi}{2} \left(b_{n-L} \cdot \sum_{k=1}^L 2^{(b_{n-L-k} - b_{n-L-k-1}) \bmod 4} \right) f(r \cdot LT - kT)}$$

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The expression for $s_i(t)$ remains unchanged.

A conventional coherent demodulator detects the first

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symbol sequence a_k by estimating the second symbol sequence b_k , and then performing the differential detection operation shown in equation (15) to determine the first sequence a_k . According to the present invention, the
5 second symbol sequence b_k includes all of the information carrying symbols, and differential detection is unnecessary. Such a b_k signal is therefore a coherently modulated CPM signal.

Referring now to FIG. 1, a block diagram of an
10 exemplary communication system according to an embodiment of the present invention is shown. The system includes a transmitter having an encoder 10 for encoding a digital information signal containing a number of data symbols, and a modulator 14 for modulating the data symbols and
15 transmitting the filtered and encoded digital signal. Modulator 14 includes a filter for pulse-shaping. The system further includes a receiver 16 for receiving the transmitted signal. The receiver 16 includes a decoder 18 for decoding the transmitted signal. The decoder 18 can
20 be implemented by a Viterbi decoder modeling a finite state machine formulated according to equation (16), or can include any other suitable decoder. The receiver 16 can also include a filter (not shown), such as a narrowband IF receive filter or multiple-tap compensation
25 filter as will be described in more detail below, for filtering the received transmitted signal.

Referring now to FIGS. 2A-B, flow charts describing two alternative exemplary communications methods for transmitting data communication signals from a transmitter
30 to a receiver according to the present invention are shown. Referring first to FIG. 2A, a transmitter obtains data bits from a source in step 100. In step 102, the transmitter maps sets of q data bits into data symbols b_k , where the symbols b_k have one of the values $\{0, 1, \dots, 2^q - 1\}$
35 and k is a time index. It will be appreciated that the



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data bits may be passed through a channel encoder prior to the mapping performed in step 102.

In step 104, the transmitter generates a data-dependent signal of the form

$$s_d(t) = e^{j2\pi h \left(b_{-L} + \sum_{k=1}^L 2^{(b_{-L-k} - b_{-L-k-1} \bmod 2^L)} f(r+LT-kT) \right)}$$

5 where h is the modulation index of the form $K/2^q$ for an integer K and $f(r)$ is a normalized phase pulse.

In step 106, the transmitter generates a data-independent signal of the form

$$s_i(t) = e^{j2\pi h \left(\sum_{k=-\infty}^{m-L} \frac{-2^{(k-1)}}{2} \cdot \sum_{k=1}^L (-2^{(k-1)}) f(r+LT-kT) \right)}$$

10 In step 108, the transmitter generates a product signal which is the product of the data-dependent signal and the data-independent signal. In step 110, the product signal is transmitted over a communication channel from the transmitter to the receiver.

15 In step 112, the receiver receives the transmitted product signal and demodulates and/or decodes the received signal by estimating the data symbols b_k from the received signal and determining the data bits from the estimation of the data symbols. The receiver preferably detects the phase angle of the received signals at predetermined
20 sampling intervals. For the case where the source bits have been passed through a channel encoder prior to the mapping performed in step 102, the received signal is passed through a demodulator which outputs soft information representative of the encoded source bits
25 generated by the channel encoder, and the soft information is passed through a channel decoder to estimate the source bits.

Referring now to FIG. 2B, an alternative communication method according to the present invention is
30 shown. Steps 100 and 102 are the same as in the method

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shown in FIG. 2A. The method of FIG. 2B differs from that of FIG. 2A in step 103, in which the transmitter generates data symbols $\hat{a}_k$ from symbols b_k generated in step 102, where

$$\hat{a}_k = (b_k - b_{k-1}) \bmod 2^q$$

5 and then generates a_k where

$$a_k = 2\hat{a}_k - (2^q - 1)$$

In step 105, the data symbols a_k are passed through a CPM modulator at the transmitter to form a continuous phase modulated data signal. The CPM modulator includes a
10 filter having a phase pulse $f(\tau)$.

In step 107, the transmitter transmits the continuous phase modulated signal over a communication channel, and the transmitted signal is received, sampled, and demodulated and/or decoded in step 112 as described above
15 with respect to FIG. 2A.

It will be appreciated that other suitable methods incorporating the principles of the present invention will be readily apparent to those skilled in the art.

The decoder/demodulator 18 for the coherent CPM can
20 be, for example, a Viterbi decoder modeling a finite state machine for the CPM signal. An exemplary state machine can be formulated using the description in equation (16). A state can be represented by the L-tuple set of values $\{b_{m-L}, \dots, b_{m-1}\}$. An input of b_m to the finite state machine
25 in the above state causes a transition to the state $\{b_{m-L}, \dots, b_{m-1}\}$, while emitting an output signal given by equations (16) and (11). The number of states is equal to 4^L in a 4-level CPM scheme.

A full response CPM scheme is one in which the
30 frequency pulse $g(t)$ has support only in the range $(0, T)$. In such a scheme, the full response due to the modulating symbol can be achieved in one symbol interval. According



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to an exemplary embodiment of the present invention, Nyquist-3 CPM scheme can be implemented by using a Nyquist-3 filter as filter 12 to use a Nyquist-3 pulse as the frequency pulse. A Nyquist-3 pulse has support over
 5 multiple symbol intervals; however, if the CPM signal is observed at symbol intervals, the behavior is similar to a full response CPM signal.

A Nyquist-3 phase pulse is defined as

$$\begin{aligned} f(kT) &= 0, \quad k \leq n \\ &= 0.5, \quad k \geq n+1. \end{aligned} \quad (17)$$

Such a pulse can be generated by passing a pulse with zero ISI (i.e., a Nyquist pulse) through a filter that has a frequency response substantially equal to the reciprocal of the sinc function. Referring now to FIG. 3A, a plot of
 15 a Nyquist-3 phase pulse is shown as an example. Although a full response pulse satisfies equation (19), a Nyquist-3 CPM scheme provides improved spectral qualities over a full response CPM scheme. In particular, the sidelobe levels in the spectrum can be reduced considerably by
 20 using Nyquist-3 CPM instead of full-response CPM. Referring now to FIG. 3B, a comparison of the spectra of 4-level CPFSK, which uses a full-response rectangular pulse as the frequency pulse, with the spectra of a Nyquist-3 CPM scheme using the phase pulse of FIG. 3A is
 25 shown.

Mathematically, this can be shown by considering the equation for the data dependent part of the coherently modulated CPM signal, i.e.,

$$s_d(t) = e^{j\frac{\pi}{2} \left(b_{n-1} + \sum_{k=0}^L (b_{n+k} - b_{n+k-1} \bmod 2) \right) f_c(t-LT-kT)} \quad (18)$$

30 At integer multiples of the symbol period using, equation (19) in the above equation yields



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$$s_d((N+n)T) = e^{j\frac{\pi}{2}(b_{N+1} + \sum_{k=1}^L (b_{N+k} - b_{N+k-1} \bmod 4))}$$

(19)

or equivalently

$$s_d((N+n)T) = e^{j0.5\pi b_n}. \quad (20)$$

Thus, by removing the data-independent part from the
5 received signal, the signal can be demodulated by simply
observing it at the sampling points. This CPM demodulator
18 offers substantially reduced complexity.

If a narrowband receive IF filter is used in receiver
16, signal distortion can occur. This distortion is small
10 if the passband of the IF filter is wider than the band
containing most of the CPM signal energy, but can lead to
a degradation in performance, as illustrated by the
scatter plot of the received symbols in FIG. 4A, and the
eye diagram in FIG. 4B. The distortion causes a loss in
15 power with respect to PSK systems; however, this loss is
offset by the fact that the peak-to-average power of the
CPM signal is lower.

A compensation filter can be used for the receive
filter in receiver 16 to improve performance. For
20 example, a five tap compensation filter having taps at
half-symbol intervals can be used. The compensation
filter is preferably designed to minimize the squared
error between the input and the known output and thereby
substantially reduce the performance-degrading effects of
25 conventional receiver filters, as described above. In
practice, an algorithm such as the LMS algorithm or an
equivalent algorithm can be used to determine the taps of
the compensation filter. FIG. 5 shows scatter plots of
the received symbols, with and without the compensation
30 filter. It will be appreciated that the compensation
filter improves performance by bringing the outlying



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points closer to the actual constellation point.

While the foregoing description includes many details and specificities, it will be appreciated that these are by way of example, only, and are not to be construed as
5 limitations of the invention. Many modifications to the disclosed examples will be readily apparent to those of ordinary skill in the art which do not depart from the spirit and scope of the invention, as defined by the appended claims and their legal equivalents.

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CLAIMS:

1. A method for transmitting and receiving a digital communications signal, comprising the steps of:

obtaining data bits from a source;

5 mapping sets of q data bits into data symbols b_k having one of the 2^q values $\{0, 1, \dots, 2^q - 1\}$, where k is a time index;

generating a data dependent signal

$$s_d(t) = e^{j2\pi h \left(b_{n-L} + \sum_{k=1}^L 2^{(q-1-k)} (b_{n-L+k} - b_{n-L+k-1} \bmod 2^q) f(\tau + LT - kT) \right)}$$

10 where h is a modulation index of the form $K/2^q$ for an integer K and $f(\tau)$ is a normalized phase pulse;

generating a data independent signal

$$s_i(t) = e^{j2\pi h \left(\sum_{k=-\infty}^{m-L} \frac{-(2^q-1)}{2} + \sum_{k=1}^L (-(2^q-1)) f(\tau + LT - kT) \right)}$$

generating a product signal from the data-dependent signal and the data-independent signal;

15 transmitting the product signal over a communication channel;

receiving the transmitted product signal; and

demodulating the received signal by estimating the data symbols b_k from the received signal determining the data bits from the estimation of the data symbols.

20

2. The method of claim 1, wherein the phase pulse has a Nyquist-3 pulse shape.

3. The method of claim 1, further comprising the step of filtering the received signal through a compensation filter prior to the step of demodulating.

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4. The method of claim 1, wherein the step of





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demodulating includes sampling the phase angle of the received signal at predetermined sampling intervals.

5 5. The method of claim 3, wherein the compensation filter is a five-tap compensation filter having taps at half-symbol intervals.

6. The method of claim 5, wherein the compensation filter minimizes a mean-squared error value between the received signal and a predetermined signal known to the receiver.

10 7. The method of claim 6, wherein the compensation filter is a multiple-tap compensation filter having taps determined by an LMS algorithm at predetermined intervals.

15 8. The method of claim 1, further comprising the step of encoding the data bits from the source prior to the step of mapping, and wherein the step of demodulating is performed by passing the received signal through a demodulator to generate soft information representative of the encoded data bits and passing the soft information through a channel decoder to estimate the data bits from
20 the source.

9. A method for transmitting and receiving a digital communications signal, comprising the steps of:

obtaining data bits from a source;

25 mapping sets of q data bits into second data symbols b_k having one of the 2^q values $\{0, 1, \dots, 2^q - 1\}$, where k is a time index;

generating first data symbols $\hat{a}_k$, where

$$\hat{a}_k = (b_k - b_{k-1}) \bmod 2^q$$





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generating new data symbols a_k , where

$$a_k = 2\hat{a}_k - (2^q - 1)$$

passing the new data symbols a_k through a CPM modulator having a modulation index h and phase pulse $f(\tau)$, to form
5 a continuous phase modulation signal;

transmitting the continuous phase modulated signal over a communication channel;

receiving the transmitted signal; and

demodulating the received signal by estimating the
10 data symbols b_k from the received signal.

10. The method of claim 9, wherein the phase pulse has a Nyquist-3 pulse shape.

11. The method of claim 9, further comprising the step of filtering the received signal through a
15 compensation filter prior to the step of demodulating.

12. The method of claim 9, wherein the step of demodulating includes sampling the phase angle of the received signal at predetermined sampling intervals.

13. The method of claim 11, wherein the compensation
20 filter is a five-tap compensation filter having taps at half-symbol intervals.

14. The method of claim 13, wherein the compensation filter minimizes a mean-squared error value between the received signal and a predetermined signal known to the
25 receiver.

15. The method of claim 14, wherein the compensation filter is a multiple-tap compensation filter having taps determined by an LMS algorithm at predetermined intervals.

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- 5 16. The method of claim 9, further comprising the step of encoding the data bits from the source prior to the step of mapping, and wherein the step of demodulating is performed by passing the received signal through a demodulator to generate soft information representative of the encoded data bits and passing the soft information through a channel decoder to estimate the data bits from the source.

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