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AN EQUIVALENT MATCHED FILTER FOR
CONTINUOUS PHASE MODULATED SIGNALS.

University — Université

CARLETON

Degree for which thesis was presented — Grade pour lequel cette thèse fut présentée

DOCTOR OF PHILOSOPHY PH.D.

Year this degree conferred — Année d'obtention de ce grade

1984

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AN EQUIVALENT MATCHED FILTER
FOR CONTINUOUS PHASE MODULATED SIGNALS

by

Lincoln Geo. Morant, B.Sc., M.Eng.

A thesis submitted to
The Faculty of Graduate Studies and Research
in partial fulfillment of the requirements for
the degree of
Doctor of Philosophy

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March 1984

1001
Ph.D.
1984
M67

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"AN EQUIVALENT MATCHED FILTER FOR CONTINUOUS PHASE
MODULATED SIGNALS"

submitted by Lincoln Geo. Morant, B.Sc., M.Eng. in partial fulfillment
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ABSTRACT

The increasing need for the use of Partial Response Continuous Phase Modulated (CPM) Systems in data transmission necessitates the design of a receiver that is easy to implement yet has good performance capabilities.

This thesis presents the concepts, theoretical development and design of a new type of receiver termed, the Equivalent Matched Filter. This receiver uses the information contained within the zero crossing intervals for demodulation and requires only automatic frequency control (afc) as far as the carrier is concerned. The detection is achieved at IF. An important feature of the receiver is that, if the modulation index, h , is changed, there is no need to change any physical component of the circuit.

A Raised Cosine (RC) Partial Response CPM modem, with $h=0.8$ and a pulse length $L=3$, has been implemented and an experimental study conducted to determine the system's bandwidth and signal to noise ratio performance.

The receiver's performance is approximately 1.5 db from optimum and shows remarkable closeness to that predicted by theory for this type of receiver.

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ACKNOWLEDGEMENTS

I would like to thank my supervisor, Professor J. S. Wight, for his help, guidance and encouragement throughout this work.

I wish also to express my gratitude to Professor D. D. Falconer, Professor H. M. Hafez and Dr. M. El-Tanany for the enlightening discussions I had with them. The graduate students and staff of the Department of Electronics also deserve credit for making life tolerable. Special thanks goes to Technologist Nagui Mikhail for his practical suggestions and help. The efficient typing of Diane Dodds also deserves credit.

I thank also Carleton University for providing me with financial assistance throughout this work.

Finally, I would like to thank my wife, Barbara, for giving me the emotional support necessary to undertake this work.

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LIST OF PRINCIPAL SYMBOLS

A	Amplitude of hard-limited signal
D^2	Squared Euclidean Distance
E_B	Energy in one transmitted bit (Joule)
f_c	Carrier centre frequency (HZ)
$g(t)$	Impulse response of frequency pulse
h	Modulation index
L	Length of frequency pulse (bits)
N_0	White noise density (W/HZ)
N_{0L}	Noise density after hard-limiting
p	Number of phase states
$q(t)$	Phase pulse
$r(t)$	Received signal
$r'(t)$	Hard-limited receiver signal
$S(t)$	Hypothesized transmitted signal
$S'(t)$	Hard-limited hypothesized transmitted signal
T	Bit interval
t_n	n th zero crossing of received or transmitted signal
t'_n	Noisy zero crossing of received or transmitted signal
W	Pulse width
Z_N	Decision variable of matched filter
E_B/N_0	Energy per bit to noise density ratio $J/(WHZ^{-1})$
α_i	Bit
$\bar{\alpha}$	Sequence of bits

$\bar{\delta}$	Pattern
θ_N	Phase node
θ_N'	Phase state
$\theta(t)$	Time dependent part of phase information
τ_n	nth zero crossing interval
$\phi(t)$	information bearing signal phase
ω_c/ω_0	Radian carrier centre frequency (rad/sec.)

CHAPTER 1

INTRODUCTION

1.1 General Background

Digital transmission over radio channels has become very popular in recent years in the telephony industry, in the mobile radio community and in other industrial applications. However most of the commonly used modulation schemes which satisfy digital transmission over telephone lines or co-axial cables have limitations for radio applications. These schemes exhibit amplitude variations and are therefore incompatible with the radio systems that require constant amplitude. Quite naturally this has led to the use of frequency modulated signals. However while FM satisfies the constant amplitude requirement, there are other constraints for which the traditional FM signal has to be modified.

In general the chief requirements that a modulation scheme for digital transmission over radio channels must meet are:

- (a) Spectral efficiency: This is concerned with the amount of bandwidth to be allocated to the signal for a given data rate. In a system, bandwidth is normally limited, therefore the modulation scheme chosen should have most of its energy concentrated in a small band of frequencies around the carrier frequency. That is, the main lobe of its power spectrum should be narrow. Its out of band radiation should be small thus lessening the filtering requirement necessary to avoid interchannel interference.

- (b) Power efficiency: This is concerned with the amount of transmitted power needed to achieve a given error probability performance. Power efficiency is also related to the requirement that the signal envelope be constant so that non-linear amplifiers (commonly found in radio systems) do not cause distortion and spectral spreading of the signal.
- (c) Synchronization and detection: It is well known that a coherent detector provides the best error performance for a signal ~~disturbed~~ by additive white gaussian noise. However, the detectors require a reference signal which is phase coherent with the transmitted reference. These are usually generated from the received signal. For some modulation schemes, and in certain situations, this can be quite a difficult task and requires complicated carrier recovery circuitry.

Although a frequency modulated signal satisfies the constraint of constant envelope, its spectrum is relatively wide. Therefore, in order that it may be compatible with the radio environment the spectrum has to be shaped. One approach, which has gained popularity, is to shape the data at the input to the frequency modulator by means of a filter. This has given rise to a class of signals called Continuous Phase Modulated (CPM) signals. Within this class of signals is a sub-class referred to as Partial Response Continuous phase modulation. These signals can yield power spectra with narrow main lobes and low side lobes which makes them spectrally efficient. They also possess the quality of constant envelopes. However the optimum receiver for this signal consists of

a bank of matched filters followed by a viterbi processor. This combination can be quite complex. Furthermore the synchronization of the receiver for some of these schemes can be quite difficult, especially in a low signal to noise ratio (S/N) environment with fast synchronization requirements such as burst transmission applications. An example of this is the case of digital transmission over the voice channels of mobile radio during silent intervals, or in the case of recent requirement in which the U.S. army proposes to use a Time Division Multiple Access (TDMA) scheme for communication with its units [1].

If synchronization cannot be accomplished quickly in these cases much of the information will be lost. One can consider sending synchronization information separately but this tends to be considered a waste of power. Consequently circuitry which provides fast synchronization without the expense of complexity is very much in demand.

It should be pointed out that one of the virtues of CPM signals is their ability to provide good power efficiency once the decision interval is extended to more than one symbol interval.

A general partial response continuous phase modulator is shown in Figure 1.1. The system consists of a tapped delay line, with coefficients f_n , in cascade with a filter with frequency response $G(\omega)$. The system response up to the frequency modulator input is given in [2] as

$$H(\omega) = F(\omega) \times G(\omega) \quad (1.1)$$

where

Figure 1.1: General Partial Response CPM Modulator

$$F(\omega) = F(D) \Big|_{D=e^{j\omega T}} = \sum_{n=0}^{N-1} f_n e^{jn\omega T} \quad (1.2)$$

$$F(D) = \sum_{n=0}^{N-1} f_n D^n \quad (1.3)$$

D is the delay operator

T is the symbol interval and

N is the length of the delay line in terms of symbol intervals.

The coefficients f_n are chosen to force the desired sample values at specific sampling instants, and $G(\omega)$ is designed to preserve those samples values and to bandlimit the resulting spectrum.

It is well known that the spectrum efficiency increases with the order N of the system polynomial, $F(D)$, however, this increases the number of signal levels at the transversal filter output and the number of phase functions $\phi(t)$ in the output signal.

$$S(t) = \sqrt{\frac{2E}{T}} \cos(2\pi f_0 t + \phi(t) + \phi_0) \quad (1.4)$$

where

E is the symbol energy

T is the symbol time

f_0 is the carrier frequency,

$\phi(t)$ is the information bearing pulse, and

ϕ_0 is some arbitrary carrier starting phase.

Unfortunately the increased efficiency achieved by increasing the order of the polynomial is negated by the need for a more complex receiver.

It is worth noting that all the information is carried in the phase $\phi(t)$ of the signal. Therefore, at the receiver it is this phase that is of primary importance.

1.2 Historical Review

Frequency shift keying in which the phase is continuous at the symbol transition time is referred to as Continuous Phase Frequency Shift Keying (CPFSK) or simply as Continuous Phase Modulation (CPM). By making the deviation ratio, h , relatively small it has been found that the spectrum of the modulated carrier is fairly compact.

de Buda [3] has performed extensive work on one form of CPFSK called Fast Frequency Shift Keying (FFSK) with index $h=0.5$. He has shown it to have good spectral properties and to have S/N performance as good as QPSK. It is also very easy to demodulate. Several other authors [4] - [7] have dealt extensively with other aspects of this scheme.

It has been found that improvement in the spectral efficiency of CPFSK signals is achievable by merely smoothing the phase transitions. Aulin et al have shown that with various baseband smoothing filters, great improvement in the spectrum can be obtained. In their paper, "Full Response Signalling" [8], one data symbol controls the phase path over one symbol interval although for continuity the actual phase values are dependent upon previous symbols.

In a sequel to this paper, Aulin et al. [9], have shown that with Partial Response Signalling [10] - [12] the spectral performance of CPFSK can be further improved. Here more than one data symbol control the phase path over a symbol interval while again the actual instantaneous phase values depend upon previous data symbols to maintain the phase continuity. The amount of phase control each symbol exhibits depends upon the weighting function $g(t)$, (the baseband filter) referred to in the literature as the frequency pulse. Actually $g(t)$ is the inverse Fourier transform of $H(\omega)$ of equation (1.1).

Another class of signals appearing in the literature is called Correlative Phase Shift Keying. This is the same as Partial Response CPM signals with the stipulation that the phase must reach specified values at the end of each signalling interval [13]. Extensive work has been performed [14] - [18] on a very useful form of this type of signal, called Tamed Frequency Modulation (TFM).

It has been reported in [19] - [22] that by exploiting the correlation between successive phase shifts and examining the signal over more than one interval, before making a decision about a particular bit, signal to noise ratio performance better than BPSK is achievable.

In [8] and [25] a quantity called the Minimum Euclidean distance is defined and calculated for different CPM schemes. This quantity is shown to greatly influence the error performance of coherent receivers. The Euclidean distance is a function of the pulse shape $g(t)$, the

modulation index h , and the number of symbol intervals N over which the received signal is to be examined before a decision is made. It is therefore one of the major factors to be considered when contemplating the use of a CPM scheme.

With improvement in spectral efficiency comes the problem of increased receiver complexity. Achieving optimum detection over N symbol interval requires complex receiver structures. Carrier synchronization can also be a problem since the difficulty of finding the exact phase of the carrier increases as the spectrum of the signal narrows. This has helped to provide the impetus for the search for receiver structures with reduced complexity, yet good S/N performance. For example in [26] and [27] the use of Quadrature receivers has been examined. The receivers were shown to be useful only for schemes with modulation indices, $h=0.5$. Aulin and Sundberg [28] have also considered differential detection of CPM signals and have concluded that differential detection will perform about 3db worse than coherent detection in an additive white gaussian noise environment. The same authors in another article [29] have also proposed a reduced complexity viterbi detector for CPM demodulation.

While some of the simplified receivers are still fairly complex to implement, others give poor performance, while still others are for specific applications only. The search for simple receivers continues with this thesis addressing a specific type of receiver simplification.

1.3 Objectives

A Partial Response Continuous Phase Modulated Signal, with its attractive spectral property and error performance capability, requires a maximum likelihood sequence estimator (MLSE) for optimum reception [19]. This receiver, typically, consists of a bank of matched filters followed by a Viterbi-like processor. The number of filters can be quite large and can also be difficult to implement. The Viterbi algorithm can also have a large number of states.

The modified offset quadrature receiver mentioned in the previous section has been found useful for CPM schemes with $h=0.5$. However, most of the published reports have not dealt with the synchronization of these receivers and have, in fact, assumed perfect synchronization in their analyses. El-Tanany [18] has addressed the synchronization problem of TFM for mobile radio application and has shown it to be fairly involved. Other receivers have been discussed in [21] but these also require large numbers of filters and in the non-coherent case require averaging the carrier phase which increases the number of filters.

The aim of this work is to investigate the usefulness of, and the application of a different type of receiver termed "The Equivalent Matched Filter". The new concept is aimed at replacing the bank of matched filters needed for optimum detection. The receiver is based

on the information contained in the relative positions of the zero crossing of CPM schemes. The approach is quite general and very flexible to changes in the modulation index. In fact, the system is designed in such a manner that if the modulation index is changed, there is no need to change any of the receiver hardware. Thus the receiver has the capability to provide very easily, an h -selectable system or even an h -hopping system for secure communication.

The detection of the Equivalent Matched Filter is done at IF and the circuit requires only an automatic frequency control loop as far as carrier synchronization is concerned. The processor of the optimum detector is still required. However due to the way the information is provided to the processor a simplified processor can easily be implemented.

In this thesis, the concepts behind the Equivalent Matched Filter and a theoretical analysis of its performance are presented.

This thesis also describes the design and prototype implementation of a 3RC, $h=0.8$ binary CPM modem using the Equivalent Matched Filter receiver. A thorough account of the design procedure, behind both the modulator and demodulator is given. Its error performance is measured and compared with theoretically derived models.

Brief discussions on related circuitry such as AFC loops and on processors are also given.

Throughout this work the channel is assumed to be an additive white gaussian channel, and only binary transmission is considered.

1.4 Organization

The thesis is organized in the following manner. Chapter 2 characterizes Continuous Phase Modulation Schemes in a general way and presents the power spectrum for a few of these schemes. It pays particular attention to the Raised Cosine (RC) CPM scheme. The implementation of both the modulator and one form of the optimum receiver for this type of scheme is discussed briefly.

In Chapter 3 the theoretical basis for the design of "The Equivalent Matched" Filter is given. This chapter also presents an analysis which provides the error performance of the receiver.

Chapter 4 deals with the implementation and experimental performance of a 3RC, $h=0.8$ modem. Chapter 5 provides a general discussion on the receiver's virtues and its disadvantages. Here also will be found the conclusions.

CHAPTER 2

CONTINUOUS PHASE MODULATED SIGNALS2.1 Introduction

Modulation methods which result in constant signal envelopes are attractive for radio equipment. Their performances are not affected by non-linear operations on the signal. Such non-linear operations are commonly found in transmitter output amplifiers and receiver IF stages. Continuous Phase Modulated systems fall into this category and this chapter characterizes these signals generally and discusses one type, the Raised Cosine, fully.

2.2 Characterization of CPM Signals

The general block diagram of a CPM modulator is shown in Figure 2.1. The output signal $g(t)$ is described mathematically [8] as

$$S(t, \bar{\alpha}) = \sqrt{\frac{2E}{T}} \cos(\omega_0 t + \phi(t, \bar{\alpha}) + \phi_0) \quad (2.1)$$

where

$\bar{\alpha}$ is an infinitely long sequence of uncorrelated m -ary data

taking values

$$\alpha_i = \pm 1, \pm 3, \dots, \pm M-1$$

and

$\phi(t, \bar{\alpha})$ is the information bearing phase given as

Figure 2.1: CPM Modulator

$$\phi(t, \alpha) = 2\pi h \int_{-\infty}^t \sum_{i=-\infty}^{\infty} g(\tau - iT) \cdot d\tau \quad (2.2)$$

$$= 2\pi h \sum_{i=-\infty}^{\infty} \alpha_i q(t - iT) \quad -\infty < t < \infty \quad (2.3)$$

Here

$g(t)$ is referred to as the frequency pulse,

$q(t)$ is the phase pulse given as

$$q(t) = \int_{-\infty}^t g(\tau) d\tau \quad (2.4)$$

A causal CPM signal is obtained if the frequency pulse, $g(t)$, satisfies

$$g(t) = 0 \quad t < 0, t > LT$$

$$g(t) \neq 0 \quad 0 \leq t \leq LT$$

where the pulse length L is measured in numbers of symbol interval T . When L equals one ($L=1$) a "Full Response" system is obtained. For $L>1$ a "Partial Response" system results. Since this thesis is only concerned with Partial Response systems and binary data symbols, then $L>1$ and $\alpha_i = \pm 1$.

The shape and length of the impulse response of the frequency pulse $g(t)$ is crucial to the eventual spectrum of the CPM signal. Three different classes of CPM signals have been chosen for analysis in [9]. The classes are obtained by considering three types of frequency pulse $g(t)$ and varying M , L and the modulation index, h . These pulses will be described here briefly since they form a fairly common starting ground for experimental CPM schemes. The pulses are:

1) The Raised Cosine frequency pulse.

This pulse is defined as:

$$g(t) = \begin{cases} \frac{1}{2LT} (1 - \cos(\frac{2\pi t}{LT})) & 0 \leq t \leq LT \\ 0 & \text{otherwise} \end{cases} \quad (2.5)$$

Note that this pulse is a raised cosine over L symbol intervals.

The various possible schemes are generally referred to as LRC such as RC, 2RC, 3RC, etc. The 3RC has been chosen for the proposed modem feasibility study. The pulse is in no sense the optimum choice but it is selected for the study because of the following reasons:

- (a) with $h \geq 0.5$ and $L < 5$ the phase trajectories are fairly well separated
- (b) the trajectories can be approximated by straight lines, hence simplifying computation
- (c) the choice provides easy implementation.

2) The second type of pulse is called a spectrally Raised Cosine (SRC) pulse. This is given as

$$G(f) = \begin{cases} \frac{1}{4} (1 + \cos(\frac{\pi fLT}{2})) & |f| \leq \frac{2}{LT} \\ 0 & \text{otherwise} \end{cases} \quad (2.6)$$

It is called a spectrally raised cosine because it is the frequency spectrum $F(g(t))$ which forms a raised cosine. Here F denotes the Fourier Transform.

(3) The third frequency pulse is given as

$$g(t) = \frac{1}{8} \{g_0(t-T) + 2g_0(t) + g_0(t+T)\} \quad (2.7)$$

where $g_0(t)$ is given in [9].

For the specific case of $h = 0.5$ and $M = 2$. This modulation scheme is referred to as Tamed Frequency Modulation (TFM) and has been dealt with in references [14] - [18].

For comparison purposes, the power spectrum of the Raised Cosine scheme for $h=0.5/0.8$ and $L=3$ are shown along with those of MSK and TFM in Figure 2.2. Note that TFM shows the best spectral properties, followed by the $h=0.5$, RC scheme.

In general for a specific CPM scheme if the modulation index is increased the spectrum will broaden.

2.3. The Raised Cosine CPM

The RC scheme has certain attractive properties with regards to the proposed experimental and theoretical study and therefore is treated in detail here. Block diagrams of its implementation are given and one form each of the optimum and sub-optimum receivers is discussed.

The Raised Cosine RC pulse is expressed in equation (2.5) as

$$g(t) = \begin{cases} \frac{1}{2LT} \left(1 - \cos \frac{2\pi t}{LT} \right) ; & 0 \leq t \leq LT \\ 0 & ; \text{otherwise} \end{cases}$$

The phase $\phi(t, \bar{\alpha})$ of the transmitted signal $S(t, \bar{\alpha})$ was given in equation (2.3) as

Figure 2.2: Power Spectra of 3RC, $h=0.8/0.5$, TFM and MSK

$$\phi(t, \bar{\alpha}) = 2\pi h \sum_{i=-\infty}^{\infty} \alpha_i q(t-iT)$$

and can be written over any bit interval as [9] ,

$$\phi(t, \bar{\alpha}) = 2\pi h \left\{ \sum_{i=N-L+1}^N \alpha_i q(t-iT) + q(LT) \sum_{i=-\infty}^{N-L} \alpha_i \right\} \quad (2.8)$$

$$\text{for } NT \leq t \leq (N+1)T$$

This equation results because $g(t)$ is time limited and $q(t)$ is 0 for $t \leq 0$ and a constant $q(LT)$ for $t \geq LT$. Therefore for a given $g(t)$ and h , and for any bit interval N , $\phi(t, \bar{\alpha})$ is defined by a correlative state vector $\bar{\alpha}_N$ defined in [9] as

$$\bar{\alpha}_N = (\alpha_{N-1}, \alpha_{N-2}, \dots, \alpha_{N-L+1}) \quad (2.9)$$

and a phase state vector defined as

$$\theta_N = q(LT) \cdot \left\{ 2\pi h \sum_{i=-\infty}^{N-L} \alpha_i \right\} \bmod 2\pi \quad (2.10)$$

For

$$h = 2K/p$$

(K and p integers and are determined by the modulation scheme) there are p different phase states with values $0, 2\pi/p, 2 \times 2\pi/p \dots (p-1) \times 2\pi/p$. This is important for the modulator implementation discussed in section 2.4.

For the case of $g(t)$ in equation (2.5) the phase pulse $q(t)$ is given as

$$q(t) = \int_{-\infty}^t g(t) = \frac{1}{2LT} \left\{ t - \frac{LT}{2\pi} \sin \frac{2\pi t}{LT} \right\} \quad (2.11)$$

and

$$\phi(T) = 1/2. \quad (2.12)$$

Choosing $L=3$ such that the phase trajectory over a bit interval is controlled by three bits $(\alpha_{N-2}, \alpha_{N-1}, \alpha_N)$, then

$$\phi(t, \bar{\alpha}) = 2\pi h \{ \alpha_{-2} q(t+2T) + \alpha_{-1} q(t+T) + \alpha_0 q(t) \} + \theta_0 \quad (2.13)$$

for $0 \leq t \leq T$.

Here N has been set equal to zero without loss of generality and θ_0 is defined as in equation (2.10). The RC frequency and phase pulses as defined by equations (2.5) and (2.11) for $L=3$ and $h=0.8$ are shown in Figure 2.3.

One of the most important tools in the analysis of CPM Signals is the so-called phase tree. The tree is formed by all phase trajectories $\phi(t, \bar{\alpha})$ having a common starting value of zero at $t=0$, that is $\phi(0, \bar{\alpha})=0$. The ensemble is over the sequence $\bar{\alpha}$ and Figure 2.4 shows a part of such a tree for the RC scheme previously described. Phase states and correlative states are assigned to the nodes in the tree. Each node in the tree is assigned the state $(\alpha_N, \alpha_{N-1}, \alpha_{N-2})$. Note the distinct path for each sequence $(\alpha_{N-2}, \alpha_{N-1}, \alpha_N)$. It is upon this property that the design of the novel receiver will be based.

2.4 Modulator Structure

Although there are several ways in which to implement a CPM modulator, the one that is most often suggested is discussed below.

Figure 2.3: 3RC Frequency and Phase Pulses

Recall equation (2.1) and note that the transmitted signal can be written as

$$S(t, \bar{\alpha}) = \sqrt{\frac{2E}{T}} \{ I(t) \cos 2\pi f_0 t - Q(t) \sin (2\pi f_0 t) \} \quad (2.14)$$

where

$$\begin{aligned} I(t) &= \cos (\phi(t, \bar{\alpha})) \\ Q(t) &= \sin (\phi(t, \bar{\alpha})) \end{aligned} \quad (2.15)$$

and from equation (2.8)

$$Q(t, \bar{\alpha}) = \theta(t, \bar{\alpha}) + \theta_N \quad (2.16a)$$

where

$$\theta(t, \bar{\alpha}) = 2\pi h \sum_{i=N-L+1}^N \alpha_i q(t-iT) \quad (2.16b)$$

and θ_N is defined in equation (2.10).

Hence for

$$NT \leq t \leq (N+1)T$$

$$I(t) = \cos (\theta(t, \bar{\alpha})) \cos \theta_N - \sin (\theta(t, \bar{\alpha})) \sin \theta_N$$

and

$$Q(t) = \cos (\theta(t, \bar{\alpha})) \sin \theta_N + \sin (\theta(t, \bar{\alpha})) \cos \theta_N \quad (2.17)$$

One may then implement the modulator by generating $I(t)$ and $Q(t)$, multiplying them by quadrature carriers $\cos \omega_c t$ and $\sin \omega_c t$ and adding both channels. The basic structure of the implementation is shown in Figure 2.5. A brief description of the system is as follows. Values of $I(t)$ and $Q(t)$ for all the different phase trajectories are calculated and stored in ROMS 1 and 2. The incoming serial data stream is converted into a parallel stream and used to address both the $I(t)$ and

Figure 2.5: Modulator Structure

the $Q(t)$ ROMS. The output of the ROMS are multiplied appropriately by $\cos \omega_c t$ and $\sin \omega_c t$ and the two streams added. A more comprehensive description of the modulator will be given in Section 4.2 where the actual implementation of the modulator for the 3RC CPM scheme is discussed.

2.5 The Optimum Receiver

It is stated in [9] that the optimum receiver is a maximum likelihood sequence estimator. This receiver examines the transmitted bit sequence over the entire time axis before making a decision. Thus the receiver observes the received signal

$$r(t) = S(t, \bar{\alpha}) + n(t) \quad (2.18)$$

(where $n(t)$ is white Gaussian noise), and maximizes the log-likelihood function

$$\ln \{P_{r(t)/\bar{\alpha}}(r(t)/\bar{\alpha})\} = - \int_{-\infty}^{\infty} [r(t) - s(t, \bar{\alpha})]^2 dt \quad (2.19)$$

with respect to the infinitely long sequence $\bar{\alpha}$ [30]. The maximizing sequence $\bar{\alpha}$ is the maximum likelihood sequence estimator and $P_{r(t)/\bar{\alpha}}$ is the probability density function for the observed signal $r(t)$ conditioned on the infinitely long sequence $\bar{\alpha}$. It has been shown in [30] that equation (2.19) is equivalent to maximizing the correlation

$$J(\bar{\alpha}) = \int_{-\infty}^{\infty} r(t) \cdot S(t, \bar{\alpha}) dt \quad (2.20)$$

In determining the optimum receiver it is first necessary to consider a sub-optimum receiver that correlates the two signals, $S(t, \bar{\alpha})$ and $r(t)$ over N symbol intervals and then to extend N to infinity to obtain the optimum. This is achieved by defining a function

$$J_N(\bar{\alpha}) \triangleq \int_{-\infty}^{(N+1)T} r(t) \cdot S(t, \bar{\alpha}) dt \quad (2.21)$$

It is then possible to write

$$J_N(\bar{\alpha}) = J_{N-1}(\bar{\alpha}) + Z_N(\bar{\alpha}) \quad (2.22)$$

where

$$Z_N(\bar{\alpha}) = \int_{NT}^{(N+1)T} r(t) \cdot \cos(\omega_0 t + \phi(t, \bar{\alpha})) dt \quad (2.23)$$

By recursively calculating $J_N(\bar{\alpha})$ using equation (2.22), N can be extended to infinity to obtain the optimum receiver. The Viterbi Algorithm [20], [31] - [35] is a recursive procedure to choose those sequences that maximizes the log-likelihood function up to the N th symbol interval.

The receiver thus computes $Z_N(\bar{\alpha}_N, \bar{\theta}_N)$ for all possible 2^L sequences $\bar{\alpha}_N = \{\alpha_N, \alpha_{N-1}, \alpha_{N-L+1}\}$ and all possible p (where p was defined on page 18), phase states $\bar{\theta}_N$. This makes $p \cdot 2^L$ different Z_N values. For the case of the 3RC scheme with $L=3$ and $p=5$ the number of Z_N values becomes $5 \times 2^3 = 40$.

Continuing with the receiver development, since $\phi(t, \bar{\alpha})$ can be written as in equation (2.16a) then $Z_N(\bar{\alpha})$ can be written as

$$Z_N(\alpha_N, \bar{\theta}_N) = \int_{NT}^{(N+1)T} r(t) \cdot \cos(\omega_0 t + \theta(t, \bar{\alpha}) + \bar{\theta}_N) dt \quad (2.24)$$

It is, therefore, seen that $Z_N(\bar{\alpha}, \bar{\theta}_N)$ is obtained by feeding the received signal $r(t)$ into filters and sampling the outputs every $(N+1)T$ seconds. In this case a bank of bandpass filters must be used.

Writing the noise $n(t)$ as

$$n(t) = x(t) \cos \omega_0 t - y(t) \sin \omega_0 t \quad (2.25)$$

and using the basic quadrature receiver, the received quadrature components are

$$\begin{aligned} \hat{I}(t) &= \frac{1}{2} \left\{ \sqrt{\frac{2E}{T}} I(t) + x(t) \right\} \\ \hat{Q}(t) &= \frac{1}{2} \left\{ \sqrt{\frac{2E}{T}} Q(t) + y(t) \right\} \end{aligned} \quad (2.26)$$

Thus $r(t)$ is given as

$$r(t) = \frac{1}{2} \left\{ \sqrt{\frac{2E}{T}} I(t) + x(t) \right\} \cos 2\pi f_0 t - \frac{1}{2} \left\{ \sqrt{\frac{2E}{T}} Q(t) + y(t) \right\} \sin 2\pi f_0 t \quad (2.27)$$

By inserting this into equation (2.24) and assuming the double frequency terms are filtered out $Z_N(\bar{\alpha}_N, \bar{\theta}_N)$ becomes

$$Z_N(\bar{\alpha}_N, \bar{\theta}_N) = \operatorname{Re} \left\{ \int_{NT}^{(N+1)T} \{ \hat{I}(t) + j\hat{Q}(t) \} e^{j(\theta(t, \bar{\alpha}_N) + \bar{\theta}_N)} dt \right\} \quad (2.28)$$

This can be written in the form

$$Z_N(\bar{\alpha}_N, \bar{\theta}_N) = Z_{NI_1} + Z_{NQ_1} + Z_{NQ_2} + Z_{NI_2} \quad (2.29)$$

where the terms of equation (2.29) are obtained by expanding equation (2.28).

If these four terms are considered to be obtained from filters with impulse responses $h_c(t, \bar{\alpha}_N)$ and $h_s(t, \bar{\alpha}_N)$ and inputs $\hat{I}(t)$ and $\hat{Q}(t)$, then $Z_{NI_1}(t)$ can be written as

$$Z_{NI_1} = \int_{-\infty}^{\infty} \hat{I}(\tau) h_c(t, \bar{\alpha}_N) d\tau \quad (2.30)$$

Now if,

$$h_c(t, \bar{\alpha}_N) = \cos(\theta(T-t, \bar{\alpha}_N)) \quad (2.31)$$

then

$$Z_{NI_1}(t) = \int_{-\infty}^{\infty} \hat{I}(\tau) \cos(\theta(T-t+\tau, \bar{\alpha}_N)) d\tau \quad (2.32)$$

also if the output of the filter is sampled every T seconds then

$$Z_{NI_1}(T) = \int_{-\infty}^{\infty} \hat{I}(\tau) \cos(\theta(\tau, \bar{\alpha}_N)) d\tau \quad (2.33)$$

This is equal to $Z_N(I_1)$ if the filter response is zero outside the interval $NT \leq t < (N+1)T$. Similar expressions can be obtained for the other three terms of equation (2.29).

This result implies that $Z_N(\bar{\alpha}_N, \bar{\theta}_N)$ can be viewed as the output of $4M^L$ (there are M^L values of $\theta(t, \bar{\alpha}_N)$) baseband matched filters. Recall that $M=2$. The impulse responses of the filters are

$$h_c(t, \bar{\alpha}_N) = \cos 2\pi h \sum_{j=-L+1}^0 \alpha_j q((1-j)T-t)$$

and

$$h_s(t, \bar{\alpha}_N) = \sin 2\pi h \sum_{j=-L+1}^0 \alpha_j q((1-j)T-t)$$

(2.34)

$$h_s(t, \bar{\alpha}_N) = \begin{cases} \text{for } 0 < t < T. \\ 0 \text{ otherwise.} \end{cases}$$

Since

$$(t, \bar{\alpha}_N) = \sum_{j=N-L+1}^N \alpha_j q(t-jT)$$

and N is set equal to zero in the RHS of equation (2.34).

Actually the number of filters can be reduced by a factor of two, as stated in [9], since each $\bar{\alpha}$ sequence has a corresponding sequence sign reversed. The block diagram of this receiver is shown in Figure 2.6. Apart from the large number of filters the output of each filter has to be multiplied by the values of $\sin \theta_N$ and $\cos \theta_N$. This adds to the complexity of the receiver. Another difficulty is the generation of $\hat{I}(t)$ and $\hat{Q}(t)$ at the receiver. This question has not yet been addressed adequately in the literature.

It should be pointed out here that this is not the only optimum receiver structure that is discussed in the literature. However the others also require large amounts of filtering that are even more difficult to design particularly if the received signal is to be observed over any extended period of time for the receiver to approach optimality. It suffices to say here that the general optimum receiver is complicated and difficult to design.

Figure 2.6: Matched Filter Receiver: $H=M^L/2$

2.6 Sub-Optimum Receiver Using Modified Offset Quadrature Filters

Because of the large number of filters and the large amount of processing required for the receiver of Section 2.5 Aulin et al [27] have performed some work on a different type of receiver. This is called the modified offset quadrature receiver and can be implemented with only two filters and a simple decision logic circuitry. The performance of this type of receiver can be quite good, less than 2db from optimum if the impulse response of the filter extends over a number of bit intervals. It should be mentioned here that the desired impulse response can be difficult to achieve at times. A brief discussion of the principle involved follows.

Recall from equation (2.3) that the information carrying phase is

$$\phi(t, \alpha) = 2\pi h \sum_{i=-\infty}^{\infty} \alpha_i q(t-iT)$$

where

$$q(t) = \int_{-\infty}^t g(t) dt$$

The frequency pulse $g(t)$ is time limited thus

$$g(t) = 0; \quad t \leq 0, \quad t \geq LT$$

and

$$q(LT) = \int_0^{LT} g(t) dt = \frac{1}{2}$$

thus

$$q(t) = \begin{cases} 0 & ; t < 0 \\ \frac{1}{2} & ; t > LT \end{cases}$$

therefore

$$\phi(t, \bar{\alpha}_N) = 2\pi h \sum_{i=N-L+1}^N \alpha_i q(t-iT) + h\pi \sum_{i \leq N-L} \alpha_i \quad (2.35)$$

$$= 2\pi h \sum_{i=N-L+1}^N \alpha_i q(t-iT) - h\pi \sum_{i=N-L+1}^N \alpha_i + h\pi \sum_{i \leq N} \alpha_i \quad (2.36)$$

$$= \underbrace{2\pi h \sum_{i=N-L+1}^N \alpha_i \left\{ q(t-iT) - \frac{1}{2} \right\}}_{\phi_{ISI}} + \underbrace{h\pi \sum_{i \leq N} \alpha_i}_{\phi'_N} \quad (2.37)$$

Note as stated in [21]

$$q(t-iT) - \frac{1}{2} = 0$$

$$\text{when } q(t) = \frac{1}{2}; t > 0$$

therefore, the first term ϕ_{ISI} of equation (2.37) can be thought of as intersymbol interference (ISI) around the phase node. The phase node being defined here as

$$\phi'_N = h\pi \sum_{i \leq N} \alpha_i \quad (2.38)$$

and this is different from the phase state

$$\theta_N = \pi h \sum_{i \leq N-L} \alpha_i \text{ mod } 2\pi$$

defined in the previous section. Therefore the phase function can be written as

$$\phi(t, \bar{\alpha}_N) = \phi_{ISI} + \phi'_N \quad (2.39)$$

This implies that it is possible to detect the phase node ϕ_N at the times $t=(N+1)T$ and by processing ϕ_N make an estimate on α_N , the symbol transmitted in the interval $NT \leq t \leq (N+1)T$. From equation (2.38)

$$\phi_{N-1} = h \pi \sum_{i \leq N-1} \alpha_i \bmod 2\pi$$

therefore

$$\alpha_N = \frac{1}{h\pi} (\phi_N - \phi_{N-1}) \quad (2.40)$$

For the specific case of $h=0.5$ there are four different phase nodes, $0, \pi/2, \pi, 3\pi/2$. A quadrature receiver can thus be constructed since the signal can be written as

$$S(t, \bar{\alpha}) = \sqrt{\frac{E}{T}} \{ \cos(\phi(t, \bar{\alpha})) \cos \omega_c t - \sin(\phi(t, \bar{\alpha})) \sin \omega_c t \}$$

and since this signal can be beaten into baseband by quadrature carrier to obtain

$$\sqrt{\frac{E}{4T}} \cos(\phi(t, \bar{\alpha})) \text{ and } \sqrt{\frac{E}{4T}} \sin(\phi(t, \bar{\alpha}))$$

deBuda's technique [3] is then used to obtain values of α_N every T seconds. Note that since the phase nodes are $0, \pi/2, \pi, 3\pi/2$, the $\cos(\cdot)$ eye pattern will be opened every $2T$ seconds but offset by T seconds relative to the $\sin(\cdot)$ pattern. It should also be noted that the interference term, ϕ_{ISI} , should be less than $\pi/2$ otherwise the polarities of $\cos(\phi(t))$ $\sin(\phi(t))$ are not maintained. There will be some phase ambiguities which

Figure 2.7: Modified Offset Quadrature Receiver

must be removed by differential encoding. However after this is accomplished it is a simple matter deciding independently in each channel whether $\cos(\phi(t))$ is less or greater than zero at odd times and whether $\sin(\phi(t))$ is less or greater than zero at even times. A simple decision rule thus obtains a_N . A block diagram of this type of receiver is shown in Figure 2.7.

Although MSK type filters can be used as the LPF's in Figure 2.7 and are in fact used in some cases they are not optimum in the sense of lowest error performance. Several techniques exist [27] for the low pass filter design. Comparisons of performance are also given.

As mentioned in [27] and [36] this type of receiver is most useful for a modulation index of $h=0.5$. For all other modulation indices the phase nodes ϕ_N do not have the attractive constellation as that for $h=0.5$. It would thus be difficult to decode the $\sin(\cdot)$ and $\cos(\cdot)$ value meaningfully. For optimum performance this type of receiver requires quadrature coherent references which could be difficult to provide.

The above discussion concerning the capabilities and limitations of existing receiver structures presents a challenge for a new receiver concept which is capable of providing the following features:

- (a) lessen the filtering and processing required in the optimum receiver of Section 2.5.
- (b) generalize on the sub-optimum receiver of this section and,
- (c) eliminate the need for generating a coherent carrier at the receiver without degrading seriously the performance of the above two receivers.

There have been constant discussion [29], [34], [35], and [37], on methods of reducing the complexity of the Viterbi algorithm. This thesis therefore addresses the need to simplify the filtering requirement necessary before the processing can be done. The next Chapter presents the method for achieving the simplifications.

CHAPTER 3

THE EQUIVALENT MATCHED FILTER RECEIVER3.1 Introduction

This chapter presents the concepts and the theory behind a new type of receiver called "The Equivalent Matched Filter". This receiver consists of two parts. The first is a matched filter which views the received signal within each bit interval as an independent transmission of different phase functions for the 2^L patterns. The receiver utilizes the phase term ϕ_{ISI} of equation (2.37) to make a decision on the transmitted pattern over each T seconds. This view is analogous to multi-frequency or multi-phase signalling with each of the 2^L patterns representing a symbol.

The second part of the receiver is a Viterbi-like processor which corrects the pretence of independent transmissions in the matched filter, by utilizing the correlative properties of the signal to optimize the pattern decision of the filter. This also optimizes bit decisions since there is a one to one correspondence between pattern sequence and bit sequence.

Recall (Section 2.5) that the optimum receiver is a maximum likelihood sequence estimator which observes the received signal

$$r(t) = S(t, \bar{\alpha}) + n(t)$$

and maximizes the correlation

$$J(\bar{\alpha}) = \int_{-\infty}^{\infty} r(t) S(t, \bar{\alpha}) dt$$

with respect to the sequence $\bar{\alpha}$. Also recall that a new variable

$$J_N(\bar{\alpha}) = \int_{-\infty}^{(N+1)T} r(t) S(t, \bar{\alpha}) dt$$

was defined thereby allowing $J_N(\bar{\alpha})$ to be written as

$$J_N(\bar{\alpha}) = J_{N-1}(\bar{\alpha}) + Z_N(\bar{\alpha})$$

where $Z_N(\bar{\alpha})$ is given in equation (2.23).

It is useful to observe here that $J_N(\bar{\alpha})$ can be written as

$$J_N(\bar{\alpha}) = Z_N(\bar{\alpha}) + Z_{N-1}(\bar{\alpha}) + Z_{N-2}(\bar{\alpha}) \dots \quad (3.1)$$

It is from this definition that the equivalent matched filter concept is developed.

The expression for $Z_N(\bar{\alpha})$ is the mathematical definition of a correlator (matched filter) [24], [28], [29]. Thus $Z_N(\bar{\alpha})$ can be considered as a matched filter for the signal transmitted in the interval $NT \leq t \leq (N+1)T$. It therefore maximizes the signal to noise (S/N) ratio at the sampling instant $(N+1)T$ and optimizes the decision on the transmitted pattern during this period. From the form of equation (3.1) it is possible to use the past and present decisions to re-optimize the receiver, since in these Partial Response CMP schemes only certain pattern sequences are allowable. The receiver

Figure 3.1: Proposed Receiver Structure

then is in two parts as shown in Figure 3.1:

- (a) A matched filter and decision device
- (b) A processor to re-optimize the decision of the filter.

So far this receiver seems similar to the optimum receiver of Chapter 2. However the input to the processor is a sequence of patterns while the input to the processor of Chapter 2 is a series of samples. Therefore the processor only needs to validate each pattern using the previous pattern decisions of the matched filter. As the chapter progresses additional differences and benefits of this type of system will be discussed.

3.2 The Matched Filter

If the received signal within T seconds is written as

$$r(t) = B \cos(\omega_c t + \phi(t, \bar{\alpha})) + n(t) \quad (3.2)$$

and the hypothesized transmitted signal is written as

$$S(t) = A \cos(\omega_c t + \phi(t, \bar{\alpha}_j)) \quad (3.3)$$

where $\bar{\alpha}_j$ is one of the 2^L patterns, i.e., $j=1, 2, \dots, 2^L$ then

$$Z_N(\bar{\alpha}_j) = \int_{NT}^{(N+1)T} (A \cos(\omega_c t + \phi(t, \bar{\alpha}_j)) \cdot \{B \cos(\omega_c t + \phi(t, \bar{\alpha})) + n(t)\}) dt \quad (3.4)$$

The $\bar{\alpha}_j$ that maximizes $Z_N(\bar{\alpha}_j)$ is the most likely transmitted pattern.

This pattern along with alternatives such as the second and third most

likely pattern would be passed on to the processor for decision optimization.

Notice that the amplitudes A and B are irrelevant in determining which pattern maximizes $Z_N(\bar{\alpha}_j)$ since the decision is not based upon absolute values of the likelihood but rather on the relative values for different sets of $\bar{\alpha}_j$ s.

It therefore seems possible to hard limit the two signals $r(t)$ and $S(t)$ (preserving the phase relationship), perform the same correlation as in equation (3.4) and the new function $Z_N(\bar{\alpha}_j)$ would still be maximized with respect to $\bar{\alpha}_j$. It is the intent of this thesis to investigate this possibility.

It has been shown in [38] and [39] that by hard-limiting a FM signal and counting the number of zero crossings during a bit interval T , a useful detector can be made. However for the more compact CPM spectrum the frequency shift between patterns is quite small so the number of zero crossings may differ only slightly if at all. Therefore, whereas this type of receiver might allow for a poor single interval bit decision detector it could not be used at all to detect patterns. Therefore the S/N advantage of being able to view the received signal over more than one bit interval, before making a decision, would be lost. This renders the counting scheme useless.

Although the number of zero crossings may or may not differ from pattern to pattern, the relative positions of these crossings change. This means that the time between consecutive zero crossings of a particular pattern differs from other patterns and therefore represents a useful set of information about the transmitted patterns.

It is unfortunately possible to find signals within the CPM or CORPSK class which have the same relative zero crossing times for two different bit patterns. However this ambiguity can be resolved in the processing section of the receiver. Careful thought will also reveal that the relative times for a particular pattern can change slightly as a result of a non-linear phase function $\phi(t, \bar{\alpha})$ and the different bit time starting phases. This is one of the factors that has to be taken into account when considering the use of this type of receiver.

The foundation of the receiver structure is established using Figure 3.2. Here the noisy received signal is amplitude limited and two consecutive zero crossing time t_n and t_{n+1} are defined. Due to the noise and the different bit time starting phase ϕ_{NT} of a particular pattern these crossing times are continually changing. To account for this change the times are rewritten as

$$\left. \begin{aligned} t_n' &= t_n \pm \tau + \Delta t_n \\ t_{n+1}' &= t_{n+1} \pm \tau + \Delta t_{n+1} \end{aligned} \right\} \quad (3.5)$$

where t_n and t_{n+1} are the zero crossing times in the absence of noise.

τ corresponds to a time shift due to the starting phase ϕ_{NT} of the original waveform being at a value other than zero, and Δt_n and Δt_{n+1} are white noise terms.

The noise is assumed white since the noise accompanying the original signal is assumed white and the limiter is assumed to have infinite bandwidth. This means all the frequency components of interest are affected equally.

In order to escape any dependence on the initial phase ϕ_{NT} , the Relative Time defined as

$$\tau'_n = t'_{n+1} - t'_n = t_{n+1} - t_n + \Delta t_{n+1} - \Delta t_n \quad (3.6)$$

is considered. With this definition the hard-limited waveform between two consecutive zero crossings can be written as a summation of Heaviside Functions $U(t)$, thus:

$$S'_N = S' + N' = A\{U(t-t_n) - U(t-t_{n+1}) + U(t-t_n + \Delta t_n) - U(t-t_n^0) + U(t-t_{n+1}) - U(t-t_{n+1} + \Delta t_{n+1})\} \quad (3.7)$$

where

$$S' = A\{U(t-t_n) - U(t-t_{n+1})\}$$

is the signal term and

$$N' = A\{U(t-t_n + \Delta t_n) - U(t-t_n) + U(t-t_{n+1}) - U(t-t_{n+1} + \Delta t_{n+1})\}$$

is the additive white noise term.

Figure 3.2: Received Signal and its Hard-Limited Version

The total hard-limited waveform within a T second interval can be written as a summation of terms such as those of equation (3.7).

This makes it possible to build a correlator, for the hard-limited version of the received signal since a correlator is a matched filter for a signal disturbed by additive white noise [40]. The decision function of equation (2.24) then becomes approximately

$$Z'_N(\bar{\alpha}) = \frac{(N+1)T}{NT} \int S'(t, \bar{\alpha}) \cdot r'(t) dt \quad (3.8)$$

where

$S'(t, \bar{\alpha})$ and $r'(t)$ are the hard-limited versions of the hypothesized transmitted and received signals respectively.

It is worth noting that the function $Z'_N(\bar{\alpha})$ can be rewritten as

$$Z'_N(\bar{\alpha}) = \int_0^{t_1} () + \int_{t_1}^{t_2} () + \dots \int_{t_m}^T () \quad (3.9)$$

Here N is set equal to zero and the integration intervals are non-overlapping. They are taken for convenience to be the zero crossing intervals of the hypothesized transmitted signal (Figure 3.3a).

It now becomes sufficient to build a receiver which multiplies and integrates the hard-limited received signal $r'(t)$ and the hard-limited hypothesized transmitted signal $S'(t, \bar{\alpha})$ over the following time intervals:

- (a) τ_0 , between $t=0$ and the first zero crossing of $S'(t, \bar{\alpha})$
- (b) τ_n , between consecutive zero crossings of $S'(t, \bar{\alpha})$
- (c) τ_m , between the last zero crossing and the end of the bit period of $S'(t, \bar{\alpha})$

A summation of these integrations results in $Z'_N(\bar{\alpha})$, the total correlation. The hypothesized signal which shows the greatest correlation is chosen as the most likely signal; hence the pattern $\bar{\alpha}_j$ is determined. In order to take advantage of the correlative properties of the signal not only the most likely $\bar{\alpha}_j$ would be determined but also the second, third, ..., most likely. These would then be used by the processor, along with previous decisions, to determine the most likely transmitted bit in the time interval N.

The correlation process of equation (3.9) can be simplified, however, if each term of the equation can be maximized separately but with all terms attaining their maxima at the same time. This would result in the maximization of $Z'_N(\bar{\alpha})$.

Examining Figure 3.3b, it can be concluded that the maximum correlation occurs if $S'(t, \bar{\alpha})$ and $r'(t)$ are phase coherent that is

$$t_{nr} = t_{ns} \quad \text{for all } n \quad (3.10a)$$

where t_{nr} and t_{ns} represent the corresponding relative times of the received and transmitted signals respectively. Thus if the hard-limited signal between zeros is written as

$$\begin{aligned} r'_0(t) &= A\{U(t-t_{nr}) - U(t-t_{(n+1)r})\} \\ \text{and} \quad S'_0(t) &= A\{U(t-t_{ns}) - U(t-t_{(n+1)s})\} \end{aligned} \quad (3.11)$$

Figure 3.3: Correlation of Hard-limited Waveforms

then each term of equation (3.9) becomes

$$Z_{N_0}(\bar{\alpha}) = \int_{t_{ns}}^{t_{(n+1)s}} S'_0(t, \bar{\alpha}) \cdot r'_0(t) dt \quad (3.12a)$$

and

$$Z_{N_0}(\bar{\alpha}) = K(\tau_c - \tau_D) \quad (3.12b)$$

where the subscript '0' indicates a zero crossing. K is a constant and τ_c and τ_D represent intervals of alignment and non-alignment of $S'_0(t, \bar{\alpha})$ and $r'_0(t)$ within τ_{ns} see Figure 3.3 .

This integral is maximized if

$$(1) \quad \tau_c = \tau_{ns} = \tau_{nr} \quad (3.13a)$$

or if

$$(2) \quad \tau_c = \tau_{ns} < \tau_{nr} \quad (3.13b)$$

Fortunately each bit pattern of a CPM signal has a particular relative time pattern, therefore an equivalent receiver can be constructed which simply seeks to satisfy equation (3.13a) for all n . This can be accomplished by measuring the times between consecutive zeros of the received signal and comparing them with the known corresponding values of the ideal transmitted signals. The decision criterion is then set as follows.

If

$$\tau_{nr} = \tau_{nsj} \quad (3.14)$$

for all $n = 0, 1, 2, \dots, m$.

and $\bar{\alpha}_j$ is one of the 2^L patterns then $\bar{\alpha}_j$ is chosen as the most likely transmitted pattern.

The amplitude limited received waveform between zeros can thus be thought of as a signal of the form

$$r_{nT} = S_T(\bar{\alpha}) + N_T \quad (3.15a)$$

where

$$S_T(\bar{\alpha}) = \begin{cases} t_{n+1} - t_n = \tau_n & \text{for } t_n \leq t \leq t_{n+1} \\ 0 & \text{otherwise} \end{cases} \quad (3.15b)$$

is the signal term and N_T is a white noise term given by

$$N_T = \Delta t_{n+1} - \Delta t_n \quad (3.15c)$$

The CPM signal within each T second interval is thus modelled as a "Frame" of Pulse Width Modulated (PWM) signal where the information is contained within the pulse width. The optimum detector for each could be described by equation (3.12a) and the optimum detector for the CPM signal is thus a series of these detectors.

Obviously the criterion set by equation (3.14) represents a noiseless situation and is too stringent. In practice one would choose the $\bar{\alpha}_j$ which gives the largest correspondence. Equation (3.14) would then be replaced by

$$\tau_{nr} = \tau_{nsj} \pm \Delta\tau_{nsj} \quad (3.15d)$$

where $\Delta\tau_{nsj}$ is set depending on

- (a) the accuracy of the measuring equipment and
- (b) the difference between the gradients of the 2^L phase slopes of the CPM signals.

The receiver as developed to this point is shown in Figure 3.4.

The received signal $r(t)$ is first band-limited and a circuit measures the distance between consecutive zero crossings for the entire bit duration. After each measurement is made, the information is sent to comparators where it is compared with the corresponding zero crossing intervals of all 2^L patterns. At the end of T seconds a tally is made of the number of correspondence for each of the patterns. These are then compared to determine in chronological order the likelihood of each pattern. This is referred to as the single count method since only one measurement per pattern is compared each time.

It might be argued that the time count could be accumulated over the entire T seconds (accumulated count method) before a comparison is made. However there is an advantage in not doing so. This can be visualized if two signals, with similar gradients as shown in Figure 3.5 are considered. An accumulated count cannot differentiate such patterns since they would show the same total count, while each pair of relative time in the single count method must correspond. In other words the time position of each measurement is important.

Figure 3.4: Theoretical Matched Filter and Decision Circuit

Figure 3.5: Phase Path With Equal Zero Crossing Intervals

$$(\dot{\phi}_{11} = \dot{\phi}_{22}; \dot{\phi}_{12} = \dot{\phi}_{21})$$

In the foregoing discussion, equations (3.5) - (3.15) form the essence of the Equivalent Matched Filter receiver. However in carrying out the analysis certain assumptions were made and these will now be clarified.

Firstly it was mentioned that due to the phase continuity of the CPM signal a particular pattern has several starting phases ϕ_{NT} , and because the phase function $\phi(t, \alpha)$ can be non-linear the relative times will change. This is illustrated by the following analysis. It is assumed that the output of the zero crossing detector is a stream of pulses defined as in [39] by the equation

$$x(t) = \begin{cases} p & t_n < t < t_n + \tau \\ 0 & t_n + \tau < t < t_{n+1} \end{cases}$$

where t_n represents the zero crossing positions of the waveform, i.e., positions such that

$$\omega_c t_n + \phi(t_n) = \frac{(2n+1)\pi}{2} \quad (3.16)$$

note that the (N+1)th pulse occurs at t_{n+1} defined by

$$\omega_c t_{n+1} + \phi(t_{n+1}) = \frac{(2n+3)\pi}{2} \quad (3.17)$$

Subtracting equation (3.16) from equation (3.17) gives

$$\omega_c (t_{n+1} - t_n) + \phi(t_{n+1}) - \phi(t_n) = \pi \quad (3.18)$$

Now if the phase function $\phi(t)$ can be approximated by a straight line between successive zero crossings $\phi(t_{n+1})$ can be replaced to within a small error by

$$\phi(t_{n+1}) = \phi(t_n) + \dot{\phi}(t_n) \cdot (t_{n+1} - t_n) \quad (3.19)$$

By substituting equation (3.19) into equation (3.18) the relative time between zeros can be obtained as

$$t_{n+1} - t_n = \frac{\pi}{\omega_c + (t_n)} = \tau_n \quad (3.20)$$

Consider now, two different starting phases $\phi_{NT}(1)$ and $\phi_{NT}(2)$ for the same bit pattern. If $\dot{\phi}_1(t_n)$ differs from $\dot{\phi}_2(t_n)$, then $\tau_n(1)$ is different from $\tau_n(2)$ where the 1 and 2 denote values corresponding to $\phi_{NT}(1)$ and $\phi_{NT}(2)$ respectively.

Theoretically one overcomes this problem by phase-locking the carrier to the bit rate clock at the modulator and by building a receiver that compares values of τ_{js} for the different values of ϕ_{NT} . Thus if there are p values of ϕ_{NT} , the number of comparisons will be p times the number for a single starting phase.

The problem though is not so simple. It is compounded by the unknown phase of the Up and Down-converter Local Oscillators and other transmission medium characteristics. Fortunately, with most of the CPM schemes the phase paths are smooth and fairly linear over several

cycles of the carrier. Therefore the phase drift should not significantly affect the relative times. However it is impossible to know τ_0 and τ_m accurately because of unknown ϕ_{NT} . Therefore these two values will have to be neglected in the receiver implementation. The signals are thus only matched in frequency.

The second assumption made is that no frequency drift exists. Obviously this has to be accounted for, since the counting circuit of Figure 3.4 requires a constant carrier frequency for proper operation. Otherwise the stored values of the relative times (τ_{js}) must be continually updated.

Depending on the CPM schemes, or more specifically depending on the values of h , the modulation index, and the length, L , of the frequency pulse $g(t)$, frequency stability can be achieved by one of the following methods:

- (1) Stabilized Crystal Oscillators. The acquisition time here is zero, however this is only suitable for large values of h and small values of L ,
- (2) Automatic Frequency Control (AFC) as used in conventional FM receivers. The use of this method is not limited by the values of h or L however the acquisition time must be kept short.
- (3) Using a microprocessor to calculate the carrier frequency and to update the stored values of τ_{js} depending on what

carrier frequency is calculated. This method is limited by the speed of the processor and the time required to get a proper estimate of the carrier frequency.

The third question to be addressed concerns the spectral density N'_0 of the noise process N' of equation (3.7). As is customary the transmitted signal is written as

$$S(t) = \sqrt{2} A_c \cos(\omega_c(t) + \phi(t))$$

and the noise as

$$n(t) = \sqrt{2} n_1(t) \cos \omega_c t - \sqrt{2} n_2(t) \sin \omega_c t$$

where $n_1(t)$ and $n_2(t)$ are Gaussian processes with one sided spectral density N_0 w/radian, therefore the received signal

$$r(t) = S(t) + n(t)$$

becomes

$$r(t) = \sqrt{2} A_c \cos(\omega_c t + \phi(t)) + \sqrt{2} n_1(t) \cos \omega_c t + \sqrt{2} n_2(t) \sin \omega_c t \quad (3.21)$$

Viterbi [41] has shown that this can be written in the form

$$r(t) = \sqrt{2[(A_c + n'(t) + n''(t))^2]} \cos(\omega_c t + \phi(t) + \tan^{-1} \left(\frac{n''(t)}{A_c + n'(t)} \right)) \quad (3.22)$$

where because of the assumed independence between the modulation $\phi(t)$ and the noise $n(t)$ the transformations

$$\left. \begin{aligned} n'(t) &= n_1(t) \cos \phi(t) + n_2(t) \sin \phi(t) \\ n''(t) &= -n_1(t) \sin \phi(t) + n_2(t) \cos \phi(t) \end{aligned} \right\} \quad (3.23)$$

have been made. These are Gaussian processes with the same distribution as $n_1(t)$ and $n_2(t)$. Therefore $r(t)$ can be written in the form

$$r(t) = R(t) \cos (\omega_c(t) + \phi(t) + \phi_n(t)) \quad (3.24)$$

where $R(t)$ and $\phi_n(t)$ are defined appropriately from equation (3.22).

Examining the phase noise term

$$\phi_n(t) = \tan^{-1} \frac{n''(t)}{A_c + n'(t)}$$

more closely it is seen that for large signal to noise ratios, the approximation

$$\phi_n(t) = \tan^{-1} \frac{n''(t)}{A_c} \quad (3.25)$$

can be made. With this in mind the spectral density, $S'_n(\omega)$ of $\phi_n(t)$ becomes

$$S'_n(\omega) = \frac{N_0}{A_c^2} \quad (3.26)$$

However since the interest lies in the time variation rather than the phase variation and since

$$\omega = \frac{d\phi}{dt}$$

then the spectral density of the process can be written as

$$T(\omega) = \frac{N_0}{A_c^2 \omega^2} \quad (3.27)$$

It was previously assumed that the phase function $\phi(t) + \phi_n(t)$ was unaffected by the limiter, therefore at the output of the limiter the spectral density of the noise process $N'_L(t)$ becomes

$$N'_{OL} = \frac{A^2 N_0}{A_c^2 \omega^2} \quad (3.28)$$

where the limiter is assumed to have transfer function $H(f)=A$.

If the signal is further passed through an amplifier as shown in Figure 3.3 such that the output signal power is equal to the signal power preceeding the limiter, then we can write,

$$A_c^2 = A^2 K^2 \quad (3.29)$$

where K is the voltage amplification (the ms value of a square wave is A^2). The power spectral density of the noise becomes

$$N''_0 = \frac{A^2}{A_c^2} \cdot \frac{N_0 K^2}{\omega^2} \quad (3.30)$$

but from equation (3.29)

$$\frac{A^2}{A_c^2} = \frac{1}{K^2}$$

Therefore

$$N''_0 = \frac{1}{K^2} \frac{N_0 K^2}{\omega^2} = \frac{N_0}{\omega^2} \quad (3.31)$$

This is an important result since it means that within a factor

of ω_c^2 the noise density is the same before the limiter as it is after the limiter/amplifier combination.

In conventional FM receivers the output noise density is parabolic. However in CPM schemes the modulation index h is small and the spectrum is narrow. Therefore the noise spectrum can be considered flat over the bandwidth of interest. This is borne out in Figure 3.6 where the spectrum of the "time" noise is shown. Also it should be noted that the higher the carrier frequency the more accurate the assumption of a flat spectrum becomes. It is safe to assume a flat spectrum for frequencies within ten percent of the carrier frequency ω_c . The noise density can thus be written as

$$N''_{OL} = \frac{N_0}{2\omega_c} \quad (3.32)$$

It was previously stated that the optimum receiver for each hard-limited waveform between consecutive zero crossings is given by equation (3.29). The total CPM waveform over T seconds is viewed as being comprised of a number of these pulses and within each τ_n seconds the pulse width can take values $w_1, w_2, \dots, w_{px2}L$. (Figure 3.7a.).

Recall that p is the number of possible starting phase ϕ_{NT} . By viewing the hard-limited received signal in this form a definite resemblance is established with a multi-amplitude system, where the widths are analogous to the different amplitude levels (Figure 3.7b).

Figure 3.6: Noise Density

Figure 3.7: Multi-Amplitude and Multi-Width Signals

Bearing this in mind and noting that a Nyquist Receiver, which compares the different amplitude levels of the received signal with the thresholds, gives similar performance to the correlator under certain conditions [42] then the receiver described in the previous sections can be considered an optimum receiver matched in frequency.

Thus although a coherent frequency reference has not been provided at the receiver and although the receiver strategy is different from the usual Matched Filter the distance properties of the signal as far as frequency is concerned is seemingly still exploited. It is, therefore, instructive to consider the Euclidean distance between the set of possible signals. The squared Euclidean distance between two signals is defined in [9] and [30] as

$$D_{12}^2 = \int_0^{NT} (S_1(t) - S_2(t))^2 dt$$

This distance should be the same as if a conventional filter matched in frequency, were used for the detection. This is because it was assumed that the phase relationships between the signals were unaffected by the limiter and the signal power was amplified after the hard limiter to its original level. However, in the Equivalent Matched filter, only time measurements are of interest and by analogy with equation (3.27) each distance should be reduced by a factor of $\frac{1}{2\omega_c}$. With this in mind it is assumed that the system error probability is calculated as though the limiter never existed.

It should be remembered, however, that the analysis is only approximate since it was assumed that the signal waveform $S(t, \bar{a})$ is a perfect sinusoid. This is based on the assumption [41] that the phase variation, $\phi(t, \bar{a})$, is slow in comparison with the carrier frequency. Also the noise analysis is valid only for high signal to noise (S/N) ratios.

It has now been shown that it is possible to hard-limit a CPM signal and construct a receiver which correlates the hard-limited received signal with a clean hard-limited version of the transmitted signal without any loss in S/N performance. Unfortunately, to do this, still requires that the receiver knows the transmitter's phase exactly. It has, however, further been established that a receiver matched in frequency can be obtained by utilizing the information contained in the zero crossing intervals and the information contained in the positions of the crossings.

The next section will briefly discuss how the information provided by the filter can be used to optimize the bit decision in the Nth bit interval.

3.3 The Processor

The objective of the processor is to optimize the decision of the receiver over N symbol intervals. The optimization process is based

upon the fact that in partial response CPM schemes only certain pattern sequences are allowable. Therefore the outputs of the matched filter are correlated with the allowable patterns over N symbol intervals and the most likely sequences chosen thereby facilitating the bit decision. The problem can be thought of as follows:

Let the transmitted patterns up to time N be defined as a set of state vectors

$$\bar{\delta}_K \triangleq \{\delta_{K1}, \delta_{K2}, \delta_{K3}, \dots, \delta_{KL}\} \quad (3.33)$$

where $K=L, L+1, \dots, N$

and L is the length of the frequency pulse.

Note that $\bar{\delta}_K$ contains all the data bits δ_{K+1} which will determine the next output pattern. With this in mind it is readily seen that there is a one to one correspondence between a sequence of state vectors $\bar{\delta}_M, \bar{\delta}_{M-1}, \dots, \bar{\delta}_N$ and a sequence of transmitted bits, $\alpha_1, \alpha_2, \dots, \alpha_N$. The problem of choosing an optimum bit sequence from the set $\delta_1, \delta_2, \delta_3$ can therefore be recast as that of choosing an optimum pattern sequence, $\bar{\delta}_M \dots \bar{\delta}_N$. Estimating the optimum sequence of pattern states can be viewed as an optimum path selection through a lattice representing the patterns. Figure 3.8 shows one such lattice for $L=3$ and $N=5$. The dashed lines represent possible transitions. For example, state vector $\bar{\delta}_4 = \{1, 0, 1\}$ can only have predecessor states $\bar{\delta}_3 = \{1, 1, 0\}$ or $\bar{\delta}_3 = \{0, 1, 0\}$; in other words, states with the same α_2 and α_3 .

Figure 3.8: Lattice Diagram: $L=3$, $N=5$

Figure 1.9: Truncated Code Tree for $L=3$

Another way of viewing the lattice diagram of Figure 3.4 is in the form of a code tree as shown in Figure 3.9. Each code tree represents a sequence of allowable patterns through time beginning with the one at the top of the tree. The trees shown are for $L=3$ and $N=3$.

Choosing the optimum sequence through a lattice of the form shown has been the subject of many papers over the years [43] - [44]. The Viterbi Algorithm [30] - [35] is one method which has been discussed for use with CPM schemes. This algorithm can be very complex to implement but in many cases can be simplified without much degradation in performance. One particularly interesting form is a simplified version by Furuyuki Yukitsuna [37] which can easily be adopted to the CPM case presented above.

As stated previously, decoding algorithms are well documented so it is sufficient to say that the method chosen is dependent upon the performance required and on the degree of complexity that is involved in the processor implementation.

The concepts behind the design of the Equivalent Matched Filter have now been presented. In the next section the error performance of the receiver in a white Gaussian noise environment will be discussed.

3.4 Error Performance Estimation

Generally the error performance of this type of receiver would be very difficult to analyze. However, the situation is improved since our receiver closely resembles a correlation receiver and since (as pointed out in Section 3.2) the limiter does not affect the calculations for appropriate noise density and signal energy assumptions. The analysis will be carried out with specific reference to the 3RC, $h=0.8$ CPM scheme, implemented in Chapter 4. This allows for the use of simplifying approximations during the analysis. It should be pointed out, however, that each CPM scheme merits a different set of approximations.

The first step in the analysis is to find the Euclidean distances between the pairs of patterns that will be error prone. These distances are then used to calculate the probability of error of the system. The analysis proceeds as follows..

The squared Euclidean distance between two signals is defined [8], [30] as

$$D_{12}^2 = \int_0^{NT} (S_K(t) - S_1(t))^2 dt \quad (3.35)$$

In the case of the filter derived in Section 3.2 the integration extends over only one bit interval therefore the distance becomes

$$D_{12}^2 = \int_0^T (S_K(t) - S_1(t))^2 dt \quad (3.35a)$$

where

$$S_K(t) = \sqrt{\frac{2E}{T}} \cos(\omega_0 t + \phi(t, \bar{\alpha}_1))$$

and

$$S_1(t) = \sqrt{\frac{2E}{T}} \cos(\omega_0 t + \phi(t, \bar{\alpha}_1))$$

It is shown in Appendix A that this can be simplified with appropriate approximations to give

$$D_{12}^2 = 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos 2\pi h \frac{d}{dt} \left(\sum_{i=0}^{-L+1} q(t-iT) (\alpha_{iK} - \alpha_i) \right) dt \right\} \quad (3.36)$$

where

L is the number of bits over which $g(t)$ extends

h is the modulation index, and

α_{iK} and α_{i1} are the data bits of the two patterns $\bar{\alpha}_K$ and $\bar{\alpha}_1$ respectively

Equation (3.36) is quite general and can be used to estimate the probability of error for any CPM scheme with Equivalent Matched Filter detection. It will now be demonstrated how it is used to estimate the error probability of the 3RC, $h=0.8$ CPM scheme.

Examining Figure 3.10 it can be seen that for patterns with α_2 (the middle bit) = +1 the phase gradient is positive and those with $\alpha_2 = -1$ have a negative gradient. Thus, Euclidean distances between pairs of patterns having different α_2 s will be greater than the minimum distance for any other pair of pattern where

corresponding bits differ. It is therefore prudent to detect the bit a_2 . From the figure it can also be observed that for reasonable signal to noise ratios the probability of error will be dominated by pattern pairs 010/100, 101,000,001; 011/100, 101001,000; 110/101,100; 111/101,100 (where / is to be used "mistaken for"). The error probability for a one bit observation interval (no processor) can thus be calculated as follows:

$$P(e) = P_{000} P_{e/000} + P_{001} P_{e/001} + P_{010} P_{e/010} + P_{011} P_{e/011} \\ + P_{100} P_{e/100} + P_{101} P_{e/101} + P_{110} P_{e/110} + P_{111/11} P_{e/111} \quad (3.37)$$

where for example P_{100} is the probability of transmission of the 100 pattern and $P_{e/100}$ is the probability of error given a 100 pattern is transmitted. Recall that it was assumed that each pattern is equally likely to be transmitted, therefore

$$P_{100} = P_{000} = P_{001} = \dots P_{111} = 1/8$$

The probability of error given a 100 pattern (for example) is transmitted is thus calculated as

$$P_e(100) = P_{100} P_{e/100} = \frac{1}{8} \left\{ Q \sqrt{\frac{D_{42}^2}{2N_0}} + Q \sqrt{\frac{D_{43}^2}{2N_0}} + Q \sqrt{\frac{D_{46}^2}{2N_0}} + Q \sqrt{\frac{D_{47}^2}{2N_0}} \right\}$$

where for example, D_{42} is the Euclidean distance between the 100(4) and the 010(2) patterns.

Figure 3.10: Phase Trajectories of JRC, $h=0.8$ CPM Scheme

Figure 4.1: 3RC, $h=0.8$ CPM Modulator

use of coherent matched filter concept to calculate the error performance of this type of receiver. These assumptions will now be tested against experimental results. It should be noted however that a set of assumptions that satisfies one system may not satisfy another:

Consequently each system has to be examined separately. However, the general theory presented here can be refined and used to calculate the error performance for CPM systems with the Equivalent Matched Filter type of receiver.

CHAPTER 4

AN EXPERIMENTAL 9.6 KB/S, 3RC, $h=0.8$ BINARY CPM MODEM4.1 Introduction

This chapter deals with the design and the implementation of a 9.6 KB/S, 3RC, $h=0.8$ binary CPM modem. The chapter also compares the experimental and theoretical error rate probabilities of the system. As mentioned earlier, the scheme is not the most spectrally efficient CPM scheme, however, because its phase trajectories are reasonably well separated it is suitable for the prototype design of the proposed receiver.

A bit rate of 9.6 KB/S is selected due to certain practical limitations that are discussed later in the chapter. In addition, this rate is suitable for coded digital speech transmission over mobile radio channels, as well as other low bit rate systems that require a spectrally efficient modulation scheme.

4.2 Modulator Design and Implementation

The modulator architecture to be described here is referred to in the literature as the I-Q design. This design is developed from

equations (2.14) and (2.15). Here the transmitted signal is written as

$$S(t, \bar{a}) = \sqrt{\frac{2E}{T}} (\cos(\omega_c t) \cos \phi(t, \bar{a}) - \sin \omega_c t \sin \phi(t, \bar{a}))$$

where

$\phi(t, \bar{a})$ is given as

$$\phi(t, \bar{a}) = 2\pi h \sum_{i=-\infty}^N \alpha_i q(t-iT) + h\pi \sum_{i=-\infty}^{N-L} \alpha_i$$

$$\text{for } NT \leq t \leq (N+1)T$$

This mathematical representation suggests a transmitter structure as given in Figure 2.5. With this implementation a fixed carrier frequency is required. Therefore a crystal controlled oscillator can be used to provide both the $\cos \omega_c t$ and the $\sin \omega_c t$ carrier components. This provides frequency stability.

The generation of the $\sin \phi(t, \bar{a})$ and the $\cos \phi(t, \bar{a})$ terms is suggested by inspection of equations (2.8), (2.10), (2.16a) and (2.16b). Recall that the phase $\phi(t, \bar{a})$ can be written as

$$\phi(t, \bar{a}) = \theta(t, \bar{a}) + \theta_N$$

where

$$\theta(t, \bar{a}) = 2\pi h \sum_{i=-L+1}^N \alpha_i q(t-iT)$$

and

$$\theta_N = h\pi \sum_{i=-\infty}^{N-L} \alpha_i \text{ modulo } 2\pi = h\pi \sum_{i=-\infty}^{N-L} \alpha_i + h\pi \alpha_{N-L} \quad (4.1)$$

The phase state θ_N takes on p different values when h , the modulation index, is given as

$$h = 2K/p$$

since

$$h = 0.8 = (2 \times 2)/p$$

we have $p=5$ and θ_N takes one of the values $0, 72^\circ, 144^\circ, 216^\circ$ or 288° .

From equation (4.1), θ_N can be written as

$$\theta_N = \theta_{N-1} + h\pi \alpha_i \quad (4.2)$$

where

$$i = N - L$$

and since $h = 0.8$

$$\theta_N = \theta_{N-1} + 144^\circ \alpha_{N-L} \quad (4.3)$$

Consequently the system phase functions over the N th bit interval may be written as

$$\phi(t, \alpha) = 2\pi h \sum_{i=N-L+1}^N \alpha_i \{q(t-iT)\} + \theta_{N-1} + 144^\circ \alpha_{N-L} \quad (4.4)$$

$$\text{for } NT \leq t \leq (N+1)T$$

Since we are considering a frequency pulse $g(t)$ of length $L=3$, and since N is assumed known and therefore can be set to zero the system phase becomes

$$\phi(t, \bar{\alpha}) = 2\pi h \sum_{i=-L}^0 \alpha_i \{q(t-iT)\} + \theta_{-1} + 144 \alpha_{-3} \quad (4.5)$$

for $0 \leq t \leq T$

This equation represents the phase of the 3RC scheme over the Nth bit interval. The first term represents the phase path. The 2^3 different paths were calculated and plotted in Figure 3.10. Note that they are reasonably well separated. They are also reasonably linear allowing piecewise linear approximation of these paths to be made for the circuit implementation. The effect of this piecewise linearization on the signal spectrum will be measured using the experimental circuit.

Returning to equation (4.5) it can be noted that the system phase, $\phi(t, \bar{\alpha})$, can be determined by the knowledge of the previous phase state θ_{N-1} and a data sequence $L+1$ bits long. This fact suggests the implementation shown in Figure 4.1 which is described in the following sections.

4.3 The Phase State ROM

The phase state memory is a look-up table which is used to update the system phase state, θ_N every T seconds in accordance with equation (4.2). The memory is addressed by the old phase state θ_{N-1} plus one bit, α_{-3} , of the incoming data. The phase state transition matrix is shown in Table 4.1a.

Address		Output v_n
Control Bit a_{-3}	Old Phase State ϵ_{-1}	
0	0 = 000	$\epsilon_3 = 011$
0	72 = 001	$\epsilon_4 = 100$
0	144 = 010	$\epsilon_0 = 000$
0	216 = 011	$\epsilon_1 = 001$
0	288 = 100	$\epsilon_2 = 010$
1	0 = 000	$\epsilon_2 = 010$
1	72 = 001	$\epsilon_3 = 011$
1	144 = 010	$\epsilon_4 = 100$
1	216 = 011	$\epsilon_0 = 000$
1	288 = 100	$\epsilon_1 = 001$

Table 4.1A

000 000 000		}	Phase State 0
000 000 111			
000 111 000			
000 111 111			
100 000 000		}	Phase State 4
100 111 111			

Address

Table 4.1E

Table 4.1E

As an example of how the ROM is read, consider the first row of the tables. Here

$$\alpha_{-3} = 0$$

$$\theta_{N-1} = \text{phase state } 0 = 000$$

and the output

$$v_n = 011$$

is a number correspond to phase state 216. This number v_n forms part of the address of the $\sin \phi$ and the $\cos \phi$ ROMS.

4.4 The Sin (ϕ) and Cos (ϕ) Look-Up Tables

Using equation (4.5) the phase functions $\phi(t, \bar{\alpha})$ are generated for all possible combinations of phase states, θ , and data vector permutation. These phase functions are then sampled eight times per bit and the sine and cosine values taken. These values are then quantized to the nearest level of 255 uniform quantization levels and the quantized values loaded into the $\sin \phi$ and $\cos (\phi)$ EPROMS, shown in Figure 4.1. The structure of the data in the EPROMS are, as shown in Table 4.1b. The $\sin (\phi)$ and $\cos (\phi)$ tables are addressed by the three bits v_n of the phase state ROM, the three bit data sequence $\bar{\alpha}_N = \{\alpha_{-2}, \alpha_{-1}, \alpha_0\}$ and the three bits of the sampling clock in the manner shown in Figure 4.1.

Figure 4.1: 3RC, $h=0.8$ CPM Modulator

The memory size of the $\sin(\phi)$ or $\cos(\phi)$ ROM is given as

$$S = C = (p \times 2^L \times \text{no. of samples/bit}) \text{ words} \quad (4.6)$$

Here $p = 5$

and the number of samples/bit is 8. Therefore $S = C = 5 \times 2^3 \times 8 = 320$ words.

4.5 The Low Pass Filters and Carrier Multiplication

The low pass filters (LPF) are used to suppress spurious signals around $\omega_c \pm 2\pi f_s$. These filters have to meet certain specification [15]. These are

- (a) The phase response should be linear over the signal bandwidth.
- (b) The filters should suppress the harmonics of the sampled baseband signal.
- (c) They should have flat amplitude response over the signal bandwidth so that there is no unwanted amplitude fluctuation.
- (d) These filters should be identical in order to prevent amplitude variation of the modulated signal $S(t, \alpha)$.

Dekker [15] suggests that the cutoff frequencies of these filters should be half the sampling rate. Therefore since the sampling frequency was chosen eight times the bit rate the cut off frequencies of these filters is four times the bit rate. Sixth order Butterworth filters with cut-off frequencies of $(4 \times 9.6 \text{ KHz})$ 38.4 KHz were used in the implementation. These were designed according to the theory given in [46].

The carrier multiplication was done according to [18]. Here two switches and an op-amp were used in both the I and Q channels to achieve the carrier multiplication.

A modulator was constructed according to the preceding design equations and some experimental results of the transmitted spectrum and the time varying waveforms taken. These are shown in Figures 4.2 to 4.7. Figure 4.2 is a picture of the modulating waveforms when an alternating sequence of ones and zeros is transmitted. Notice the carrier is being modulated by a triangular waveform in this case. The spectrum of the modulated signal for this sequence is shown in Figure 4.3 and, for comparison, the theoretical spectrum is shown in Figure 4.4a. The calculations are given in Appendix B. As a comparison the theoretical spectrum for an all one sequence is shown in Figure 4.4b and the corresponding experimental one shown in Figure 4.5. Notice that for the alternating sequence there are two important frequency components, one for the 101 and the other for the 010 pattern. In contrast there is only one important frequency component when an all one sequence is transmitted, this is also true for the all zero sequence. Other sequences will have several frequency components.

The unwanted tones shown in the figures are a result of the amplitude and phase imbalances in the I and Q channels of the modulator. Apart from the spreading of the spectrum, the amplitude components are of no trouble, however, the frequency modulation components cause

Scale
Vert. 1v/cm

Hor. 2ms/cm

(a)

Scale

Vert. 5v/cm

Hor. 2ms/cm

Vert. 1v/cm

Hor. 2ms/cm

(b)

(a) Baseband ($\sin(\phi(t))$ and $\cos(\phi(t))$), waveforms for the 1010 ... bit sequence.

(b) $\cos(\phi(t))$ for 1010 ... bit sequence.

Figure 4.2: Baseband Waveforms and Bit Sequences

Experimental Power Spectrum of 1010 ... Bit Sequence

Figure 4.4: Theoretical Spectrum of 1010 ... and 1111 ... Sequences

(a) 1010 ... and (b) 1111 ... sequences.

Figure 4.5: Experimental Spectrum of 1111 ... Bit Sequence

Scale

Vert. 10db/div.

Hor. 5 KHz/div.

2 2
OF / DE

excessive problems because they lessen the receiver's noise margin.

This will be discussed with more clarity later.

In order to give an idea of what the receiver will do, the down-converted and hard-limited waveforms for both the alternating and all-one sequences are shown in Figure 4.6. Notice that for the alternating sequence the zero crossings are spread out indicating different frequency components while for the all one sequence all the relative times are the same. The receiver's function is to differentiate between these times in the presence of noise.

Finally the experimental power spectrum of the modulated signal for a pseudo-random bit sequence is given in Figure 4.7. For comparison the theoretical spectrum is also shown. The spectrum does not fall off as sharply as expected. This could be due to

- (a) The piecewise linearization of the phase function $\phi(t)$ that was used in the implementation to facilitate easy calculation of the relative times.
- (b) The amplitude and phase imbalances in the two channels of the modulator.
- (c) The mixer used in the down-conversion process also produces noise which contributes to the spectrum spreading.

The implementation of the transmitter has been demonstrated and the following sections will deal with the receiver.

(a)

Scale

Vert. 5v/cm.

Hor. 10 ms/cm.

(b)

Figure 4.6: Downconverted and Hard-Limited Waveforms For

(a) 1010 sequence and

(b) 1111 sequence.

Figure 4.7: Output Spectrum for Psuedo Random Bit Sequence

Scale

Vert. 10db/div.

Hor. 5 KHz/div.

4.6 The Receiver

The block diagram of the receiver is shown in Figure 4.8. A brief description is given below followed by a detailed explanation of the essential building blocks of the system.

The incoming signal $r(t)$ is first down-converted to a suitable IF frequency to facilitate easy digital processing without unduly stressing the speed capabilities of the logic components. In this case the IF is chosen to be 50 KHz. The zero crossings of the down-converted signals are then detected and the time between each adjacent pair of crossings is measured. These values are fed to comparators, the other inputs of which arrive from ROMs containing the ideal values of the relative times for each bit pattern. These values are compared, and for each matched pair per pattern one of $2^3=8$ pattern counters is advanced by one. It should be mentioned here that although in the theoretical receiver, Section 3.2, there were $p \times 2^L$ phase paths, corresponding to the p values of the starting phase, ϕ_{NT} , and the frequency pulse response L , it is assumed here that it is possible to group all the phase paths of one pattern together since the relative times will be affected little by the different starting phases. At the end of T seconds the outputs of these counters are latched and fed to another set of comparators. The second set of comparators determines the relative magnitudes of these counters. In this way the highest, second highest (etc.) counts are determined in chronological

Figure 4.61, Receiver Block Diagram

Figure 4.8: Cont. Receiver Block Diagram

order. In this receiver only the largest count is determined since it is not intended to implement the processor.

Since each counter is associated with one and only one pattern, simple decoding logic is used to choose the pattern or patterns. These patterns would then be used in the processor to optimize the bit decision. In this case the middle bit of each pattern was chosen as the bit decision.

4.7 The ZCD and Relative Time Counting Circuit, Figure 4.9

After the signal is down-converted it is fed to a voltage comparator where the zero crossings are detected. The output of this comparator feeds two monostable multivibrators which triggers on opposite edges of the comparator waveform. The monostables feed two S/R flip-flops which generate the waveforms that enable the counting circuits of two eight-bit, high speed synchronous counters c_{11} and c_{22} . These waveforms are shown in Figure 4.9. The clock input of these two counters are connected to a 20 MHz crystal oscillator. The counters c_{11} and c_{22} thus measures the time between two consecutive zero crossings on alternate half of cycles of the incoming waveforms.

The outputs of each counter is then ANDed with the enable output of the other counter and then OR-ed with the corresponding output of the other counter. This allows each counter to be fed alternately

Figure 4.9: The ZCD Counting Circuit and Timing Waveforms

to the output latches which operates on the zero crossings of the incoming waveform. The use of two counters in this manner, enables the counters to be latched then cleared before being restarted.

The clearing of the counters is done by the zero level of the waveform shown in Figure 4.9. Notice that when each counter is disabled, the count is held for a short period of time (for latching purposes) before it is cleared. Delays of the logic components used to develop these waveforms are very important.

The output of the latches feeds the comparator circuits which will now be described.

4.8 The Comparator Circuit, Figure 4.10

For each of the $2^L=8$ bit patterns the theoretical relative times between adjacent zero crossings were measured using a logic analyser and stored in a ROM organized as shown in Figure 4.11. In fact, each theoretical value was given a 'window' having an upper and a lower bound. These bounds were determined by the measurements and by the frequency separation between the different patterns. The upper and lower bound values for each pattern and for each zero crossing ($1 \dots M$), were stored in EPROMs referred to, in Figure 4.10, as High and Low ROM respectively. The total number of EPROMs utilized is $2 \times 2^L=16$. This is p times less than the number required if each starting phase, ϕ_{NT} , is matched.

Figure 4.10: Comparator Circuit

The sixteen EPROMs are addressed in parallel by counter c_{13} . The counter c_{13} counts the incoming zero crossing pulses of the input waveform and each time one is received the counter instructs each EPROM which of the M values of the stored times is to be read into the comparator. While the EPROMs are being addressed the output of the latches of Figure 4.9 would also be feeding the comparators. The EPROM memory is organized in such a way that if the value of the address is c_N then values of τ_{N-1} are read from the memory except for c_0 which reads τ_m (see Section 3.2 for the definition of τ_m). For example if the value of counter c_{13} is 011, (decimal three) then the values of the relative times being read from memory into the comparatoes would be τ_2 . It was necessary to place τ_m in memory location zero because the address counter is cleared every T seconds at which time, the time counter latches would have the values of τ_m of the incoming signal. It is easy to see that some patterns will have less zero crossings than others. For these patterns memory location τ_m is loaded with a value that is never expected to occur. It is worth noting that the EPROMs are only enabled after the data from c_{11} and c_{22} are latched. The timing waveforms are shown in Figure 4.12.

Corresponding to the High and the Low EPROMs are High and Low Magnitude Comparators. These comparators are fed simultaneously by the EPROMs and the latched outputs of counters c_{11} or c_{22} . In this way the theoretical relative times of each pattern are compared with the relative times of the incoming signals, in a chronological order.

Figure 4.11: EPROM Memory Organization

Figure 4.12: Comparator Circuit Timing Waveforms

The outputs of the comparators are connected through a system of AND and OR gates to 4-bit pattern counters $c_1 - c_8$. Within each T second interval, whenever a measurement from counter c_{11} or c_{22} matches corresponding stored value, the corresponding pattern counter is incremented by one. For example, if τ_2 , the measurement from c_{11} or c_{22} , is between τ_{2H4} and τ_{2L4} (H=High and L=Low) of EPROM number 4, then the value of the count on pattern counter c_4 would be increased by one. All other counters would be unchanged.

The pattern counters are latched and cleared every T seconds. Notice that the clearing pulses in Figure 4.12 have to be delayed with respect to the latching pulses. It is also worth noting that the pattern counters are enabled by a slightly delayed EPROM enabling waveform in order to allow transients to die away and not affect the counters.

The latched value of the pattern counters are then compared, in the following manner, in order to determine the most likely transmitted pattern. The counters are compared in a tree-like structure of comparators as shown in Figure 4.13. Here the counters are first paired into four groups and each member of a pair compared with its partner. The survivor which is the greater of the two is passed on for comparison with the other survivor. These survivors are now paired into two groups which are compared as before to give a pair of survivors. These survivors are now finally compared to give the most likely transmitted pattern.

Figure 4.13: Pattern Counter Comparisons

To clarify the operation of the circuit let us assume that at the end of T seconds $c_1 > c_2 > c_3 > c_4 > c_5 > c_6 > c_7 > c_8$. Then after the first comparison the survivors would be c_1, c_3, c_5 and c_7 since counters $c_1/c_2, c_3/c_4, c_5/c_6, c_7/c_8$ were used to form the originating pairs. After the second comparison the survivors would be c_1 and c_5 assuming that after the first comparison survivor counters c_1/c_3 and c_5/c_7 formed pairs. After the final comparison c_1 is the only survivor. An explicit circuit for the comparison of each pair of counters is shown in Figure 4.14. The circuit is shown comparing c_1 and c_2 , however, all the comparisons are done with similar circuits.

Now since each counter is associated with one and only one pattern most likely transmitted pattern is obtained by use of the pattern decision circuitry which is described in the following section. As mentioned previously, for processing, it might be required to obtain the second, third, fourth ... most likely patterns. The choices would be made in a similar fashion to that used in choosing the most likely.

4.9 The Pattern Decision Circuit, Figure 4.15

Let, (see Figure 4.13)

$$\begin{array}{llll} a \equiv c_1 > c_2 & b \equiv c_3 > c_4 & c \equiv c_5 > c_6 & D \equiv c_2 > c_8 \\ \bar{a} \equiv c_2 > c_1 & \bar{b} \equiv c_4 > c_3 & \bar{c} \equiv c_6 > c_5 & \bar{D} \equiv c_8 > c_7 \end{array}$$

Figure 4.14: Comparison of Two Counters C_1 and C_2

$$\begin{aligned}
 e &\equiv c_1 \text{ or } c_2 > c_3 \text{ or } c_4 & f &\equiv c_5 \text{ or } c_6 > c_7 \text{ or } c_8 \\
 \bar{e} &\equiv c_3 \text{ or } c_4 > c_1 \text{ or } c_2 & \bar{f} &\equiv c_7 \text{ or } c_8 > c_5 \text{ or } c_6
 \end{aligned}$$

and

$$g \equiv c_{1234} > c_{5678}$$

where c_{1234} is to be read c_1 or c_2 or c_3 or c_4 , then the pattern decision circuit is designed in the following manner.

$$c_1 = a e g$$

$$c_2 = \bar{a} e g$$

$$c_3 = b \bar{e} g$$

$$c_4 = \bar{b} \bar{e} g$$

$$c_5 = c f \bar{g}$$

$$c_6 = \bar{c} f \bar{g}$$

$$c_7 = d \bar{f} \bar{g}$$

$$c_8 = \bar{a} \bar{f} \bar{g}$$

Since c_1 is associated with pattern 000, c_2 with pattern 001, etc., then these could be used to address devices containing these patterns. In the actual circuit implementation, each output pattern was obtained from a system of logic gates in the manner shown in Figure 4.15. Those patterns are then stored in the pattern storage circuit where they are used to make the bit decision. In this receiver the choice was simply made on the middle bit of each pattern due to the amount of hardware involved in implementing a well established processor about which much has already been written.

Figure 4.15: Pattern Decision Circuitry

4.10 Automatic Frequency Control

Although the automatic frequency control (afc) loop has played an important rôle in communication circuits over the years, there is very little published literature which deals with the afc loop implementation and performance. This is in contrast to the large body of literature available on phase locked loops (PLL). Automatic frequency control (afc) loops are known, however, to be able to provide faster synchronization and to be more robust to noise than PLL [48] - [50].

It was previously mentioned that because of the detection method used in this type of receiver only an afc loop is necessary as far as carrier synchronization is concerned. However the problem is complicated, because in the $h=0.8$ 3RC scheme it is difficult to generate a single carrier tone or multiples thereof. A brief discussion follows of the author's thoughts on how the afc loop can be implemented.

For this purpose the transmitted signal is viewed as a multi-frequency shift keyed signal with equal positive and negative (with respect to the carrier) frequencies (see Figure 3.10). Therefore over a long period of time the average frequency should be the carrier frequency. Thus, as shown in Figure 4.16a, if the relative times between zero crossings are averaged over a long period of time (long enough such that the time boundaries are unimportant), then the average frequency should be the IF. If this is compared with the value for the (crystal) local oscillator, then an error signal would be generated.

Figure 4.16: Automatic Frequency Control

The ROMs which contain the relative times for the hypothesized patterns would have been sectionalized. The partition is done in such a way that depending on the error signal a different portion of the memory would be used, that is, each section of memory would contain the times pertinent to the received IP. A decision device placed after the error signal would then be used to form part of the address of the ROMs to facilitate reading of the correct section of the ROM.

It also seems possible to close the loop as shown in Figure 4.16b.

These circuits have not been implemented or analyzed and therefore their practicality and performance are unknown.

This concludes the design and implementation of the Equivalent Matched Filter. Some of the circuits have not been shown explicitly because of space. The receiver performance will now be discussed.

4.11 Bit Error Rate Performance

The modem was implemented as discussed in the previous sections and white Gaussian noise was added to the system in the manner shown in the block diagram of Figure 4.17. After adding the noise, the signal was down-converted to 50 KHz for processing by the receiver. A spectrum analyzer was used to measure the signal energy to noise

density (E_b/N_0) ratio for different levels of noise, while the error rate was measured by the error detector.

The error performance curve is shown in Figure 4.18. This is compared to the theoretical error rate curve developed in section 3.4. For comparison purposes the theoretical error rate curve of QPSK is also shown. The performance of the one interval detection scheme in this experiment is less than QPSK as is expected. However the receiver is nearly optimum in high signal to noise ratio conditions. For performance improvement, a Viterbi or other similar processor can be added. The theoretical improvement for a two-bit Viterbi processor is added in Figure 4.18 to indicate how the moderate performance of the one-interval detection scheme can be significantly improved. Further improvement can be achieved by increasing the number of intervals over which the processor operates. Thus one can trade-off receiver complexity against signal to noise ratio S/N performance. The particular application of the system will dictate performance requirements.

Further investigation was done on how the receiver would perform for slight changes in the IF. Figure 4.19 is a plot of the error rates for different IF. Notice that over a region of about 400 HZ the error rate is fairly constant, indicating that the system is insensitive to frequency changes within this range.

Figure 4.17: Bit Error Rate Test Set Up

Figure 4.18: Experimental and Theoretical Error Performance

Figure 4.19: Measured Error Performance at Constant E_B/N_0 for Different IF

In conclusion to this chapter, the design, implementation and performance of a 3RC, $h=0.8$, CPM, 9.6 KB/S modem has been presented.

One of the main results of this chapter, is demonstrated by Figure 4.19. Here it is seen that the error performance of the system remains fairly constant for frequencies within 400 HZ of the IF centre frequency. This makes the circuit very useful in application with low doppler rates such as in a mobile radio environment.

The circuit, although fairly elaborate, is not difficult to design. Its size may even be reduced by the use of microprocessors in certain sections of the circuit. It is fair to say that good error rate performance is achievable. In the next chapter, the virtues, limitations and applications of the system will be discussed.

CHAPTER 5

DISCUSSION AND CONCLUSION5.1 Introduction

In this final chapter, the results of the thesis will be summarized. Here, also, are discussions regarding the practicality and application of the modem, as well as suggestions for further investigations.

5.2 Advantages and Disadvantages

A very useful quality of the Equivalent Matched Filter is the simplicity of the circuit design and its relative ease of modification for different modulation indices and frequency pulse length, L . In the selection of a bit pattern all the components are the same and the only parameters which are different are the data stored in the ROMs.

Therefore, if the value of h is changed, the only modification that would be required is that of the data in the ROMs. A change in the pulse length, of the frequency pulse $g(t)$, would require additional circuitry to account for the increased number of patterns as well as changes in the data stored in the ROMs to account for phase trajectory changes. However the circuit construction would remain the same. All the pattern selection circuits contain the same number and type of components.

With such common digital circuit components and with advances in VLSI technology in recent years, the receiver attractively lends itself to VLSI implementation. Furthermore since each pattern path through the receiver contains the same circuit elements the actual layout could be quite simple, thereby lessening manufacturing cost. Because of the size reduction capability of the receiver it could potentially be retro-fitted into existing systems without much concern for space.

Another advantage of the receiver is that it is applicable to all CPM schemes and all modulation indices provided equation (5.1) is satisfied. This is a distinct advantage over the offset quadrature receiver discussed in Chapter 2 which is only useful for index $h=0.5$.

The elimination of the need for elaborate carrier recovery schemes that require precise transmitter phase knowledge is in itself quite an achievement. However, the real advantage is that the Equivalent Matched Filter receiver lends itself to fast synchronization. This clearly makes it more useful in certain applications than the coherent receiver of Chapter 2. In certain situations an automatic frequency control circuit (afc) circuit may be required but this provides faster synchronization than phase locked carrier recovery circuits [48] - [50].

From Figure 4.19 it should also be seen that the receiver exhibits a certain amount of insensitivity to frequency offset from the intermediate frequency (IF). Therefore a certain amount of frequency error in the afc loop can be tolerated. Also the receiver should show reasonably low Doppler sensitivity.

One of the major disadvantages of this circuit is that it is cumbersome, if designed with discrete components, especially for frequency pulses with large values of L . Here the number of possible bit patterns would become large and the number of circuit components becomes excessive. For example, an increase in L from two to three would increase the number of ROMs, pattern counters and comparators in the patterns decision circuitry by a factor of two. Furthermore the phase trajectories would be more closely bunched, thus making resolution of the relative times difficult.

Another disadvantage of this receiver is that a processor is needed to extend the observation interval over more than one interval. If the Viterbi-processor is used in its unmodified form the number of states in the processor could be quite large and this would make it difficult to implement. Fortunately, a lot of simplification can be incorporated in order to reduce the complexity of the processor without seriously affecting its performance. Furthermore, as alluded to previously VLSI technology can be used to reduce the seemingly elaborate size.

5.3 Practical Considerations

Three main factors should be taken into account when implementing the Equivalent Matched filter receiver. Firstly, the relative zero crossing times for a particular bit pattern changes with different

starting phases ϕ_{NT} . Therefore for this type of receiver to be uncomplicated the phase paths have to be fairly linear over a bit interval. Otherwise the receiver would require all the phase paths for all the different starting phases to be matched. That is to say, if there are p values of ϕ_{NT} and there are 2^L phase paths the number of comparison circuits of Figure 4.10 will be increased p fold.

Fortunately, in most of the CPM schemes, this condition is satisfied and small changes in these relative times can be ignored. As a matter of fact they are taken care of in the window (equation (3.15)) allowed for each pattern. A general limiting condition is, however, set on this unwanted variation.

Consider a system where there are p different values of ϕ_{NT} , and consider two bit patterns, $\bar{a}_1$ and $\bar{a}_2$. If the relative times for any two consecutive crossings are written as

$$\tau_{j1} \text{ and } \tau_{j2} \quad (\text{where } j \text{ is an index for } \phi_{NT})$$

and

$$\tau_{j1} < \tau_{j2} \quad \text{for } j=1 \text{ (for example).}$$

then

$$\tau_{j1} < \tau_{j2} ; \quad \text{for } j=2 \dots p \quad (5.1)$$

Once the condition is satisfied any fluctuations due to the phase non-linearity can be ignored.

The choice of the receiver IF frequency is the second factor which has to be considered. This frequency is upper-bounded by the ability of the circuit to resolve the "Relative Times" of the different patterns. This may not be a factor if fast logic circuits (for example, Emitter Coupled Logic, (ECL)) is used to detect the zero crossings and measure the interval times. As well the IF frequency is lower bounded by the need to have a fair number of cycles per bit interval to give reliable data. Too high a bit rate reduces the number of cycles per bit for a given IF.

The third important factor which may affect the performance of the receiver is spike noise. In an environment containing significant spike noise, the circuit could become virtually useless since the zero crossing times would be severely affected. However, the window effect of equation (3.15a) helps to eliminate the effects of spike noise.

For example, if a spike causes a "time" that does not fit into any of the windows then none of the pattern counters will advance. Consequently, a spike will affect a decision only because of the reduced data available, on which, to make a decision.

5.4 Applications

An initial objective for this thesis was the design of a modem that could be used in the mobile radio environment or any other low bit rate

bandwidth restricted system. While the modem is useful for these systems, the most unusual aspect about the design is the ability to change the modulation index, h , without having to change any hardware in the circuit. The only modifications required are the values of the relative times in the ROMs. This aspect opens the door for the system to be used to provide secure communications using a h -hopping system. In addition an h -selective feature can be provided with the modem which can be set in the field to minimize channel interference due to the specific interfering environment.

5.5 Discussion of the Experimental Prototype

A major undertaking in this thesis has been to design, build and test a 3RC modem and to show that the receiver behaves like a Matched Filter. This has been achieved in spite of the fact that the laboratory model of the 3RC modem was particularly time consuming to implement due to the number of discrete logic components used.

From Figure 4.7 it can be seen that the experimental spectrum of the 3RC modulator approaches the theoretical. It is affected little by the linearization of the phase characteristic in the implementation. The output signal for each bit pattern (for example the all one pattern in Figure 4.5), has a number of unwanted side tones. These are due to unwanted amplitude and frequency modulation caused by amplitudes and phase imbalances in the two channels of the modulator. The amplitude

modulation is not harmful to the signal reception, however, it causes some spectrum spreading. The frequency modulation however was quite troublesome to the receiver since the relative times were caused to vary. This effect lessened the noise margin of the system. Checking the system thoroughly showed that most of the unwanted components were generated in the carrier multiplication circuitry.

Although the receiver was successfully implemented, the circuit was particularly susceptible to glitches. These affected the relative time counters and the pattern counters.

The error performance of the system has been shown to be about 1.5 db worse than that of the optimum single interval detector. It is however only about 0.5 db worse than that predicted by the theory for this type of receiver. The difference in the error performance between the optimum receiver and the Equivalent Matched Filter receiver is due to the fact that the Equivalent Matched filter receiver is matched only in frequency. It therefore fails to use the constant part of the phase path over each interval. The small differences between the theoretical curve and the experimental curve of the Equivalent Matched filter is attributable to the transmitter limitations mentioned above as well as imperfections in the receiver proof of conception implementation. For reference purposes the results also show the theoretical error performance for a two bit observation interval using a Viterbi-processor and the error performance of QPSK. One can thus get an idea of the trade off between receiver complexity and signal to noise ratio performance.

In Figure 4.19 the error performance at different frequency offsets from the IF are shown. This gives an idea of the degree of insensitivity of the receiver to a frequency shift. It shows that within about ± 200 HZ of the IF, the error performance is fairly constant. This allows a faster afc loop to be designed since the loop bandwidth can be widened and frequency lock can be considered to be achieved even before the exact IF is reached. It should be pointed out that the degree of insensitivity depends on the pulse shape $g(t)$, the pulse length L , and on the modulation index h , since the phase paths are controlled by these three factors.

5.6 Conclusions and Suggestions for Further Work

The theoretical design and experimental implementation of a 3RC, $h=0.8$ CPM modem has been described. The results show that the receiver behaves as a filter matched to the frequency of the signal in high signal to noise ratio conditions. Experimental results have been presented for decisions made by observing the received signal over one bit interval. A theoretically estimated curve shows the additional performance achievable with longer observation intervals.

The experimental receiver proved to be insensitive to delays and Doppler frequency shift. Consequently the receiver design will perform satisfactorily in many mobile environments.

While the results presented are for the 3RC scheme, the approach is quite general. The receiver can be used for digital FM systems in general. Techniques similar to those in this thesis can predict error performance of other systems.

The system can be used in bandwidth restricted low data rate environments and can easily be applied to secure communication applications.

While the number of logic gates is quite large, the modem design is fairly simple and the receiver can quite easily be implemented using VLSI technology.

Several questions remain unanswered and need to be investigated. These are discussed below.

5.6a Suggestions For Further Work

1. Fading. The receiver's performance has only been analyzed for additive white Gaussian noise disturbances. However, its performance in a fading environment such as the mobile radio environment needs to be investigated. It is suggested that either a theoretical or a simulation study be conducted since an experimental implementation will be very involved.

2. A theoretical analysis of the receivers performance under low signal to noise ratio conditions is required. Here click noise is of importance and must be accounted for.

3. The design and analysis of an automatic frequency controlled loop and a timing recovery circuit is required.

4. It is suggested that the system can be easily applied to achieve secure communication by hopping. The design, analysis and implementation of such a modem would make excellent work.

APPENDIX A

A.1 Euclidean Distance Calculations

The squared Euclidean distance between two signals was defined in equation (3.35) as

$$D^2 = \int_0^{NT} (S_K(t) - S_1(t))^2 dt$$

For an interval of one bit ($N=1$) this is given as

$$D^2 = \int_0^T (S_K(t) - S_1(t))^2 dt$$

where the two signals were defined as

$$S_K(t) = \sqrt{\frac{2E}{T}} \cos(\omega_0 t + \phi(t, \bar{\alpha}_K) + \phi_{OK})$$

and

$$S_1(t) = \sqrt{\frac{2E}{T}} \cos(\omega_0 t + \phi(t, \bar{\alpha}_1) + \phi_{O1})$$

Now

$$S_K(t)^2 = \frac{2E}{2T} \{1 + \cos[2(\omega_0 t + \phi(t, \bar{\alpha}_K) + \phi_{OK})]\} \quad (A.1)$$

$$S_1(t)^2 = \frac{2E}{2T} \{1 + \cos[2(\omega_0 t + \phi(t, \bar{\alpha}_1) + \phi_{O1})]\} \quad (A.2)$$

and

$$S_{K1}(t) = \frac{4E}{2T} \{ \cos[(2\omega_0 t + \phi(t, \bar{\alpha}_K) + \phi(t, \bar{\alpha}_1) + \phi_{OK} + \phi_{O1})] + \cos[\phi(t, \bar{\alpha}_K) - \phi(t, \bar{\alpha}_1) + \phi_{OK} - \phi_{O1}] \} \quad (A.3)$$

Therefore

$$D^2 = \int_0^T \frac{E}{T} \{ 2 + \cos 2(\omega_0 t + \phi'(t, \bar{\alpha}_K)) + \cos 2(\omega_0 t + \phi'(t, \bar{\alpha}_1)) - 2 \cos (2\omega_0 t + \phi'(t, \bar{\alpha}_K) + \phi'(t, \bar{\alpha}_1)) - 2 \cos (\phi'(t, \bar{\alpha}_K) - \phi'(t, \bar{\alpha}_1)) \} dt \quad (A.4)$$

where

$$\phi'(t, \alpha_K) = \phi(t, \bar{\alpha}_K) + \phi_{OK}$$

and

$$\phi'(t, \bar{\alpha}_1) = \phi(t, \bar{\alpha}_1) + \phi_{01}$$

Now assuming the variation of $\phi'(t, \bar{\alpha}_K)$ and $\phi'(t, \bar{\alpha}_1)$ are slow in comparison to $\omega_0 t$. This is a reasonable approximation for these CPM signals. Then

$$D^2 = \frac{E}{T} \left\{ 2T + \frac{\sin 2(\omega_0 t + \phi'(t, \bar{\alpha}_K))}{2\omega_0} + \frac{\sin 2(\omega_0 t + \phi'(t, \bar{\alpha}_1))}{2\omega_0} - \frac{\sin (2\omega_0 t + \phi'(t, \bar{\alpha}_K) + \phi'(t, \bar{\alpha}_1))}{2\omega_0} - \frac{2}{T} \int_0^T \cos (\phi'(t, \bar{\alpha}_K) - \phi'(t, \bar{\alpha}_1)) dt \right\} \quad (A.5)$$

Since $2\omega \gg 1$ (the maximum value of $\sin x$) then

$$D^2 = 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos (\phi'(t, \bar{\alpha}_K) - \phi'(t, \bar{\alpha}_1)) dt \right\} \quad (A.6)$$

which can be written in the form

$$D^2 = 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos (\theta(t, \bar{\alpha}_K) - \theta(t, \bar{\alpha}_1) + \theta_{OK} - \theta_{O1} + \phi_{OK} - \phi_{O1}) dt \right\} \quad (A.7)$$

where

θ_{OK} , θ_{O1} , $\theta(t, \bar{\alpha}_K)$ and $\theta(t, \bar{\alpha}_1)$ were defined in equation (2.16).

Now since the Matched Filter developed in Chapter 3 only matches the frequency component of the signal the distance becomes

$$D'^2 = 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos \frac{d}{dt} (\theta(t, \bar{\alpha}_K) - \theta(t, \bar{\alpha}_1) + \theta_{OK} - \theta_{O1} + \phi_{OK} - \phi_{O1}) dt \right\} \quad (A.8)$$

$$= 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos \left(\frac{d}{dt} (\phi(t, \bar{\alpha}_K) - \frac{d}{dt} (\phi(t, \bar{\alpha}_1)) dt \right) \right\}$$

$$= 2E \left\{ 1 - \frac{1}{T} \int_0^T \cos 2\pi h \frac{d}{dt} \left(\sum_{i=0}^{-L+1} q(t-iT) (\alpha_{iK} - \alpha_{i1}) \right) dt \right\}$$

(A.9)

where α_{iK} and α_{i1} are the individual bits of a pattern.

It has been stated in [9] that squared Euclidean distance for coherent CPM schemes are additive and this will be assumed true here since the Matched Filter appears to provide a "coherent frequency reference". Therefore if

$$D^2(N) = x$$

then

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APPENDIX B

3RC h=0.8 SPECTRUM FOR AN ALTERNATING1010 ... BIT SEQUENCE

In order to gain more insight into the operation of the transmitter it is instructive to show the spectrum calculation for a particular bit sequence. For ease of calculation and convenience this sequence is chosen to be the 1010 ... sequence.

Consider the modulating waveform $f(t)$ shown in Figure B.1a. The instantaneous frequency can then be written as

$$\omega_i = \omega_c + K_1 f(t) = \omega_c + \Delta\omega f(t) \quad (B.1)$$

where

ω_c is the carrier frequency

and

$\Delta\omega$ is the maximum frequency deviation

$$\theta(t) = \int_0^t \omega_i(\tau) d\tau = \omega_c t + \Delta\omega \int_0^t f(\tau) d\tau \quad (B.2)$$

$$= \omega_c t + \psi(t)$$

where

$$\begin{aligned} \psi(t) = (\Delta\omega)t & \quad -T'/4 < t < T'/4 \\ (\Delta\omega)(T'/2-t) & \quad T'/4 < t < 3T'/4 \end{aligned} \quad (B.3)$$

Figure B.1: Frequency Modulation

(a) Bit Sequence

(b) Phase Function

This waveform is shown in Figure B.1b.

It is worthwhile noting that although there was a 3RC filter between the input bit stream and the FM modulator in the implementation of Chapter 4, this filter only served to reduce the maximum frequency deviation,

Continuing with the deviation, recall that the FM signal can be written as

$$\phi_{FM} = \text{Re}\{Ae^{j\theta(t)}\} = \text{Re}\{Ae^{j\omega_c t} e^{j\psi(t)}\} \quad (\text{B.4})$$

The phase function $\psi(t)$ is periodic with T and can be expressed in a Fourier series [47] thus

$$e^{j\psi(t)} = \sum_{n=-\infty}^{\infty} F_n e^{jn\omega_0 t} \quad (\text{B.5})$$

where

$$F_n = \frac{1}{T'} \int_{-T'/4}^{3T'/4} e^{j\psi(t)} e^{-jn\omega_0 t} dt \quad (\text{B.6})$$

and

$$\omega_0 = \frac{2}{T'}$$

Note $T' = 2T$, where T is the bit interval.

Using equation (B.3) F_n becomes

$$F_n = \frac{1}{2} \left\{ \text{Sa}\left(\frac{\pi}{2} (B_1 - n)\right) + (-1)^n \text{Sa}\left(\frac{\pi}{2} (B_1 + n)\right) \right\} \quad (\text{B.7})$$

where

$$\text{Sa} \triangleq \frac{\sin x}{x}$$

and

$$B_1 \triangleq \Delta\omega/\omega_0$$

The modulated signal thus becomes

$$\begin{aligned} \phi_{\text{FM}} &= \text{Re}\{A e^{j\omega_c t} \sum_{n=-\infty}^{\infty} F_n e^{jn\omega_0 t}\} \\ &= A \sum_{n=-\infty}^{\infty} F_n \cos(\omega_c + n\omega_0) t \end{aligned} \quad (\text{B.8})$$

This expression is plotted in Figure 4.4. For small modulation indices B_1 , the spectrum is insignificant for $n > 1$.

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