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Noncoherent Detection with Post-Detection Processing for Digital FM Over Fading Channels

**by
Abdulbaset Z. Atweri, B. Sc.**

**A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfillment of the requirements
for the degree of
Master of Engineering**

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
Ottawa, Ontario, Canada**

March 1995

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**"Noncoherent Detection with Post-Detection Processing for Digital FM Over
Fading Channels"**

Submitted by

Abdulbaset Z. Atweri, B. Sc.

in partial fulfillment of the requirements
for the degree of Master of Engineering.

Thesis Supervisor
Thesis Co-supervisor
Chair, Department of Systems
and Computer Engineering

Carleton University
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Abstract

The thesis deals with the error rate performance of three binary narrow-band digital FM signals using Limiter-discriminator detection. The performance improvement due to several post-detection processing algorithms has been investigated. The signals considered are: minimum shift keying (MSK), Duobinary coded MSK and Gaussian filtered MSK (GMSK). Computer simulation was used extensively to study the receiver performance. The introduction of Inter-symbol Interference (ISI) through pre-modulation and predetection filtering severely degrades the receiver performance of the simple bit-by-bit detection algorithm. A K-state Viterbi equalizer, following an integrate and dump (I&D) post-detection filter, was used to reduce the effect of ISI and therefore improving the error rate performance of the receiver.

The BER performance was first investigated over an AWGN channel. It was shown that the I&D filter is more effective with 2-bit integration period. We considered two implementations of the Viterbi algorithm: A simple version based on the assumption that the noise is Gaussian at the output of the receiver, and a second complex version that takes the exact noise distribution. We found that the second implementation outperforms the first one by 0.8 dB and 0.1 dB at error rates of 10^{-2} and 10^{-5} , respectively. For the cases considered, the receiver provides the best performance when the 2-bit I&D filter is followed by a 4-state Viterbi equalizer. For the receiver pre-detection IF filter, a bandwidth-time product BT of 1 was approximately optimum.

The performance investigation of the receiver was then extended to the case of mobile fading channel using GMSK with B_1T product of 0.35 and $h=0.5$. The post-detection arrangements investigated are the 1-bit and 2-bit I&D filter followed by 4-state Viterbi equalizer. The performance of the receiver over a channel corrupted by cochannel interference and flat Rayleigh fading has been greatly improved compared to the conventional detection techniques. The performance of the receiver over a frequency-selective fading channel was also studied using a two-ray channel model. The effects of time dispersion on the eye pattern characteristics was first examined. Finally the error rate performance was studied. It was found that great improvements can be achieved when the equalizer is adapted to the channel conditions. With perfect adaptation the error floor can be reduced an order of magnitude compared to the fixed equalization case.

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List of Abbreviations

AWGN	Additive White Gaussian Noise
AGC	Automatic Gain Control
FM	Frequency Modulation
CPM	Continuous Phase Modulation
CPFSK	Continuous Phase Frequency Shift Keying
SD	Soft Decision
HD	Hard Decision
QPSK	Quadrature Phase Shift Keying
FSK	Frequency Shift Keying
FFSK	Fast Frequency Shift Keying
MSK	Minimum Shift Keying
GMSK	Gaussian Minimum Shift Keying
TFM	Tamed Frequency Modulation
GTFM	Generalized Tamed Frequency Modulation
BT	Bandwidth-Time product
BER	Bit Error Rate
I&D	Integrate and Dump
VE	Viterbi Equalizer
IF	Intermediate Frequency
TMTR	Transmitter
RCVR	Receiver
MLSE	Maximum Likelihood Sequence Estimation
LRC	L-symbol Raised Cosine
LSRC	L-symbol Spectrally Raised Cosine
LREC	L-symbol Rectangular
PN	Pseudonoise
SNR	Signal-to-Noise Ratio
ISI	Intersymbol Interference

List of Symbols

E_b	Signal Energy per Transmitted bit
$\Phi(t)$	Information bearing phase function
ϕ_0	Phase of the carrier at $t=0$
f_c	Carrier Frequency
a_k	Input Data sequence
h	Modulation index
$g(t)$	Frequency pulse
M	Number of Modulation levels
T	Symbol period
B_t	3-dB Bandwidth of pre-modulation filter
B	3-dB Bandwidth of post-detection filter
f_i	Instantaneous frequency
f_d	Frequency deviation
$m(t)$	Time varying signal amplitude
$\psi(t)$	Arbitrary phase function
$c(\tau;t)$	Channel impulse response
τ_{rms}	rms delay spread of the power delay profile
B_{rms}	rms Doppler spread
T_c	Channel coherence time
B_c	Channel coherence Bandwidth
$s(t)$	Transmitted signal
$s_I(t)$	In-phase component of the transmitted signal
$s_Q(t)$	Quadrature component of the transmitted signal
$n(t)$	Complex Gaussian noise
$n_I(t)$	In-phase component of the noise
$n_Q(t)$	Quadrature component of the noise
$\eta(t)$	Phase noise
$\rho(t)$	Instantaneous signal-to-noise ratio
$h(t)$	IF filter impulse response
$H(f)$	IF filter transfer function
$a(t)$	Normalized filtered signal amplitude
$\theta(t)$	Filtered signal phase angle
$\Delta\varphi(t)$	Differential phase angle at the output of the I&D filter
$N(t-lT,t)$	Number of clicks in the interval $[t-lT,t]$
U	Average SNR in the interval $[t-lT,t]$
V	Spread in SNR in the interval $[t-lT,t]$
W	Geometrical mean of SNR in the interval $[t-lT,t]$
m	Length of the ISI memory

List of Symbols

σ	Noise variance
L	Length of the impulse response of the pre-modulation filter
L_F	Length of the impulse response of the pre-detection filter
NS	Number of states
μ_i	Branch metrics of Viterbi algorithm
g_i	Polynomial coefficients of the correlative encoder
d	Delay Variable
P_e	Probability of error

1

Introduction

Narrow-band digital frequency modulation (or digital FM), has been of widespread use on account of various superior features, such as constant envelope property [1] which consequently exhibits high degree of immunity to some channel non-linearities compared to other linear modulation techniques. For example, in applications where the peak power of the modulated signal must be kept independent from the modulating waveform (this requirement is usually imposed to avoid certain channel impairment such as non-linearity), FM is more suited to such application than any other modulation scheme. This feature of FM signal results from the fact that only the frequency of the sine wave is modulated and the amplitude, and consequently the power level, is constant. [2]. Moreover, from spectrum utilization point of view, it is possible by proper selection of the modulation parameters to generate a signal with a very compact spectrum and well suppressed out of band side-lobes. The very well known examples of narrow-band digital FM signals are the Minimum shift keying (MSK) [2], Duobinary FM [11], Gaussian-filtered MSK (or GMSK)[30] and tamed frequency modulation (TFM) [31].

In addition to the inherent features of constant-envelope and narrow spectrum, the appeal of narrow-band digital FM techniques derives also from instrumentation advantages which results from the relative ease of signal generation and the availability of fully non-

coherent demodulators. These demodulators operate directly on the received wave without the need to recover the carrier frequency and phase as in some other modulation techniques. The non-coherent demodulation advantage makes digital FM a very attractive candidate for mobile communications applications. In such applications the mobile channel is severely distorted because of fading and dispersion, carrier recovery and synchronization becomes very difficult and complex to implement, therefore coherent systems become less attractive and inefficient to use, in that case non-coherent systems can provide an excellent performance with minimum equipment complexity. Non-coherent demodulation can be achieved using *differential* detection or *Limiter-discriminator* detection. Differential detection does not perform as good as limiter-discriminator in mobile communication environment because of the sensitivity of this technique to carrier frequency drift¹ which makes it less reliable and inefficient to use[24]. As a matter of fact part of the appeal of *digital FM* lies in the use of Limiter-discriminator demodulation. This detection technique is very immune to many problems suffered by differential detection, at the same time it is very simple to build, consequently it is very robust to use in mobile communication environment.

1.1 Thesis objective

This thesis will investigate the limiter-discriminator receiver of narrow-band digital FM signal with the objective of using different post-detection processing techniques to improve the bit error rate performance of such receiver over mobile communication channels.

An evolutionary approach has been in our investigations that are covered here. The investigation will start by studying the performance of the receiver in it's simplest

1. Carrier frequency drift is caused by the relative instability in the local oscillators used in mobile units.

arrangement. The different post-detection processing techniques will later on be added to the receiver and the error rate performance will be investigated accordingly over a simple AWGN channel. The combination of post-detection processing techniques which provides the best BER performance will then be used over mobile communication channel.

The bit error rate performance of digital FM limiter-discriminator receiver is severely degraded by the increased intersymbol interference (ISI). This kind of impairment is introduced to the receiver through pre-detection IF filtering and multipath delay spread of the mobile radio channel. The behavior of the ISI at the output of the discriminator receiver will be closely examined with the aid of eye diagram for different IF filters and different channel conditions. The use of proper IF filtering that provides less signal distortion and excellent out-of-band noise rejection will also be addressed.

The target post-detection processing techniques will include filtering and data detection algorithms. The Post-detection filter will be used in the form of simple integrate and dump filter (I&D) filter, and the key parameter will be the integration period of the filter. Different hard decision detection techniques will be investigated. It is well known that in the presence of ISI, the use of soft decision detection can improve the error rate performance of the receiver over that of hard decision detection [47]. More emphasis will be given to this detection technique and it will be implemented in the form of maximum likelihood sequence estimator (MLSE) following the post detection filtering.

1.2 Thesis organization

This thesis report is organized into five primary chapters: chapter two provides the necessary background for those readers who are not well familiar with the theory of digital FM signalling. It comprises three sections. In section 2.1 we review the basic properties of digital FM signal including the concept of full and partial response signalling with some

examples and special cases. In section 2.2, the operation of the non-coherent Limiter-discriminator detector is described. In the last section, the basic characteristics of the multipath mobile fading channel is discussed.

Chapter three discusses the issue of performance analysis of limiter-discriminator detection of narrow-band digital FM signal. This chapter comprises six main sections. In section 3.1 the general system model and the main system parameters of the digital FM system incorporating the discriminator receiver are presented. In section 3.2 the noise statistics at the output of the discriminator detector are derived and discussed, in particular the noise is modeled as two components, the continuous noise and the impulsive noise (clicks). The total noise distribution which results from the convolution of the two noise components is also given. In section 3.4 the issue of ISI, caused by pre-modulation and pre-detection filtering, is addressed. Some examples are discussed and presented graphically with the aid of eye diagram. In section 3.5 the performance analysis of the discriminator receiver, using hard decision detection at the output of the postdetection filter, is presented. In particular some preliminary assumptions for simplifying the analysis are made and the probability of error is derived. Section 3.6 presents the performance of the discriminator receiver employing soft decision detection technique. The theory of MLSE as a technique of estimating the transmitted data, in the presence of ISI, is presented. Viterbi algorithm is discussed as an efficient implementation of MLSE. The metrics of the Viterbi algorithm are derived for different noise assumption at the receiver output. Finally the design of M-state Viterbi equalizer is presented. No attempt is made for deriving the probability of error using soft decision technique due to the mathematical difficulties and we relied heavily on simulation for estimating the bit error rate.

Chapter four describes the system components of the communication system model used for computer simulation. This chapter comprises three main sections. In section 4.1

and 4.2 the different transmitter and receiver components are described, their functions and the way they were implemented for simulation is also discussed. Section 4.3 describes the channel model used to simulate the mobile radio channel.

Chapter five presents the results obtained through computer simulation of the system described in chapter four. Three digital FM signalling schemes are considered namely; MSK, Duobinary coded MSK and GMSK. This chapter begins by discussing the manner in which the BER performance is achieved and the accuracy of the simulation results. In the following subsections the results are presented for different receiver arrangements and different channel conditions.

Chapter six provides a concluding discussion based on the outcome of the research. It also makes recommendations as to what further research would be worthwhile.

2

Background

This chapter is intended to provide first a review of the basic properties of the digital FM signal, secondly a brief description of the non-coherent discrimination detection technique is introduced, and finally the basic characteristics of the multipath mobile fading channel is discussed

2.1 The Digital FM Signal

The general class of modulation to which *digital FM* belongs is usually referred to as *continuous phase modulation* or *CPM*. A superior performance and efficient spectrum utilization can be achieved using this class of signalling by proper selection of the parameters. This class of digital modulation is a form of phase modulation where the phase is constrained to remain continuous, i. e. no discontinuity in the phase function as is the case in *QPSK* for example [35]. A general CPM signal is representable as [1]

$$s(t) = \sqrt{\frac{2E_b}{T}} \cos(2\pi f_c t + \Phi(t) + \phi_0) \quad (2.1)$$

in this representation E_b is the signal energy per transmitted symbol, T is the symbol

duration, f_c is the frequency of the carrier in the absence of modulation, ϕ_0 is the phase of the carrier at $t=0$, and $\Phi(t)$ is the information bearing phase function and it is given by

$$\Phi(t) = 2\pi \int_{-\infty}^t \sum_{k=-\infty}^n a_k h_k g(\tau - kT) d\tau = \quad (2.2)$$

by changing the order of integration and summation equation 2.2 can be rewritten as

$$\Phi(t) = 2\pi \sum_{k=-\infty}^n a_k h_k q(\tau - kT) d\tau \quad nT \leq t \leq (n+1)T \quad (2.3)$$

where a_k is an M -ary data sequence such that $a_k \in \{\pm 1, \pm 3, \dots, \pm(M-1)\}$, (M is normally a power of 2), h_k is a sequence of modulation indices, commonly $h_k = h$ for all k , in some applications h_k varies with k in a predetermined manner, in that case the CPM signal is called *multi-h CPM*. And $g(t)$ is a frequency pulse and

$$q(t) = \int_0^t g(\tau) d\tau \quad (2.4)$$

Normally, the frequency pulse $g(t)$ is a smooth pulse shape over a finite time interval $0 \leq t \leq LT$ and normalized in such a way that

$$\int_0^{LT} g(t) dt = 1/2 \quad (2.5)$$

from the definition of CPM signal in equation 2.1 we see that the smoothness of the information carrying phase function is determined by the shape of the frequency pulse

$g(t)$. In table 2-1 a list of some commonly used frequency pulses is given (the shaded

Table 2-1: Commonly used Frequency Pulse functions $g(t)$

L-symbol Raised Cosine LRC[†]	$g(t) = \begin{cases} \frac{1}{2LT} \left[1 - \cos\left(\frac{2\pi t}{LT}\right) \right]; & 0 \leq t \leq LT \\ 0 & \text{otherwise} \end{cases}$
Generalized Tamed Frequency Modulation GTfM[‡]	$g(t) = \frac{1}{8} [a g_0(t-T) + b g_0(t) + a g_0(t+T)]$ $g_0(t) = \sin\left(\frac{\pi t}{T}\right) \left(\frac{1}{\pi t} - \frac{2 - \frac{2\pi t}{T} \cot\left(\frac{\pi t}{T}\right) - \frac{\pi^2 t^2}{T^2}}{\frac{24\pi t^2}{T^2}} \right)$
L-symbol Spectrally Raised Cosine LSRC	$g(t) = \frac{1}{LT} \cdot \frac{\sin\left(\frac{2\pi t}{LT}\right)}{\frac{2\pi t}{LT}} \cdot \frac{\cos\left(\beta \frac{2\pi t}{LT}\right)}{1 - \left(\frac{4\beta}{LT} \cdot t\right)^2}; \quad 0 \leq \beta \leq 1$
Gaussian Minimum Shift Keying GMSK	$g(t) = \frac{1}{2T} \left[Q\left(2\pi B_b \frac{t - \frac{T}{2}}{\sqrt{\ln(2)}}\right) - Q\left(2\pi B_b \frac{t + \frac{T}{2}}{\sqrt{\ln(2)}}\right) \right]$ $; 0 \leq B_b T \leq \infty$ $Q(t) = \int_t^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\tau^2/2} d\tau$
L-symbol Rectangular LREC^{††}	$g(t) = \begin{cases} \frac{1}{2LT} & ; 0 \leq t \leq LT \\ 0 & ; \text{otherwise} \end{cases}$

†. L is the pulse length, e.g., 3RC has $L=3$.

‡. By changing a , b and $g_0(t)$ a class of GTfM pulses is obtained, the case where $a=1$ and $b=2$ is the well known TFM.

††. $L=1$ yields 1REC which is commonly referred to as CPFSK

pulses are those investigated in this thesis). Since $g(t)$ is finite in length we have

$$\Phi(t) = 2\pi \sum_{k=n-L+1}^n a_k h_k q(t-kT) + \pi \sum_{k=-\infty}^{n-L} a_k h_k, \quad nT \leq t \leq (n+1)T \quad (2.6)$$

It is a common practice to use the following terminology: The case when $L = 1$ is called *full response CPM*, and when $L \geq 2$ is called *partial response CPM*. Figure 2.1 and 2.2 show respectively two full and partial response pulses $g(t)$ and their corresponding $q(t)$.

Therefore, by choosing different pulse shapes $g(t)$ and varying the size of the alphabet M and the modulation index h a broad class of CPM signaling schemes can be produced. For example, by fixing the value of h for all k 's (this will be the case from now on since the *multi-h* CPM system is not the topic of this work) and by choosing the frequency pulse $g(t)$ as a rectangular pulse of length 1 (i.e. 1REC from table 2-1), an important subclass of CPM signals is generated, this class is usually referred to in the literature as *continuous phase frequency shift keying* or *CPFSK* [1][35]. The phase function of CPFSK in interval $nT \leq t \leq (n+1)T$ is then given by

$$\Phi(t) = \frac{1}{T} \pi h a_n (t - nT) + \pi h \sum_{k=-\infty}^{n-1} a_k \quad (2.7)$$

from the above equation we see that the instantaneous frequency is given by

$$f_i(t) = \frac{1}{2\pi} \frac{d}{dt} \Phi(t) = \frac{h a_n}{2T} \quad (nT \leq t \leq (n+1)T) \quad (2.8)$$

i.e the instantaneous frequency in each signaling interval is fixed and chosen from a set of finite values M . It is a usual practice to define the increment of frequency shift as f_d which is given by

$$f_d = \frac{h}{2T} \quad (2.9)$$

from the above equation the parameter h can be defined as the ratio of the peak-to-peak frequency separation $2f_d$ to the symbol rate $1/T$, and in fact that is the reason why they call

it also the frequency deviation ratio [5].

A special popular form of binary ($M=2$) CPFSK, known as *Minimum Shift Keying MSK* or *Fast Frequency Shift Keying FFSK*, has recently found a wide range of applications. In this form the modulation index h is set to 0.5 and therefore the phase function becomes

$$\Phi(t) = \frac{\pi}{2T} a_n (t - nT) + \frac{\pi}{2} \sum_{k=-\infty}^{n-1} a_k \quad nT \leq t \leq (n+1)T \quad (2.10)$$

where the second part of the left hand side of equation 2.10 represents the phase accumulation (memory) of all symbols up to time $(n-1)T$. The information carrying phase function of MSK signal is illustrated in Figure 2.3 for a particular data sequence.

The power spectral density of a digitally modulated signal consists of a mainlobe centered around the carrier frequency and a number of sidelobes symmetrically distributed on the two sides of the mainlobe. However, from spectrum utilization point of view, it is usually desirable to have a signal with a smaller width mainlobe and fast dying sidelobes. In CPM signaling, these two desirable properties are mainly controlled by the length and the shape of the frequency pulse $g(t)$ for a fixed value of h and M . The width of the main spectral lobe is determined by the length of $g(t)$, where the longer the pulse length L , the narrower the mainlobe. The rate of reduction of the spectral sidelobes is determined by the smoothness of the pulse shape $g(t)$. For example, MSK has the worst spectral properties among the CPM signaling schemes since it uses a rectangular pulse of length T . To improve these properties a longer pulse of length $2T$ is used and the resulting scheme is the well known *duobinary FM* signaling [55] which has a higher spectral efficiency than MSK. However, the improvement in spectral efficiency can not be achieved for free. The price for such an improvement is a degradation in terms of bit error rate performance. This

Figure 2-1: Example of full response frequency pulse function for CPM (a) 1REC (b) 1RC.

Figure 2-2: Example of partial response frequency pulse function for CPM (a) 2REC (b) 2RC.

performance degradation results from the fact that the longer the pulse length, the more memory the system will have and consequently more pulse overlapping in time or intersymbol interference (ISI).

2.2 Limiter-Discriminator Detection of FM Signal

A classical non-coherent method of demodulating an FM signal is called discriminator detection. As a matter of fact, part of the appeal of digital FM system derives from the availability of this simple non-coherent detection technique. The ideal response of this detector is to produce an output signal, proportional to the time derivative of the phase function of it's input signal. The circuit used to carry out the function of this non-coherent detector is usually referred to as *limiter-discriminator*. The basic structure of the limiter-discriminator detector is shown in Figure 2.4.

The operation of this detector can be described as follows [46]: the limiter is a circuit whose output is a constant amplitude for all inputs above a critical value. It's function in an FM receiver is to remove any residual (unwanted) amplitude modulation¹ and the amplitude variations due to noise. Both of these variations would have an undesirable effect if carried through to the baseband circuits. If we have an arbitrary input signal given by

$$y_{in}(t) = m(t) \cos(\omega_o t + \psi(t)) \quad (2.11)$$

where $m(t)$ is the time varying signal amplitude and $\psi(t)$ is the phase function, then the output of an ideal limiter is given by

$$y_o(t) = V_o \cos(\omega_o t + \psi(t)) \quad (2.12)$$

1. One possible cause of this amplitude modulation is the narrow-band IF filter.

Figure 2-3: Example of the information bearing phase function using MSK (a) input binary data (b) phase function

where V_o is a positive constant representing the maximum amplitude of the output of the limiter. In addition to that, the limiting function also provides some sort of AGC action since signals from the critical minimum value (threshold) up to the same maximum value provide a constant input level to the following circuit (which is the discriminator in our case).

The FM discriminator (detector) following the limiter extracts the information that has been modulated onto the carrier via frequency variation. It should provide an output signal whose amplitude is dependent on instantaneous frequency deviation and whose frequency is dependent on the carrier rate of frequency deviation. To carry out the function of the discriminator, the signal in (2.12) is differentiated to obtain

$$e(t) = V_o \left[\omega_o + \frac{d}{dt} \psi(t) \right] \sin(\omega_o t + \psi(t)) \quad (2.13)$$

we observe that the information bearing signal appears in the envelope of (2.13). By applying the signal $e(t)$ to an envelope detector the information-bearing signal can be recovered. This process is accomplished by first rectifying the signal and then passing the result through a low-pass filter as shown in figure 2.4.

2.3 Mobile Communication Channel

In mobile radio communications there are usually more than one path through which the transmitted signal propagates towards the receiver. This multipath environment results from the fact that the transmitted signal being reflected and scattered from buildings, terrain and other objects. The net received signal then becomes the sum of all the transmitted signals traveled through different paths and picked up by the receiver. However, each potential received signal has a random time-varying amplitude and phase. Therefore, they can add up either constructively (increasing the signal level) or

Figure 2-4: Functional block diagram of a limiter-discriminator circuit

destructively (decreasing the signal level). This random process of signal variation is commonly referred to as *multi-path signal fading*. In the absence of a line-of-sight path between the transmitter and the receiver, the random nature of the multipath fading channel is characterized by Rayleigh statistics¹. In particular, the received signal is modeled by a Rayleigh distributed amplitude and a uniformly distributed phase.

2.3.1 Characteristics of Multipath Fading Channels

In general, multipath Rayleigh fading channel can be sufficiently characterized by the second order statistics of it's complex low-pass impulse response $c(\tau; t)$ defined as [37]

$$c(\tau; t) = \sum_m A_m(t) e^{j2\pi\tau_m(t)} \delta(t - \tau_m(t)) \quad (2.14)$$

where $A_m(t)$ and $\tau_m(t)$ are, respectively, the attenuation and the propagation delay along the m^{th} path at time t . These second order statistics are the autocorrelation function and power spectral density of the channel. Under the assumption that the channel is wide sense stationary, and the radio transmission medium is such that the attenuation and propagation delay through different multipaths are assumed to be independent of each other, a number of important characteristics of the channel can be obtained.

First the autocorrelation function can be written as

$$\frac{1}{2} E [c^*(\tau; t) c(\tau + \Delta t; t + \Delta t)] = \phi_c(\tau; \Delta t) \quad (2.15)$$

The case when $\Delta t = 0$ or $\phi_c(\tau; 0)$ is usually referred to as the *multipath power delay profile* of the channel and it is written as $\phi_c(\tau)$. The characteristic is basically a measure

1. If the line-of-sight path is present the channel is characterized by Rician statistics.

of the variance of the channel impulse response as a function of the propagation delay τ . It characterizes the time dispersive nature of the channel due to multipath fading. Due to the fact that the signal power is vanishingly small for a large propagation delay, the power delay profile is non-zero for only a finite range of propagation delay. The signal distortion due to channel time dispersion can be more quantified by the *rms* delay spread of the power delay profile which is given by

$$\tau_{rms} = \sqrt{\frac{\int_0^{\infty} \phi_c(\tau) (\tau - \bar{\tau})^2 d\tau}{\int_0^{\infty} \phi_c(\tau) d\tau}} \quad (2.16)$$

where $\bar{\tau}$ is the average delay spread defined as

$$\bar{\tau} = \frac{\int_0^{\infty} \tau \phi_c(\tau) d\tau}{\int_0^{\infty} \phi_c(\tau) d\tau} \quad (2.17)$$

By taking the Fourier transform of the power delay profile $\phi_c(\tau)$, an equivalent characteristic of the time dispersion of the multipath channel can be obtained in the frequency domain, i.e.;

$$\phi_c(\Delta f) = \int_{-\infty}^{\infty} \phi_c(\tau) e^{-j2\pi\Delta f\tau} d\tau \quad (2.18)$$

$\Delta\phi_c(\Delta f)$ is usually referred to as *space-frequency* correlation function. Analogous to the

power delay profile *rms* delay spread τ_{rms} , the coherence bandwidth $B_c \sim \frac{1}{\tau_{rms}}$ of the space-frequency correlation function can be defined as a measure of the spectral flatness of the multipath channel affecting the transmitted signal propagating through it. In other words, it is the bandwidth that represents the largest frequency separation Δf in which any two frequencies within the signal bandwidth are still essentially correlated.

Another important characteristic of the multipath fading channel is the so called Doppler frequency spread. By taking the Fourier transform of $\phi_c(\tau; \Delta t)$ with respect to Δt , we get

$$S(\tau; f_d) = \int_{-\infty}^{\infty} \phi_c(\tau; \Delta t) e^{-j2\pi f_d \Delta t} d\Delta t \quad (2.19)$$

which is commonly referred to as the scattering function. The Doppler power spectrum, $S(f_d)$, which characterizes the Doppler frequency dispersion of the fading channel, can be obtained by setting $\tau = 0$ in the scattering function, i.e;

$$S(f_d) = S(0; f_d) = \int_{-\infty}^{\infty} \phi_c(\Delta t) e^{-j2\pi f_d \Delta t} d\Delta t \quad (2.20)$$

This power spectrum specifies the intensity of the signal as a function of the Doppler spread of the channel. The degree of signal fading can be quantified by the rms Doppler spread of the channel (analog to the rms delay spread of the power delay profile), which is defined as

$$B_{rms} = \sqrt{\frac{\int_{-\infty}^{\infty} S(f_d) f_d^2 df_d}{\int_{-\infty}^{\infty} S(f_d) df_d}} \quad (2.21)$$

By taking the inverse Fourier transform of the Doppler power spectrum with respect to f_d , an important function called the *space-time* correlation function $\phi_c(\Delta t)$ is obtained and it is given by

$$\phi_c(\Delta t) = \int_{-\infty}^{\infty} S(f_d) e^{j2\pi f_d \Delta t} df_d \quad (2.22)$$

this function characterizes the time dynamics of the channel by means of coherence time $T_c \sim \frac{1}{B_{rms}}$ which can be defined as a measure of the time variation of the channel.

We conclude this section by saying that the mobile channel can, in general, be characterized by a dispersive signal propagation (due to multipath delay spread) and a spatial variation in signal level (due to multipath fading).

2.3.2 Types of signal fading

The severity and the manner in which the transmitted signal fades can be classified in terms of the ratio of the signalling rate $\frac{1}{T}$ (or equivalently the signal transmission bandwidth B) to the channel coherence bandwidth B_C , and the rms Doppler spread B_{rms} .

If the signal bandwidth B is less than the coherence bandwidth of the channel B_C , then the channel is *frequency non-selective* or *flat fading*, i.e., the propagation delay is approximately the same for all frequencies of the transmitted signal which implies also that

there is no time dispersion in the received signal. When the signal bandwidth B is greater than the coherence bandwidth B_c , the channel is *frequency selective* and the received signal is dispersed in time.

The ratio of the rms Doppler spread of the channel B_{rms} to the signal bandwidth B determines the degree of signal fading. Specifically if the ratio $\frac{B_{rms}}{B}$ is much less than unity then the channel is said to be *a slowly fading channel*, otherwise the channel is considered to be *a fast fading channel*.

3

Performance Analysis of Narrow-band Digital FM

3.1 Introduction

Digital FM transceivers usually involve a non-linear processing function in the transmitter and the receiver which implies that the principle of superposition does not hold in such systems. In fact this non-linear nature of the modulation and demodulation process makes the analysis of bit error rate performance of FM systems more complicated than other linear systems. The observable decision variable at the output of the receiver has, in general, non-Gaussian statistics, even if the input noise is Gaussian and, as we shall see later in this chapter, that the output noise is a mixture of a continuous noise and an impulsive noise commonly referred to as click noise. Moreover, due to the narrow-band nature of the receiver, intersymbol interference is inherent and it enters into the decision variable in a non-linear fashion.

In this chapter, we will discuss the statistics of the received signal samples and the ISI distortion considerations at the output of the digital FM discriminator receiver. These noise statistics will later be used, with a suitable assumption, for evaluating the error rate performance using hard and soft decision detection. Throughout the rest of this thesis we

will confine our study to the binary digital FM system.

3.2 System Model and Parameters

The model of a digital FM system over an arbitrary channel (this channel can be, for example, a mobile fading channel, satellite channel, etc) and employing a discriminator detector at the receiver is depicted in Figure 3.1. The data source at the transmitter is assumed to generate a sequence of independent, identically distributed random variables that takes the values ± 1 with equal probability. The received signal is assumed to be corrupted by an additive white Gaussian noise with one-sided spectral density N_0 . The IF filter $H(f)$ is assumed to be symmetric around the carrier frequency f_c with an equivalent low-pass impulse response $h(t)$. The channel has a complex low-pass impulse response $c(t)$. Since the noise is additive, the overall impulse response of the channel and the pre-detection IF filter, $h(t)$, can be expressed as the convolution of $c(t)$ and $h(t)$ and will be denoted as $h_f(t)$ which can be expressed as

$$h_f(t) = h_{fi}(t) \cos \omega_c t - h_{fq}(t) \sin \omega_c t \quad (3.1)$$

where $h_{fi}(t)$ and $h_{fq}(t)$ are respectively the in-phase and quadrature components of $h_f(t)$ about the carrier frequency ω_c . The channel input signal is given by

$$s(t) = s_i(t) \cos \omega_c t - s_q(t) \sin \omega_c t \quad (3.2)$$

where

$$s_i(t) = \sqrt{\frac{2E_b}{T}} \cos \Phi(t) \quad (3.3)$$

and

Figure 3-1: System model of Digital FM (a) Transmitter (b) Limiter-discriminator receiver

$$s_q(t) = \sqrt{\frac{2E_b}{T}} \sin \Phi(t) \quad (3.4)$$

The noise-free input to the limiter-discriminator $y(t)$, which is the convolution of $s(t)$ and $h_f(t)$, can be expressed as

$$y(t) = p(t) \cos \omega_c t - q(t) \sin \omega_c t \quad (3.5)$$

where $p(t)$ and $q(t)$ are defined as

$$p(t) = \frac{1}{2} \int_{-\infty}^{\infty} [s_i(t-\tau) h_{ii}(\tau) - s_q(t-\tau) h_{iq}(\tau)] d\tau \quad (3.6)$$

and

$$q(t) = \frac{1}{2} \int_{-\infty}^{\infty} [s_i(t-\tau) h_{iq}(\tau) + s_q(t-\tau) h_{ii}(\tau)] d\tau \quad (3.7)$$

By incorporating the additive noise at the receiver input, the noise corrupted version of (3.5) can be expressed as

$$y_c(t) = R(t) \cos(\omega_c t + \eta(t)) \quad (3.8)$$

where the envelope $R(t)$ and phase $\eta(t)$ are, respectively, defined as

$$R^2(t) = [p(t) + n_i(t)]^2 + [q(t) + n_q(t)]^2 \quad (3.9)$$

$$\eta(t) = \tan^{-1} \left(\frac{q(t) + n_q(t)}{p(t) + n_i(t)} \right) \quad (3.10)$$

and $n_i(t)$ and $n_q(t)$ are the respective in-phase and quadrature components of the input narrow-band gaussian noise $n(t)$ about the carrier frequency.

The discriminator detector produces an output signal $r(t)$ which represents the derivative of the phase function $\eta(t)$, i.e.:

$$r(t) = \frac{d}{dt} \eta(t) \quad (3.11)$$

by carrying out the derivative of (3.10) we get

$$r(t) = \frac{(p(t) + n_i(t))(\dot{q}(t) + \dot{n}_q(t)) - (\dot{p}(t) + \dot{n}_i(t))(q(t) + n_q(t))}{R^2(t)} \quad (3.12)$$

3.3 Output noise statistics for the AWGN channel

The system model for the gaussian channel can be obtained as a special case of the system model shown in Figure 3.1 by setting the impulse response of the channel $c(t)$ to an impulse at $t=0$. In that case, and assuming that the two Gaussian processes, $n_i(t)$ and $n_q(t)$, are zero-mean and unity variance, the phase function of the IF filter output signal, i.e. (3.10), can be written as [10]

$$\eta(t) = \tan^{-1} \left(\frac{\sqrt{2\rho(t)} \sin\theta(t) + n_q(t)}{\sqrt{2\rho(t)} \cos\theta(t) + n_i(t)} \right) \quad (3.13)$$

In the above representation, $\rho(t)$ is the instantaneous signal-to-noise ratio defined as

$$\rho(t) = \frac{E_b}{N_o} \cdot \frac{a^2(t)}{T \int_{-\infty}^{\infty} |H(f)|^2 df} \quad (3.14)$$

where $a(t)$ is the filtered normalized signal amplitude in the absence of noise whose square is given by:

$$a^2(t) = \left(\int_{-\infty}^t h(t-\tau) \cos \Phi(\tau) d\tau \right)^2 + \left(\int_{-\infty}^t h(t-\tau) \sin \Phi(\tau) d\tau \right)^2 \quad (3.15)$$

The phase angle $\theta(t)$ is defined as

$$\theta(t) = \tan^{-1} \left(\frac{\int_{-\infty}^t h(t-\tau) \sin \Phi(\tau) d\tau}{\int_{-\infty}^t h(t-\tau) \cos \Phi(\tau) d\tau} \right) \quad (3.16)$$

The post detection filter following the discriminator is approximated by an integrate and dump (I&D) filter whose output at any time t is the integration of it's input over an interval from $t-lT$ to t . Therefore, the I&D filter reintegrates the derivative in (3.11) producing a phase difference $\Delta\phi(t)$ which is commonly written as:

$$\Delta\phi(t) = \Delta\phi(t) + 2\pi N(t-lT, t) \quad (3.17)$$

where $\Delta\phi(t) = [\eta(t) - \eta(t-lT)] \bmod 2\pi$ represents the signal phase change due to modulation and noise (the mod 2π operation means that the phase angle is specified over some 2π interval), and N represents a discrete random variable which indicates the number of $\mp 2\pi$ phase rotations (jumps) in the interval from $t-lT$ to t , and usually called clicks.

The output of the integrate and dump filter is then sampled and applied to a decision device to recover the input data. From (3.16) we can see that the receiver ability to recover the transmitted data is governed by the statistics of the random processes

$$\Delta\phi(t) \quad \text{and} \quad N(t-lT, t)$$

3.3.1 Distribution of the continuous Noise

Pawula *et al.* [9] have derived a general expression for the probability distribution of continuous noise $\Delta\phi(t)$ or equivalently the mod 2π phase angle between two vectors perturbed by Gaussian noise. Following [9] the probability density function of $\Delta\phi(t)$ in it's general form is given by

$$p(\Delta\phi) = \frac{1-r^2-\lambda^2}{4\pi} \int_{-\frac{\pi}{2}}^{\frac{\pi}{2}} \frac{e^{-E \cos t}}{[1-Y \cos t]^2} \left[1 - E + \frac{2U - 2W(r \cos \Delta\theta + \lambda \sin \Delta\theta)}{1-r^2+\lambda^2} \right] dt \quad (3.18)$$

where

$$Y = r \cos \Delta\theta + \lambda \sin \Delta\theta$$

and

$$E = \frac{U - V \sin t - W \cos(\Delta\theta - \Delta\phi) \cos t}{1 - Y \cos t}$$

The different parameters in (3.18) are defined as follows: r and λ are, respectively, the autocorrelation and the crosscorrelation functions between the noise components given by

$$r(v) = E\{n_i(t) n_i(t+v)\} = E\{n_q(t) n_q(t+v)\} \quad (3.19)$$

$$\lambda(v) = E\{n_i(t) n_q(t+v)\} = E\{n_q(t) n_i(t+v)\} \quad (3.20)$$

and U , V and W are respectively given by

$$U = \frac{1}{2} [\rho(t) + \rho(t-lT)] \quad (3.21)$$

$$V = \frac{1}{2} [\rho(t) - \rho(t-lT)] \quad (3.22)$$

$$W = \sqrt{U^2 - V^2} \quad (3.23)$$

where U represents the average signal-to-noise ratio in the interval $[t-T, t]$, W is its geometrical mean and V is a measure of the spread in SNR. If the noise process $\Delta\phi(t)$ is such that its samples at $t-T$ and T are uncorrelated; i.e the correlation parameters r and λ are equal to zero, the density function in (3.18) can be written as

$$p(\Delta\phi) = \frac{1}{4\pi} \int_{-\pi/2}^{\pi/2} e^{-E} (1 + 2U - E) \cos t \, dt \quad (3.24)$$

where, in this case, $E = U - V \sin t - W \cos(\Delta\theta - \Delta\phi) \cos t$. The distribution of $\Delta\phi(t)$ as it is given in (3.23) is plotted in Figure 3.2 for the case when $\rho(t) = \rho(t-T)$ equals to 10 dB, 3 dB and -3 dB. As can be seen from the figure, the 10 dB case resembles the Gaussian noise distribution. For this reason some researchers approximate the distribution of the noise at the output of the I&D filter by a Gaussian pdf for high SNR.

3.3.2 Distribution of the Clicks

The number of clicks N is assumed to be a discrete random variable with a poisson distribution [3], and it is given by

$$p(N = k) = \frac{e^{-\bar{N}} \bar{N}^k}{k!}; \quad k = 0, 1, \dots \quad (3.25)$$

where $\bar{N}$ is the average number of clicks in the interval $[t-T, t]$ and is given by

$$\bar{N} = \int_{t-T}^t \frac{\dot{\theta}(\tau)}{2\pi} e^{-\rho(\tau)} d\tau \quad (3.26)$$

Figure 3-2: Probability density function of $\Delta\phi$ for $\rho(t) = \rho(t - IT)$

3.3.3 Total output noise Distribution

It is a usual practice to assume that the two random processes $\Delta\phi(t)$ and $\bar{N}$ are statistically independent. Therefore, the probability density function of the sum $\Delta\phi(t) + 2\pi N(t - lT, t)$ is given by the convolution of the two noise components $\Delta\phi(t)$ and $\bar{N}$ which is (assuming a "1" bit is transmitted or $\dot{\theta}(t) > 0$)

$$p(\Delta\phi) = \left(\sum_{i=0}^{\infty} p(\Delta\phi + 2\pi i) p\{N = i\} \right) \quad (3.27)$$

From (3.27) we can see that the probability density function of the noise, at the output of the post detection filter, consists of a main lobe centered at the transmitted data phase and extends to $\pm\pi$ on each side, plus identically shaped lobes displaced by integral multiples of 2π towards lower values. These displaced lobes are weighted by the probability of getting a given number of clicks to affect the displacement, i.e., the lobe centered at $-2i\pi$ is weighted by the probability of getting i clicks in the interval $[t - lT, t]$. There are lobes on the other side as well (i.e. towards higher values) but for large signal-to-noise ratios they are very small compared to those at the opposite side and hence they are neglected. However, if the transmitted bit is "0" or $\dot{\theta}(t) < 0$ the situation is reversed.

3.4 Intersymbol Interference Considerations

Intersymbol interference or ISI is one of the major impairments inherent to the narrow-band digital FM systems. The duration over which the ISI components have a significant value is commonly referred to as the ISI memory and it is usually given in terms of the number of bits (symbols) that the ISI spans. This memory span is controlled by the

length of the impulse response of the channel filters. For the purpose of demonstration, consider the case of MSK signaling with no transmitter filter and a Gaussian IF filter at the receiver whose low-pass impulse response is defined as

$$h(t) = B\sqrt{2}e^{-2\pi B^2 t^2} \quad (3.28)$$

where B is the 3-dB bandwidth of the filter. The impulse response of this filter is shown in Figure 3.3 for a range of normalized bandwidth BT , where T is the bit duration.

As we can see, for $BT = 1$, $h(t)$ has significant values only for a duration of $2T$. Therefore, only the bits on either side of the bit under consideration have significant ISI effects. Similarly, if $BT = 0.5$, the impulse response has significant values only over a $4T$ duration and therefore the ISI effects extend only to the two bits preceding and two bits succeeding the one under consideration. As a matter of fact, the ISI can be better visualized with the aid of eye diagrams which provide a qualitative measure of the degree of distortion, caused by ISI. Using the same Gaussian filter in (3.28), the eye diagram at the output of the discriminator was simulated and the resulting pattern is shown in Figures 3.4a and 3.4b for two different BT products. The severity of the ISI can be estimated from the eye opening at the sampling instant. It is evident from the eye diagram that as the filter bandwidth becomes narrower (i.e., a longer impulse response) the contribution from adjacent bits becomes more significant (more ISI) and consequently more signal distortion.

In the above discussion we have considered the input data to the FM modulator as square pulses of duration T (i.e. full response signaling). If these data pulses are shaped using a premodulation filter with an impulse response of length greater than T (i.e. partial response signalling), the ISI memory becomes longer and the distortion increases accordingly. Figure 3.4c and 3.4d show the eye diagram of two partial response FM signalling cases.

Figure 3-3: Impulse response of the low-pass equivalent of a Gaussian IF filter for different BT products.

Thus far we have only considered the effect of the transmitter and receiver filters, and we assumed that the channel $c(t)$, in Figure 3.1, has ideal characteristics in the sense that it does not introduce any distortion to the transmitted data. However, in practice the channel characteristics are of random nature and the distortion due to ISI behaves in an unpredictable manner and can be more severe especially if the channel has a dispersive nature as we will see when we study the performance over the mobile radio channel.

In general, for binary digital FM signaling over an ideal channel, the total span of the ISI memory, m , in bits, can be approximated by

$$m = L + L_F - 1 \quad (3.29)$$

where L is the length, in T sec, over which the impulse response of the pre-modulation filter has significant values, and L_F is the same but for the pre-detection IF filter.

To calculate the error rate performance in the presence of ISI, one has to consider all the possible bit combinations of ISI memory that have a significant effect on the bit under consideration. For example, consider the binary FM signaling case and that the memory extends to one bit after and one bit before the bit under consideration. In this case, all the possible 3 bit combinations of 1's and 0's need to be considered for a reliable estimation of the bit error performance.

3.5 Performance Analysis

Due to the non-linear nature of the system under study and the resulting non-Gaussian noise distribution at the receiver output, mathematical evaluation of the error performance is, in general, very complicated. With hard decision detection it is possible under suitable assumptions to have an approximate expression for the bit error probability especially if the channel is static. The problem gets more complicated for soft decision detection or if

Figure 3-4: Eye pattern at the output of the discriminator detector (a) MSK with Gaussian IF filter of bandwidth $BT=1$ (b) MSK with Gaussian IF filter of bandwidth $BT=0.5$ (c) Duobinary coded MSK with Gaussian IF filter of bandwidth $BT=1$ (d) GMSK with pre-modulation filter B_1T of 0.35 and Gaussian IF filter of bandwidth $BT=1$.

the channel is not static such as the mobile radio channel. However, under those circumstances, we resort to computer simulation for evaluating the bit error rate performance. In this thesis our main concern is with soft decision detection over a fading channel, so most of the error performance results were obtained from simulation. Some analytical results were obtained for the purpose of verifying the validity of simulation results.

3.5.1 Preliminary assumptions

For the digital FM system under study, the following assumptions are made in order to simplify the evaluation of the error rate performance: (1) The modulation index h is restricted to the range $0.5 \leq h \leq 1$, and the reason is that our primary concern in this work is the case of narrow-band digital FM. (2) The IF filter normalized bandwidth BT is confined to the range $1 \leq BT \leq 3$. The case of $BT \geq 1$ is required so that the filter transient response dies away very quickly and thereby reducing the effect of ISI in the received data sequence, moreover under this constraint the noise samples at the output of the filter which are separated by $1/B$ can be regarded as uncorrelated samples. The case of $BT \leq 3$ is required in order for the filter not to pass an excessive amount of out of band noise to the receiver. (3) The I&D filter integrates over an integer multiple of bit duration. In the case of fractional bit integration over less than one bit duration, the noise samples can no longer be regarded as uncorrelated, the incorporation of such correlation makes the analysis more difficult.

3.5.2 Performance with hard decision detection

For MSK signaling with hard decision detection, the decision device after the I&D post-detection filter has a threshold at zero. With reference to equation (3.17), the receiver decides that a "1" was sent if $\Delta\phi(t) > 0$ and "0", otherwise. The decision times are

multiples of the bit duration T . The probability of making an error when a "1" was transmitted is then given by

$$P_e = P \{ \Delta\phi(t) \leq 0 \} \quad (3.30)$$

This probability contains contributions from both continuous and the impulsive (click) component of the discriminator output noise. In the presence of ISI, the calculation of the overall bit error probability requires averaging over various bit patterns. In other words, for the bit error probability to be estimated correctly, one has to use a long data sequence which contains all the possible ISI patterns and then calculate the overall probability by averaging over all these bit patterns. Cartier [4] has noticed that for $BT \geq 1$ the ISI effect is caused by one bit after and one bit before the one under consideration and thereby he was able to reduce the error probability calculations to an average over only four bit patterns. Following [10], the probability of bit error for binary FM is well approximated by

$$P \{ \Delta\phi(t) \leq 0 \} = P \{ \Delta\phi(t) \leq 0 \} + \bar{N} \quad (3.31)$$

where the first part in the right hand side represents the error probability due to the continuous noise and the second part represents the one due to clicks. Using Cartier's approximation, Pawula [8] reduced the calculation of bit error probability in (3.31) to an average over three main pattern namely: 111, 010, and 011 (or 110). In this approximation Pawula considered the 111 pattern as a result of the "all one" bit pattern, and the pattern 010 as result from the "alternating one-zero" bit pattern and finally the 011(or 110) as a result of "all ones" until the middle of the center bit and then the "alternating one-zero" bit pattern. Therefore, the probability of bit error in (3.31) can now be written as

$$P_e = P_{continuous} + P_{click} \quad (3.32)$$

where

$$P_{continuous} = \frac{1}{4} [P\{\Delta\phi \leq 0|111\} + P\{\Delta\phi \leq 0|010\} + 2P\{\Delta\phi \leq 0|011\}] \quad (3.33)$$

and

$$P_{click} = \frac{1}{4} [\bar{N}_{111} + \bar{N}_{010} + 2\bar{N}_{011}] \quad (3.34)$$

where in (3.33) the probability density function, $p\{\Delta\phi\}$, takes the form given in (3.24), and $\bar{N}$ in (3.34) is the average number of clicks given by (3.26). This procedure has been adopted by other researchers for calculating the BER performance of other FM signalling schemes such as Duobinary coded MSK [11].

3.6 Discriminator with Soft Decision Detection

In the presence of ISI, it is well known that the use of maximum-likelihood sequence estimation (MLSE), or soft decision¹ detection, improves the bit error rate performance of the receiver over that of bit-by-bit detection. The basic operation of the MLSE can be described as follows: Due to ISI the ensemble of the signal at the output of the discriminator forms a trellis where the states are defined to be the ISI memory bits $a_{n-1} \ a_{n-2} \ \dots \ a_{n-L-L_F+1}$ and the number of states is given by

$$NS = 2^{(L+L_F-1)} = 2^m \quad (3.35)$$

A brute-force MLSE detector would calculate the likelihood of the received data sequence on all paths through the trellis. The path with the largest likelihood would then be selected

1. In soft decision detection the demodulator operates on the unquantized samples and utilizes the additional information contained in those samples to recover the transmitted data sequence with higher reliability.

and the information bits corresponding to that path would form the detector output. To see how the likelihood of the received data is calculated, suppose we have received a sequence of samples $\bar{x}_N = (x_1, x_2, \dots, x_N)$ and we would like to estimate the transmitted sequence $\bar{a}_N = (a_1, a_2, \dots, a_N)$. Using the knowledge of the noise statistics at the output of the receiver, the MLSE would select the sequence $\bar{a}_N$ which maximizes the joint probability density function (likelihood) given by

$$p(\bar{x}_N | \bar{a}_N) = p(x_N, x_{N-1}, \dots, x_1 | a_N, a_{N-1}, \dots, a_1) \quad (3.36)$$

i.e. a decision is made in favor of the sequence $\bar{a}_N$ with the largest likelihood (which corresponds to the most likely path in the state trellis). If the noisy samples at the receiver output are uncorrelated, the joint probability density function $p(\bar{x}_N | \bar{a}_N)$ can be simplified as the product of the marginal densities, i.e.

$$p(\bar{x}_N | \bar{a}_N) = \prod_{i=1}^N p(x_i | a_i, a_{i-1}, \dots, a_{i-m}) \quad (3.37)$$

where m is the ISI memory span defined by (3.29). A brute-force implementation of the MLSE is usually impractical due to the fact that it requires the calculation of M^N probabilities or metrics, (M is the alphabet size), and therefore the complexity grows quickly as N increases, moreover one has to wait until the entire data sequence becomes available at the receiver output which implies a very long decoding delay.

The maximization of the likelihood function $p(\bar{x}_N | \bar{a}_N)$ can be performed efficiently using a non-linear recursive processor called *Viterbi Algorithm*. By using this algorithm it is possible to greatly reduce the computational complexity suffered by the MLSE by taking advantages of the special structure of the state trellis. A more detailed discussion of Viterbi

algorithm can be found in [34] [46] and [50]. Our primary concern here is how to use the algorithm for improving the error performance of the digital FM limiter-discriminator receiver. In the analysis to follow the Viterbi algorithm will be applied to the output of the I&D post-detection filter and will be called the M-state Viterbi Equalizer¹, in this case the sequence of received samples $\bar{X}_N$ is replaced by $\overline{\Delta\phi_N}$ which represents the sequence of noisy samples at the output of the I&D filter.

The multiplication of probabilities in (3.37) is usually avoided in practice by taking the natural logarithm of both sides of (3.37), i.e.,

$$\ln p(\overline{\Delta\phi_N} | \overline{a_N}) = \sum_{i=1}^N \ln p(\Delta\phi_i | a_i, a_{i-1}, \dots, a_{i-m}) \quad (3.38)$$

In general, upon receiving the signal sample $\Delta\phi_{i+m}$ at the output of the I&D filter the metrics of the algorithm are calculated according to the following recursive relation [35]

$$\mu_i(\bar{a}_{i+m}) = \max_{a_i} [\ln p(\Delta\phi_{i+m} | a_{i+m}, \dots, a_i) + \mu_{i-1}(\bar{a}_{i+m-1})] \quad (3.39)$$

As the next sample is received, the algorithm first computes the M^{m+1} probabilities (metrics)

$$[\ln p(\Delta\phi_{i+m} | a_{i+m}, \dots, a_i) + \mu_{i-1}(\bar{a}_{i+m-1})] \quad (3.40)$$

which correspond to M^{m+1} sequences. These sequences are a continuation of the M^m surviving sequences from the previous stage of the algorithm. Each of the M^{m+1} sequences are then subdivided into M^m groups with each group containing M sequences, which terminate in the same data bits except the last bit (i.e., the same $a_{i+m}, \dots, a_{i+1}$, and

1. Loosely speaking, an "Equalizer" is a signal processing system which is designed to deal with ISI[60].

different a_i). Among each M sequences entering the same state the one having the largest probability is selected as indicated by (3.39), and the other sequences are discarded. Thus at the end of each stage we are left with M^m surviving sequences having the metrics given by (3.39). Therefore, the viterbi algorithm requires calculation of M^{m+1} metrics at each stage, hence a total of NM^{m+1} metrics are computed for the detection of a sequence with length N instead of M^N probabilities using a brute-force MLSE implementation.

As it can be seen from (3.39), the calculation of the metrics depends primarily on the density function of the output noise. In many previous works [18] [21] [22] [23] [25] and [42] the signal samples at the output of the I&D filter are viewed as if they are received in an additive Gaussian noise. Strictly speaking, this assumption is not valid because as we have seen in Section 3.3, the distribution at the I&D filter output is not Gaussian and it contains an impulsive part or clicks. Nevertheless the use of Gaussian approximation has effectively improved the error performance over that of bit-by-bit detector of the FM discriminator receiver. However, for the Viterbi algorithm used in this thesis, we have used the exact distribution which will be called the non-gaussian, as well as the Gaussian approximation to see how the receiver performance compares in the two cases.

3.6.1 Metrics Computation for The Gaussian approximation

The conditional probability density function in (3.39) for the Gaussian noise approximation is given by

$$p(\Delta\varphi_{m+i}) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-(\Delta\varphi_{m+i} - \Delta\theta_{a_{m+i}, \dots, a_i})^2 / 2\sigma^2} \quad (3.41)$$

where σ is the noise variance at the I&D filter output, and $\Delta\theta_{a_{m+i}, \dots, a_i}$ is the output of the I&D filter at the sampling instant in the absence of noise taking into account the effect of

m ISI memory bits. In terms of the eye pattern these values correspond to the different receiving points in the eye pattern at the sampling instant. These values will be referred to as the reference points of the M-state Viterbi equalizer. By taking the natural logarithm of (3.41) and factoring out the parameters which are not function of the input data, we get

$$\ln p(\Delta\phi_{m+i} | a_{m+i}, \dots, a_i) = -(\Delta\phi_{m+i} - \Delta\theta_{a_{m+i}, \dots, a_i})^2 \quad (3.42)$$

3.6.2 Metrics Computation for the Non-Gaussian distribution

Using the exact noise distribution at the output of the I&D filter, the conditional density function in (3.39) is given by

$$p(\Delta\phi_{m+i} | a_{m+i}, \dots, a_i) = \sum_{k=-\infty}^{\infty} p(\Delta\phi_{m+i} + 2\pi k | \Delta\theta_{a_{m+i}, \dots, a_i}) p(N = k) \quad (3.43)$$

In general, there is no simple formula of the above representation, for this reason the metrics of the Viterbi equalizer are calculated numerically when using (3.39). For computational convenience the following assumptions are made: (1) For positive instantaneous frequency most of the clicks occur in the negative side, clicks in the positive side are less likely to occur and therefore they will be neglected [3]. (2) The number of positive (negative) clicks considered are those who have significant probabilities (i.e, the summation in (3.43) is taken over only a finite number of values). (3) At high signal-to-noise ratios the click noise becomes rare and the continuous noise becomes dominant and therefore the conditional density in (3.43) reduces to one term of the form given by (3.24).

3.6.3 Design of the K-state Viterbi Equalizer

In designing the K-state Viterbi equalizer, the selection of the number of states is governed by the amount of ISI memory we have considered. The more ISI memory we account for, the more states we need to use and consequently the more complex the

algorithm becomes. Two types of Viterbi equalizers were used in this study; a 2-state and 4-state Viterbi equalizer. In the 2-state Viterbi equalizer we assume that the received signal sample is determined by two consecutive bits; the previous bit and the one under consideration, i.e. a one bit ISI memory is assumed. One situation where this approximation holds is the Duobinary FM, in this case the receiving signal points of the eye pattern at the sampling instants are clustered around 0 and right below $\pm\pi h$. Another situation is the MSK or GMSK with 2-bit I&D filter, where the signal points are clustered around 0 right below $\pm 2\pi h$. Figure 3.5a shows the eye pattern of Duobinary FM signal at the output of a 1-bit I&D filter. The receiving signal points are labeled R_j ($j=0,1\dots3$). From the figure, the receiving point R_0 is produced from the bit pattern 00, R_1 from 01, R_2 from 10 and R_3 from 11. These ISI rules are depicted as the 2-state trellis diagram shown in Figure 3.6.

In the 4-state viterbi equalizer, we limit the ISI memory to approximately 2 bits. The output received sample is determined by three consecutive bits; the one under consideration and another two neighboring bits. The selection of the other two bits depends on the shape of the channel filters impulse response. For example, using MSK scheme, with a Gaussian pre-detection IF filter of the form shown in figure 3.3 with $BT=1$, the two neighboring bits are the previous and the next bit to the one under consideration (this pattern results from the symmetry of the filter impulse response). For this example, the receiving signal points at the sampling instant are clustered around 8 symmetrical points. Figure 3.5b shows these receiving points on the eye diagram of MSK signal at the output of 1-bit I&D filter, also shown are the bit pattern combination which generates these points. In figure 3.5b, the signal point R_0 is produced from the bit pattern 000, R_1 is from 001, R_2 from 100, R_3 from 101, R_4 from 010, R_5 from 011, R_6 from 110, and R_7 from 111. Since the adjacent past and future bits influence the current bit exactly by the same amount of ISI

Figure 3-5: Examples of eye pattern at the output of I&D filter to demonstrate the dispersion of receiving signal points due to the ISI caused by pulse shaping and IF filtering (a) Duobinary coded MSK signal (b) MSK signal. In both cases the IF filter is Gaussian with $BT=1$.

Figure 3-6: Trellis diagram of a 2-state Viterbi Equalizer

the receiving points R_1 and R_2 has the same position in the eye pattern, also the points R_5 and R_6 have the same position. These ISI rules are depicted as the 4-state trellis diagram shown in Figure 3.7. From the state trellis, the state 00 represents the one that the previous bit is 0 and the current bit is 0 and so on for the rest of the states. Along the trellis diagram are shown the receiving signal points R_j ($j=0,2,\dots,7$), which are used as the reference points of the algorithm in the implementation of the 4-state Viterbi equalizer. The reference points¹ of the 4-state Viterbi equalizer are measured from the eye pattern at the output of the I&D filter and are shown in Table 3-1 for different receiver arrangements.

Table 3-1: Reference points of the 4-state Viterbi equalizer using different receiver arrangements.

Receiver [†] Arrangement	Reference points							
	R_0	R_1	R_2	R_3	R_4	R_5	R_6	R_7
MSK LD+1-bit I&D filter	$-\pi h$	-1.22		$-.873$	$.873$	1.22		πh
	$-\pi h$	-1.3	$-\pi h$	-1.3	1.3	πh	1.3	πh
GMSK LD+1-bit I&D filter	$-\pi h$	-1.07	$-.893$	$-.410$	$.410$	$.893$	1.07	πh
	$-\pi h$	-1.13	-1.05	$-.623$	$.623$	1.05	1.13	πh
GMSK LD+2-bit I&D filter	-2.80	$-.434$	-2.21	$.244$	$-.244$	2.21	$.434$	2.80
	-2.89	$-.183$	-2.43	$.259$	$-.259$	2.43	$.183$	2.89

†. Gaussian IF filter

8th order Butterworth IF filter

1. In case the Viterbi equalizer does not utilize the full ISI memory of the channel filters, the middle of each cluster is considered to be one reference point.

Figure 3-7: Trellis diagram of a 4-state Viterbi Equalizer

4

System Description and Simulation Model

This chapter describes the model of communication system that was used for computer simulation. The simulation was carried out using the SUN-SPARC and HP workstations. A software program has been written in MATLAB based on the functional block diagram of figure 4.1 and 4.2 to study the system performance characteristics under different conditions. Low pass equivalent models have been used to represent all signals and system components. In this chapter, section 4.1 and 4.2 describes the different transmitter and receiver components, their functions and the way they were implemented for simulation. Section 4.3 describes the channel model used to simulate the mobile radio channel.

4.1 Transmitter

4.1.1 Data Source

The binary input data are driven from a Pseudo-random binary sequence generator (Pseudonoise, PN, generator). The PN sequence is generated using a feed back shift register arrangement shown in Figure 4.2a. The output binary sequence is periodic with a

Figure 4-1: FM Transmitter

Figure 4-2: Limiter-discriminator FM Receiver

period $2^m - 1$, where m is the number of shift register stages. The PN sequence has many properties that make it an attractive binary data source for simulating communication systems. It has the following properties [52]: (1) The output sequence of 0's and 1's are statistically independent with occurring probability of 1/2. (2) The number of 1's always exceeds the number of 0's by only one in each period of the sequence. (3) The autocorrelation function of the PN sequence resembles that of a random binary sequence. (4) Every possible pattern of m -bit combination, of a given m -bit shift register, exists and appears only once in each period (The only pattern which doesn't exist is the all 0's pattern). This property is very useful for simulating intersymbol interference (ISI) caused by band limiting filters in digital communication systems, for example, if the ISI memory of the system spans m bits, we need to run the system with a binary sequence that has all the possible m -bit combinations.

In the simulation a program similar to [52] is used to generate a binary Non-Return to Zero (NRZ) PN sequence with a selectable period that can go up to 1023 (i.e. an ISI memory of up to 10 bits). The number of transmitted bits can go up to 10^6 . Figure 4.3b shows a sample PN sequence with a period length of 63 bits (i.e., $m=6$).

4.1.2 Differential Encoder

Binary differential encoding is the process of mapping the input binary sequence $\{a_k\}$ into a new binary sequence $\{b_k\}$ which conveys the same information. The mapping process is achieved by using the current source bit and the previous output bit of the encoder according to the following equation

$$b_k = a_k \oplus b_{k-1} \quad (4.1)$$

where $\oplus$ indicates modulo-2 addition.

(a)

(b)

Figure 4-3: PN sequence Generation (a) Shift register arrangement (b) sample output of period 63 ($m=6$)

The differential encoder is used to avoid error propagation problem. This problem is usually encountered when the detector decision is made based on the previously detected bit and the current bit. An example, of this detection techniques is the case when we have a three-level eye pattern at the decision instant (e.g., Duobinary MSK and 1/2-bit offset GMSK signals).

4.1.3 Correlative encoder

Correlative encoding is the process of mapping an M-ary sequence, $\{b_k\}$, into a Q-ary sequence, $\{c_k\}$, for some desirable properties especially in the frequency domain. The correlative encoder operation is usually described by the following characteristic polynomial

$$F(D) = \sum_{i=0}^l g_i d^{l-i} \quad (4.2)$$

where d is the delay variable which represents a T -seconds delay and the polynomial coefficients, g_i 's, are such that $\sum_{i=0}^l g_i = 1$. The output sequence of the encoder, $\{c_k\}$, is given by

$$c_k = \sum_{i=0}^l g_i b_{k-i} \quad (4.3)$$

For MSK

$$g_i = \begin{cases} 1 & \text{for } i = 0 \\ 0 & \text{otherwise} \end{cases} \quad (4.4)$$

In this case

$$c_k = b_k \quad (4.5)$$

For the Duobinary MSK

$$g_i = \begin{cases} \frac{1}{2} & \text{for } i = 0, 1 \\ 0 & \text{otherwise} \end{cases} \quad (4.6)$$

in this case we have

$$c_k = \frac{b_k + b_{k-1}}{2} \quad (4.7)$$

The two encoder in Figure 4.1 are shown in a shaded blocks to indicate that they are not always used.

4.1.4 Pre-modulation Filter

The purpose of the pre-modulation filter (or the pulse shaping filter), $g(t)$, is to shape the input NRZ data waveform before application to the FM modulator. Two types of pre-modulation filters are used in the simulation; the *rectangular* filter with an impulse response given by

$$h_t(t) = \begin{cases} \frac{1}{2T} & |t| \leq T/2 \\ 0 & \text{otherwise} \end{cases} \quad (4.8)$$

where T is the bit duration. This filter is used for the generation of the MSK signal. The other filter type used is the *Gaussian* filter with an impulse response given by

$$g(t) = \frac{1}{2} \operatorname{erf} \left(\pi B_f T \sqrt{\left[\frac{2}{\ln 2} \right]} \left(\frac{t}{T} + 0.5 \right) \right) - \frac{1}{2} \operatorname{erf} \left(\pi B_f T \sqrt{\left[\frac{2}{\ln 2} \right]} \left(\frac{t}{T} - 0.5 \right) \right) \quad (4.9)$$

where $\operatorname{erf}()$ is the error function and B_f is the filter 3-dB bandwidth. Figure 4.3 shows the impulse response of the Gaussian pre-modulation filter with Bandwidth-Time product, $B_f T$, as a parameter. The Gaussian filter is used for the generation of the Gaussian filtered Minimum Shift Keying signal (GMSK).

The impulse response of the Gaussian filter will be considered to be nonzero only over a 4 bits interval. The value of $g(t)$ is negligible outside this interval for the range of $B_f T$ product we considered. The filter is implemented as 64 taps linear phase FIR filter.

4.1.5 FM Modulator

The FM modulator accepts, as its input, the filtered NRZ data, $d(t)$, from the pre-modulation filter and produces an output sinusoidal signal, $s(t)$, with a constant envelope and a phase function varying in proportion to $d(t)$. The output signal $s(t)$ is given by

$$s(t) = \operatorname{Re} \{ A e^{j(\omega_c t + \Phi(t))} \} \quad (4.10)$$

where A is the signal amplitude and $\Phi(t)$ is the modulating phase given by

$$\Phi(t) = \frac{\pi h}{T} \sum_{n=-\infty}^{\infty} c_n \int_{-\infty}^t g(t-nT) dt \quad (4.11)$$

where h is the modulation index as defined in chapter 2, and c_n is the input NRZ bit stream (if the differential encoders are not used then c_n is replaced by a_n).

The simulated FM signal is implemented by first generating the phase function $\Phi(t)$ recursively using (4.11) (the initial phase of the signal is assumed to be zero), and since

Figure 4-4: Impulse response of a Gaussian pre-modulation filter with B_1T as a parameter

low-pass equivalent model is used in the simulation, Equation (4.5) is implemented as a complex quadrature envelope according to the following equation

$$\hat{s}(t) = A \cos\Phi(t) + jA \sin\Phi(t) \quad (4.12)$$

4.2 Receiver

4.2.1 IF Pre-detection Filter

The primary function of the IF filtering is to select the desired signal and to minimize the effect of noise and interference from adjacent channels. In the design of digital FM systems IF filtering plays an important role in determining the error rate performance of the system. Two conflicting requirements are usually encountered in designing such filters. A narrow bandwidth is required for noise limiting, while a wide bandwidth is desirable for minimizing the signal distortion (ISI).

Two types of filters are used; Gaussian and Butterworth. The Gaussian filter has an impulse response defined as

$$h(t) = B\sqrt{2}e^{(-2\pi B^2 t^2)} \quad (4.13)$$

where B is the 3-dB bandwidth of the filter (the frequency response of a Gaussian filter is also Gaussian). The duration of the impulse response of this filter is considered to be non zero only over 4 bits and is implemented as a 64-tap linear phase FIR filter.

The second filter type considered is an 8th order Butterworth filter which has a maximally flat magnitude response. The filter is designed using the standard IIR filter design procedure outlined in [48]. The filter realization is the transposed direct form II. Figure 4.5 and Figure 4.6 show the impulse response and the magnitude response of the Gaussian and Butterworth filter both with normalized 3-dB bandwidth of 1.

Figure 4-5: Low-pass equivalent of a Gaussian IF filter with $BT=1$ (a) Impulse response (b) Magnitude response

Figure 4-6: Low-pass equivalent of an 8th order Butterworth IF Filter with $BT=1$ (a) Impulse response (b) Magnitude response

4.2.2 Limiter-Discriminator

The function of the limiter, as it is defined in chapter 2, is to deliver a constant amplitude signal to the discriminator input, however it does not alter the phase of the carrier. In the simulation, the limiter function is modeled by dividing the complex signal at the output of the IF filter by its absolute value; the resulting output signal has a unity amplitude.

Following the limiter is the discriminator which delivers an output signal, $r(t)$, proportional to the time derivative of the phase function, $\Phi(t)$, of its input signal (the instantaneous frequency), i.e.;

$$r(t) = \frac{d}{dt} \Phi(t) \quad (4.14)$$

In the simulation the discriminator is modeled by the following equation

$$r(t) = I(t) \frac{d}{dt} Q(t) + Q(t) \frac{d}{dt} I(t) \quad (4.15)$$

where $I(t)$ and $Q(t)$ are, respectively, the real and imaginary components of the complex signal at the output of the limiter. Since the differentiation is carried out using difference between adjacent samples, the number of these samples must be large enough to minimize the computation errors.

4.2.3 Post-detection Filter

The primary function of the low-pass post-detection filter is to compensate for the undesirable enhancement of the noise at the band edges caused by the discriminator (the process of differentiation is the main cause of such noise enhancement). Throughout the simulation the post detection filter is implemented as a simple integrate and dump (I&D)

filter, where the input signal is integrated over some interval and at the end of the integration period the output is dumped to zero. Depending on the integration period, two filter types are investigated in the simulation; these are the one bit and the two bits I&D filter. The response of the one bit I&D filter at any time t is the integration of its input over one bit duration according to the following equation

$$x(t) = \int_{(n-1/2)T}^{(n+1/2)T} r(t) dt \quad (4.16)$$

The output signal, $x(t)$, is then sampled every T sec and applied to the detection device (note that the output in this case represents a phase change over a one bit period). Figure 4.5a shows a block diagram of a one bit I&D post-detection filter arrangement. The two bits I&D filter is implemented by applying the output of the discriminator to a bank of two I&D filters with integration window over the interval $(n-1)T \leq t \leq (n+1)T$ and $(n-2)T \leq t \leq nT$, i.e., each integration window is two bits in duration (a phase change over two bits period). The output signal of the filter bank is then sequentially sampled and applied to the detection device as shown in Figure 4.5b.

4.2.4 Detection Device

The final stage of the receiver is the detection device and its main function is to estimate the transmitted data sequence with the least possible errors. Both hard and soft decision detection were used in the simulation. The sampling clock is assumed to be generated from an ideal clock recovery circuit.

4.2.4.1 Bit-by-bit Detection

A bit-by-bit, or hard decision¹ detector, decides whether a received data bit is a 0 or 1

Figure 4-7: Post-detection filter arrangement (a) 1-bit I&D filter (b) 2-bit I&D filter

according to whether the data signal is above or below a decision threshold. With the aid of the eye pattern at the output of the post-detection filter, the decision threshold is set according to the number of levels of the eye pattern at the decision instant (i.e., the time when the eye is open). If the eye pattern has a 2-level opening, in this case the output sequence, x_k , is passed through a simple binary threshold comparison device with a threshold setting at 0. The decision is made as follows

$$\hat{a}_k = \begin{cases} 1 & \text{if } x_k > 0 \\ 0 & \text{if } x_k \leq 0 \end{cases} \quad (4.17)$$

If the eye pattern has a 3-level opening, the data are first differentially encoded at the transmitter and x_k is then passed through a full wave rectifier and then through a binary threshold device. The threshold setting in this case is $\alpha\pi h$ for the one bit I&D filter and $2\alpha\pi h$ for the two bit I&D filter, where the parameter α is an adjustment factor selected to yield a minimum probability of error and h is the modulation index. However, in any of the above cases the decision is made as follows

$$\hat{a}_k = \begin{cases} 1 & \text{if } |x_k| < \text{threshold} \\ 0 & \text{if } |x_k| \geq \text{threshold} \end{cases} \quad (4.18)$$

4.2.4.2 Maximum-Likelihood Sequence Estimation MLSE

The MLSE is used as a soft decision detection technique to enhance the performance of the limiter-discriminator receiver in the presence of ISI. The 2-state and 4-state Viterbi algorithm described in chapter 3 is used for the implementation of the MLSE. The algorithm was implemented as a software program which is described by the flow chart

1. In hard decision detection the demodulator makes a firm decision on each received sample by passing it through an L-level quantizer, where L is the number of signaling levels.

Figure 4-8: Flow chart showing the implementation of MLSE Equalizer with Viterbi algorithm

shown in figure 4.8. In the initialization process an arbitrary¹ state is chosen as a starting state and the algorithm reference points are set to their values depending on the number of states and the type of FM signal. The program traverses through the state trellis for N times which represents the pass memory size (or the path length). The pass memory size for the Viterbi algorithm was set to 256 bits which seems to be sufficiently long compared to the ISI memory size. After each N loops the surviving sequence is traced back and the algorithm outputs an estimate of the transmitted data bits. These estimated data are then compared with the original transmitted data for estimating the bit error rate.

4.3 Channel Model

For simulating the additive white Gaussian noise (AWGN) channel, a random number with a Gaussian distribution is added to each signal sample at the input of the receiver (IF filter). For estimating the BER performance of the system as a function of signal-to-noise ratio, E_b/N_0 , in the simulation, the mean signal level is held constant while the added noise power is changed. Cochannel interference is simulated by passing an FM signal with different PN sequence through a simulated channel and adding it to the desired signal at the input of the receiver.

The channel model of the flat fading and the selective fading used in the simulation is the two-ray channel model. The discrete impulse response of this model is given by

$$h_c(kT_s) = i_1(kT_s) \delta(kT_s) + i_2(kT_s) \delta(kT_s - T_{ds}) \quad (4.19)$$

where the channel coefficients, i_1 and i_2 , are two independent complex gaussian random variables, T_s is the fractional symbol sampling period and T_{ds} is the time delay between

1. It has been shown through simulation [51] that there is a high probability that all the surviving sequences will merge with the correct path, thus the algorithm synchronizes it self without any special procedure.

the two rays. The power sum $E\{|i_1|^2\} + E\{|i_2|^2\}$ (where E indicates the expected value) is made equal to 1, so that the channel has a unity gain throughout the simulation runs. The flat fading channel is generated by setting i_2 to zero. For the frequency selective fading channel the power ratio $E\{|i_1|^2\}$ to $E\{|i_2|^2\}$ will be denoted as C/D , where C is the main ray and D is the delayed ray. The channel coefficients, i_1 and i_2 , are generated using Jakes' sum of nine sinusoids technique[36]. The resulting Doppler power spectrum is a discrete approximation of

$$S_c(f) = \left[1 - \left(\frac{f}{f_m}\right)^2\right]^{-\frac{1}{2}} \quad (4.20)$$

where f_m is the maximum Doppler frequency of the channel. A block diagram of the experimental channel model is shown in Figure 4.8 and a sample realization of a Rayleigh channel envelope is shown in Figure 4. 9.

Figure 4-9: Channel Model used in the simulation

Figure 4-10: Simulated Rayleigh channel (a) power envelope (b) Histogram of the Rayleigh Amplitude (c) Histogram of the Rayleigh power level

5

Experimental Results

In this chapter, we will present the results obtained from the computer simulation of the system described in chapter 4. In this work we have considered three CPM signaling schemes that have become very attractive for mobile communications applications [1] [2] [11] and [30]; these are MSK, Duobinary coded MSK and GMSK, where the MSK signaling represents an example of full response CPM signaling and the other two represent an example of partial response CPM signaling.

5.1 BER Estimation and Simulation Accuracy

In analog transmission systems the criterion of signal-to-noise ratio is usually adequate as a performance measure. In digital data transmission this criterion by itself does not provide an adequate performance judgement, the performance is usually judged in terms of the number of bits in error (commonly referred to as *bit error rate*, or *BER* performance). This performance measure will be solely used in this thesis.

The estimation of the BER performance is achieved using the conventional Monte Carlo error counting technique. In order for the average BER estimation to be reliable, one has to run the simulation over a considerable amount of time, or in other words one has to

run the simulation with sufficiently long data sequences. In order to save the simulation time, the length, L , of the data sequence is estimated using the well known rule-of-thumb [57] given by

$$L \equiv \frac{100}{p_e} \quad (4.1)$$

where p_e is the bit error probability. To implement this rule-of-thumb in the computer simulation program, the error probability for the AWGN channel is reported after the occurrence of at least 100 errors. In mobile communication environment care must be taken in calculating the BER performance using this technique. For example, in a slowly fading channel, a sudden deep fade in the channel might generate numerous bit errors and causes the simulation to terminate prematurely before the mobile experience a sufficient number of Doppler periods, especially at low SNR. To avoid this situation we have to average the BER performance over a longer span of channel response in order to approximate the Rayleigh statistics more accurately¹. In terms of simulation, this implies counting more errors than the AWGN channel case before reporting the BER performance so that the program cycles through an adequate length of channel response before termination.

To see how accurate the simulation results are, one of the BER measurements was run repeatedly using different seeds for the sequence generator used to create the transmitted data sequence and different seeds for the Gaussian noise generator. After 15 runs, it was found that the difference in the required SNR to achieve an average BER of 10^{-4} is approximately ± 0.2 dB. Moreover some published results were reproduced by our simulation program and our results compare very favorably with those published ones as it will be shown later.

1. In other words the longer the channel response, the better averaging we get, hence more reliable results.

5.2 Performance using HD detection over an AWGN channel

The average BER performance of the three CPM signaling schemes was investigated with HD detection following the post-detection I&D filter. The IF filters used are the Gaussian and the 8th order Butterworth filter. Although some of the BER performance results using the Gaussian IF filter were previously produced by other researchers, it has been reproduced here for simulation verifications and comparison. The Butterworth or maximally flat IIR filter was selected because of its flat pass band response and the sharp cut-off frequency. Note that this filter does not exhibit a linear phase characteristics (most of the IF filters studied in the literatures were assumed to have a linear phase characteristics). This would allow a demonstration of the effects of a nonlinear phase response on the performance of digital FM receivers. The Butterworth filter also represents an example of a more practical filter design than the Gaussian filter.

5.2.1 Error rate with 1-bit I&D post-detection filter

Through the computer simulation, we have found that a BT product of 1 provides an excellent compromise between signal distortion (ISI) and noise rejection for both filters at any value of $h < 1$ for MSK signalling. The BER performance curves were generated and are shown in Figure 5.1 and 5.2 for the Gaussian and 8th order Butterworth IF filters, respectively. Both filters have a normalized bandwidth BT of 1. The post-detection filter is 1-bit I&D filter. From the figures, it is evident that the performance of the 8th order Butterworth filter is superior to that of the Gaussian filter. The modulation index was used as a parameter, and it can be seen, the best performance is achieved at an index of 0.7 for both filters. From the figures, the required SNR to achieve an average BER performance of 10^{-4} for $h=0.7$ is 10.8 dB for the Gaussian filter and 9.9 dB for the 8th order Butterworth

Figure 5-1: BER performance of MSK signaling using 1-bit I&D filter and HD detection. The IF filter is Gaussian with $BT=1$.

Figure 5-2: BER performance of MSK signaling using 1-bit I&D filter and HD detection. The IF filter is 8th order Butterworth with $BT=1$.

filter, i.e., an improvement of approximately 1 dB in using the Butterworth filter. It is also interesting to see that the required SNR to achieve an average BER of 10^{-4} is the same for $h=0.5$ and $h=1$ when using the Gaussian IF filter and it is approximately 12.5 dB with the performance at $h=1$ starting to improve, whereas for the 8th order Butterworth filter, the SNR is 10 dB at $h=0.5$ and 11.8 dB at $h=1$ for the same BER performance. The reason for the for the BER performance superiority of $h=0.5$ over that of $h=1$ is that the BT product of 1 is not optimum for $h=1$. This justification is confirmed by looking at the BER performance of MSK at an index $h=1$ and $BT=1.25$ in Figures 5.1 and 5.2, which provides almost the same performance as MSK with $h=0.5$ and $BT=1$ (with the performance of $h=1$ being slightly better at high SNR).

The improvement of the MSK signaling performance when using the 8th order Butterworth filter over that with the Gaussian filter can be more visualized with the aid of eye diagram. Figure 5.3a and 5.3b show simulated eye patterns of an MSK signal at the output of 1-bit I&D filter for the two IF filters used, both with BT product of 1. The eye opening at the sampling instant for the case of 8th order Butterworth filter provides more noise margin (approximately 1.8 dB) compared to the Gaussian IF filter case. This improvement in the eye opening agrees favorably with the difference in the BER performance curves of the two filters, where the Butterworth filter is approximately 2 dB superior to the Gaussian filter at high SNR.

The BER performance for the Duobinary coded MSK signaling are shown in Figures 5.4 and 5.5 for the Gaussian and the 8th order Butterworth IF filters, respectively. It can be seen here also that the 8th order Butterworth filter exhibits a better BER performance than the Gaussian filter for all modulation indices used. From the figures, the required SNR to achieve an average BER of 10^{-4} with an index $h=0.7$ and a Gaussian filter is approximately

Figure 5-3: The eye pattern at the output of 1-bit I&D filter (a) MSK with Gaussian IF filter (b) MSK with 8th order Butterworth IF filter (c) GMSK with Gaussian IF filter (d) GMSK with 8th order Butterworth IF filter. The IF filters' BT product is 1 and $h=0.5$.

Figure 5-4: BER performance of Duobinary FM signaling using 1-bit I&D filter and HD detection. The IF filter is Gaussian with $BT=1$

Figure 5-5: BLR performance of Duobinary FM signaling using 1-bit I&D filter and HD detection. The IF filter is 8th order Butterworth with $BT=1$

14.3 dB which is 3.5 dB higher than the MSK signaling. Using an 8th order Butterworth filter the required SNR to achieve the same BER is approximately 12.5 dB which is 2.5 dB higher than the MSK and 1.8 dB better than the Duobinary coded MSK with the Gaussian filter. The degradation in the performance of Duobinary MSK compared to the MSK results from the use of correlative encoding which introduces more ISI and provides a signal with more compact spectrum.

Figures 5.6 and 5.7 show the BER performance of GMSK signaling using Gaussian and 8th order Butterworth IF filters, respectively, at an index $h=0.5$ with a pre-modulation filter $B_T T$ product as a parameter. Performance curves are shown for both 2-level and 3-level eye opening HD detection. In the 3-level eye detection the integration period is offset by $T/2$, where T is the bit duration, i.e., the I&D filter integrates over the interval $[t-T, t]$ instead of $[t-T/2, t+T/2]$, and the resulting eye pattern has a 3-level opening at the decision instant which can have an eye opening wider than the 2-level eye depending on the pre-modulation filter $B_T T$ product and the IF filter bandwidth. In the case of 3-level eye detection differential encoding of the input data is used to avoid the problem of error propagation. From the figures we first notice again that the 8th order Butterworth filter provides a superior performance compared to the Gaussian filter for both detection techniques. We also notice that the 3-level eye HD detection with a pre-modulation filter $B_T T$ product of 0.35 and a Butterworth filter has an inferior performance compared to the 2-level eye detection and the reason behind this is that for $B_T T$ product of 0.35, the 2-level eye has a wider opening at the decision instant than the 3-level eye. In contrast with the duobinary coded MSK, at an index $h=0.5$ the required SNR to achieve an average BER of 10^{-3} with $B_T T$ product of 0.35 and 2-level eye detection is approximately 12.5 dB for the Butterworth IF filter which is approximately 1 dB better than the Duobinary coded MSK.

Figure 5-6: BER performance of GMSK ($h=0.5$) signaling using 1-bit I&D filter and HD detection. The IF filter is Gaussian with $BT=1$

Figure 5-7: BER performance of GMSK ($h=0.5$) signaling using 1-bit I&D filter and HD detection. The IF filter is 8th order Butterworth with $RT=1$

This suggests that if the system constraint is to operate at an index $h=0.5$, GMSK with B_fT product of 0.35 will be the choice over Duobinary coded MSK not only in terms of BER performance but also in terms of spectrum efficiency. The eye pattern at the output of the 1-bit I&D filter is shown in figure 5.3c and 5.3d for a GMSK signal with an index $h=0.5$ and a B_fT product of 0.35 using both types of IF filters, which again demonstrates the improved eye opening in the case of Butterworth filter.

5.2.2 Error rate with 2-bit I&D post-detection filter

The BER performance of MSK signaling at an index $h=0.5$, using a 2-bit I&D post-detection filter and HD detection is shown in Figure 5.8 for the Gaussian and 8th order Butterworth IF filters. The improvement in the BER performance with the 2-bit I&D filter over that of 1-bit I&D filter is quite evident with the Butterworth filter performance being superior to the Gaussian filter. The required SNR to achieve an average BER of 10^{-4} is 10 dB with the 8th order Butterworth filter (i.e., an improvement of 0.8 dB over that of 1-bit I&D filter), and 11.8 dB with the Gaussian filter (i.e., an improvement of 0.7 dB over that of 1-bit I&D filter). For comparison, the BER performance of MSK with 2-bit differential detection from [17] is also shown in the figure. The eye pattern of an MSK signal with $h=0.5$ at the output of 2-bit I&D filter is shown in Figure 5.9a and 5.9b using Gaussian and Butterworth IF filters.

The BER performance of GMSK signaling at an index $h=0.5$ is plotted in Figure 5.9 and 5.10 using a pre-modulation filter B_fT product of 0.25 and 0.35 respectively, which also confirms the superiority of the 2-bit I&D filter over the 1-bit I&D filter and that the Butterworth filter still exhibit a superior performance even though it has a nonlinear phase characteristics. For example, in figure 5.11 the required SNR to achieve an average BER of 10^{-3} is 10.5 dB using a Butterworth IF filter, which is 2 dB better than the 1-bit I&D filter.

Figure 5-8: BER performance of MSK signaling using 2-bit differential filter and HD detection.

2 of/de 2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

Figure 5-9: The eye pattern at the output of 2-bit I&D filter (a) MSK with Gaussian IF filter (b) MSK with 8th order Butterworth IF filter (c) GMSK with Gaussian IF filter (d) GMSK with 8th order Butterworth IF filter. The IF filters' BT product is 1 and $h=0.5$.

Figure 5-10: BER performance of GMSK signaling ($B_1T=0.25, h=0.5$) using 2-bit I&D filter and HD detection.

Figure 5-11: BER performance of GMSK ($B_1T=0.35, h=0.5$) signaling using 2-bit I&D filter and HD detection.

For the Gaussian filter the required SNR to achieve the same BER is 12.5 dB, which is approximately 3.8 dB better than the 1-bit I&D filter.

The performance improvement of the 2-bit I&D filter over the 1-bit I&D filter can be justified as follows; the process of differentiation enhances the high frequency contents of the discriminator input signal with the majority of these frequency contents being noise. The function of the I&D filter is to compensate for these undesirable noise enhancement. The degree of this compensation is proportional to the length of the integration interval; where the longer the integration period the smoother the I&D filter output. In terms of the eye pattern, the operation of 2-bit I&D filter provides an eye opening with a larger noise margin than that of the 1-bit I&D filter. Figure 5.9c and 5.9d show the eye pattern of a GMSK signal at the output of 2-bit I&D filter with a pre-modulation filter B_T product of 0.35 and an index $h=0.5$. Since the eye pattern has a 3-level eye opening at the sampling instant, differential encoding of the transmitted data sequence is used to avoid error propagation

5.3 Performance using SD detection over an AWGN channel

The error rate performance of MSK, Duobinary coded MSK and GMSK is evaluated for the discriminator receiver with SD following the post-detection I&D filter. The SD detector was implemented as an M-state Viterbi equalizer as described in Chapter 3. The performance is evaluated using the approximation that the noise at the I&D filter is Gaussian [18] [21] [23], and in this case the metrics for the Viterbi algorithm are calculated using the Euclidean distance. The performance is also evaluated using the exact or the non-Gaussian noise distribution which contains the clicks and in this case the metrics are calculated numerically as mentioned in Chapter 3.

5.3.1 Error rate with 1-bit I&D filter

The BER performance of Duobinary coded MSK signalling at an index $h=0.5$ and employing a 2-state Viterbi Equalizer is shown in Figures 5.12 and 5.13 with Gaussian and 8th order Butterworth IF filters, respectively. In each figure, the BER is plotted for both the Gaussian output noise approximation and the non-Gaussian noise distribution. From the figures, we notice that the BER performance has been improved with SD detection compared to HD detection regardless of the noise distribution at the output of the post-detection I&D filter. The improvement in using SD detection is approximately 2 dB at an average BER of 10^{-4} using SD detection with both IF filters, with the 8th order Butterworth filter being better than the Gaussian IF filter by approximately 1.5 dB. An interesting observation from the figures is that the difference in the BER performance of the 2-state Viterbi equalizer based on the Gaussian and non-gaussian output noise assumption is very small in the range of interest (i.e. an error rate below 10^{-2}). This difference is almost negligible at a high SNR value.

For GMSK signalling, if the sampling instant is offset by $T/2$ (the same as the 3-level eye detection technique) the dependency in the input data sequence can be assumed to be between the current bit and the previous bit. In this case, the 2-state Viterbi equalizer can be applied successfully for the detection of GMSK signal. Since the resulting eye pattern has a 3-level eye, no differential encoding is required in this case because the data are detected using sequence estimation.

Figures 5.14 and 5.15 show the BER performance of a GMSK signalling at an index $h=0.5$ and a pre-modulation filter B_1T product of 0.35, employing a 2-state Viterbi equalizer. The improvement in BER compared to the HD detection is very evident with the Butterworth filter still maintaining the superiority over the Gaussian filter. It is also seen that the relative position of the error rate curves based on the Gaussian and the non-Gaussian

Figure 5-12: Error Rate performance of Duobinary FM signaling ($h=0.7$) with Gaussian IF filter and SD detection over an AWGN channel.

Figure 5-13: Error Rate performance of Duobinary FM signaling ($h=0.7$) with 5th order Butterworth IF filter and SD detection over an AWGN channel.

Figure 5-14: Error Rate performance of GMSK signaling ($B_1T=0.35$, $h=0.5$) with a Gaussian IF filter and SD detection over an AWGN channel.

Figure 5-15: Error Rate performance of GMSK signaling ($B_1T=0.35$, $h=0.5$) with an 8th order Butterworth IF filter and SD detection over an AWGN channel.

assumption is still the same as it is the case of Duobinary coded MSK, i.e., the performance is almost the same for both cases in the range of interest. In Table 5-1 we show the required SNR in dB to obtain an average BER performance of 10^{-4} with the 2-state Viterbi equalizer based on the non-Gaussian output noise assumption.

Table 5-1: Required SNR to achieve an average BER of 10^{-4} with the 2-state Equalizer

Signalling scheme	IF Filter Type(BT=1)	SNR
Duobinary MSK $h=0.5$	Gaussian	14.1 dB
	8 th order Butterworth	12.5 dB
GMSK $h=0.5$ $B_1T=0.35$	Gaussian	13.5 dB
	8 th order Butterworth	12.2 dB

A 4-state Viterbi equalizer was also used at the output of 1-bit I&D filter. The trellis diagram of such equalizer is shown in Figure 3.7 of Chapter 3. As it was the case with the 2-state equalizer, the performance has been examined with the Gaussian and the non-Gaussian output noise assumptions. This type of equalizer was used for the detection of MSK and GMSK signals. The BER performance of these two signaling schemes at an index $h=0.5$, using the 4-state Viterbi equalizer, are shown in Figures 5.16 through 5. 19. The GMSK pre-modulation filter B_1T product used is 0.35. The improvement in the BER performance is quite evident compared to the HD detection techniques. Moreover the 4-state equalizer performs better than the 2-state equalizer, the reason for this is that the equalizer is taking care of more ISI memory. We also observe that the Viterbi equalizer error rate curves are almost the same for the two noise assumption. In Table 5-2 we show the required SNR to achieve an average BER of 10^{-4} with the 4-state Viterbi equalizer based on the non-Gaussian output noise assumption.

From the error rate performance curves discussed in the previous paragraphs we conclude that regardless of the signalling scheme or the equalizer complexity, the relative

Figure 5-16: Error Rate performance of MSK signaling with Gaussian IF filter and SD detection over an AWGN channel.

Figure 5-17: Error Rate performance of MSK with 8th order Butterworth IF filter and SD detection over an AWGN channel.

Figure 5-18: Error Rate performance of GMSK signaling ($B,T=0.35$, $h=0.5$) with a Gaussian IF filter and SD detection over an AWGN channel.

Figure 5-19: Error Rate performance of GMSK signaling ($B,T=0.35$, $h=0.5$) with an 8th order Butterworth IF filter and SD detection over an AWGN channel.

position of the error rate curves, based on the Gaussian and the non-Gaussian output noise, is the same, i.e. the performance is almost the same in the range of interest with the performance based on the non-Gaussian assumption being slightly better at low SNR. The justification of this situation is that as the SNR decreases the average number of clicks increases at the output of the I&D filter, for a fixed value of h , and therefore the noise becomes more and more impulsive in nature, and as a consequence the Viterbi equalizer based on the Gaussian assumption is not optimum in this environment, hence the poorer performance. As the SNR progresses towards the higher values the clicks start vanishing and the continuous noise part starts dominating the output of the I&D filter with its Gaussian like distribution, this leads to a small difference in the BER performance using both noise assumptions.

Table 5-2: Required SNR to achieve an average BER of 10^{-4} with the 4-state equalizer

Signalling scheme	IF Filter Type(BT=1)	SNR
MSK $h=0.5$	Gaussian	10.5 dB
	8 th order Butterworth	9.9dB
GMSK $h=0.5$ $B_1T=0.35$	Gaussian	11.7 dB
	8 th order Butterworth	10.9 dB

5.3.2 Error Rate with 2-bit I&D filter.

In section 5.2.2, we have seen how the BER performance of the discriminator receiver has been improved when using a 2-bit I&D filter, an additional improvement can be achieved when using Viterbi equalization at the output of the 2-bit I&D filter. Both 2- and 4-state Viterbi equalizers were used for the detection of MSK and GMSK signals. The performance was evaluated only for the non-gaussian output noise assumption, since the difference between the two assumptions is almost negligible in the range of interest. The resulting BER performance for MSK at an index $h=0.5$ is shown in Figure 5.20. For

Figure 5-20: BER performance of MSK signaling, using 2-bit I&D filter and SD detection.

comparison, the BER performance of coherent MSK signalling is also shown in the figure. As one can see, the BER performance with the discriminator detection has been brought to approximately 10.5 dB at an average BER of 10^{-5} and using an 8th order butterworth IF filter, which is 1.5 dB worse than the coherent MSK signalling (about 1 dB from differentially coherent MSK).

Table 5-3: Required SNR to achieve an average BER of 10^{-5} using 2-bit I&D filter and SD detection

Scheme	IF Filter Type (BT=1)	SNR
MSK $h=0.5$ (2-state Viterbi Equalizer)	Gaussian	10 dB
	8 th order Butterworth	9.1 dB
GMSK $h=0.5$, $B_1T=0.35$ (2-state Viterbi Equalizer)	Gaussian	12.5 dB
	8 th order Butterworth	11.5 dB
GMSK $h=0.5$, $B_1T=0.35$ (4-state Viterbi Equalizer)	Gaussian	11 dB
	8 th order Butterworth	10.5 dB

The BER performance of GMSK signalling at an index $h=0.5$ is shown in Figures 5.21 and 5.22 using the 2-state and the 4-state viterbi equalizers following the 2-bit I&D post-detection filter, respectively. From the figures, it can be seen that the BER performance of GMSK with the 2-bit I&D filter and SD detection outperforms the one with 1-bit I&D filter and SD detection. Compared to the coherent MSK for example, with the 4-state Viterbi equalizer, the performance of GMSK signalling is now 3.3 dB away from the coherent MSK when using a Gaussian IF filter and 2.5 dB when using an 8th order nonlinear phase Butterworth IF filter, both at an average BER performance of 10^{-5} . Table 5-3 shows the required SNR in dB to achieve an average BER of 10^{-4} using MSK and GMSK signalling

Figure 5-21: BER performance of GMSK signaling ($B_1T=0.35, h=0.5$), using 2-bit I&D filter and SD detection. The IF Filter is Gaussian.

Figure 5-22: BER performance of GMSK signaling ($B_1T=0.35, h=0.5$), using 2-bit I&D filter and SD detection. The IF filter is 8th order Butterworth.

and employing Viterbi equalizer following the 2-bit I&D post-detection filter.

5.4 Performance over a Mobile Communication channel

In this section, we shall present the BER performance results when using the discriminator receiver with SD detection over a Mobile communication channel. This section considers only GMSK¹ signalling scheme. The modulation index used is 0.5 and the pre-modulation filter B_rT product is 0.35. The IF filter used is an 8th order Butterworth filter because of it's superior performance over the Gaussian filter. The mobile channel impairments investigated are; cochannel interference, Flat and frequency selective Rayleigh fading.

5.4.1 Performance in the presence of co-channel interference

For efficient utilization of the frequency spectrum in the mobile communication environment the carrier frequency is reused in spatially separated cells. The performance of the limiter-discriminator receiver of GMSK signal employing SD detection was investigated in the presence of co-channel interference and AWGN. This interference was simulated as we have mentioned in section 4.3 of chapter 4. The case studied here is the one with a single cochannel interferer. The resulting BER performance curves are shown in Figure 5.23. As it is evident from the figure, the BER performance of the two SD detection techniques are far superior to the HD detection. From the BER curves, the GMSK with 2-bit I&D filter followed by a 4-state Viterbi equalizer is approximately 1.2 dB superior in SNR compared to GMSK with 1-bit I&D filter followed by a 4-state Viterbi equalizer, both at an average BER performance of 10^{-4} .

1. This decision was made because of the simulation time, for the fading channel case, is significantly longer by comparison to the AWGN channel case.

Figure 5-23: Performance of GMSK signaling ($h=0.5$, $B_fT=0.35$), using different detection techniques, in the presence of Co-channel interference. AWGN limited environment. The IF filter is 8th order Butterworth with $BT=1$.

5.4.2 Performance over a flat Rayleigh fading channel

The performance of GMSK limiter-discriminator receiver was investigated over a flat Rayleigh fading channel. Two cases of flat fading were considered. In the first case we made the assumption that the fading rate is very slow compared to the data rate $1/T$, so that the random amplitude and the random phase of the carrier stay relatively constant over a long sequence of transmitted bits. In terms of Doppler spread, for this kind of channel the normalized doppler bandwidth $f_d T$ is approximately zero. The other case considered is the one when the channel fading rate is fast. In the two cases the channel is simulated using Jakes' sum of nine cosines technique [37].

The BER performance of a GMSK employing different detection techniques, over a flat Rayleigh fading channel, is plotted in Figure 5.28 and 5.29 for a slowly fading and fast fading channels, ($f_d T = 0.0005$ and 0.005 respectively). From the figures, one can see that the 2-bit I&D filter followed by 4-state Viterbi equalizer still maintains the superiority over the other detection techniques. It has an improvement of approximately 2 dB over the GMSK with 1-bit I&D filter followed by a 4-state Viterbi equalizer, at an average BER of 10^{-4} . It is also noticed that for the fast fading case, after certain value of SNR the error rate curve starts leveling off with increasing SNR. This irreducible error rate is due to the fast fading rate of the channel and the value of SNR after which this error rate floor is produced depends on the fading rate (or Doppler spread) of the channel. From the figures we can say that, for fast fading case, at $\text{SNR} < 30$ dB Gaussian noise dominates and the BER is a linear function of SNR. As the SNR increases, the BER performance is dominated by fading rate. As it is evident from Figure 5.27, the error floor has been reduced from 10^{-4} using 1-bit I&D filter followed by HD detection to 10^{-5} using 2-bit I&D filter followed by a 4-state Viterbi equalizer. For comparison, some results from published literatures are also shown in the figures.

Figure 5-24: Performance of GMSK ($h=0.5$, $B_rT=0.35$), using different detection techniques, over a very slowly fading flat Rayleigh channel ($f_dT=0$). The IF filter is 8th order Butterworth $BT=1$.

Figure 5-25: Performance of GMSK ($h=0.5$, $B_rT=0.35$), using different detection techniques, over a fast fading flat Rayleigh channel ($f_dT=0.005$). The IF filter is 8th order Butterworth $BT=1$.

5.5 Performance over a Two-Ray Frequency-Selective Channel Model

The performance of GMSK limiter-discriminator receiver over a frequency-selective Rayleigh fading channel was investigated. The frequency-selective fading was modeled by a Rayleigh faded delayed signal added to the Rayleigh faded main-path signal as described by Figure 4.9. This model is usually referred to as the two-ray model and it was adopted by the IS-54 standards to evaluate the tolerance of digital modems to delay spread in digital cellular systems. As a matter of fact, this simple model does not represent a realistic mobile fading channel since it has only a discrete two-ray impulse response which in turn produces a single frequency null in the frequency domain. Practically there could be more than two-ray and the frequency response may exhibit more than one frequency null. However, it provides a good approximation of the behavior of frequency-selective fading and also simplifies the analysis of such channels. The main parameters of this model are: (1) The average power ratio of the main ray to the delayed ray or C/D (As this ratio increases the channel becomes less selective). (2) The channel delay spread which in our case ranges from 0 to T sec and can be incrementa in steps of $0.0625T$. (3) the mobile speed (the Doppler spread of the channel). It has been shown in [39] that the Doppler spread of the channel has a little effect on the BER performance for the Doppler frequency in the range of interest for land mobile communications when the channel exhibits a frequency selective fading.

Quasi-static analysis will be used for evaluating the BER performance. The quasi-static analysis is carried out by making a series of channel snap shots and freezing that condition while obtaining the performance and then averaging over the possible snap shots. For simulation this approach is simple for implementation and it was carried out by

generating an array of 10000^1 sets of channel parameters drawn from channel parameter distributions and set up as an input file in the simulation program. The degree of severity of the channel time dispersion was first investigated with the aid of eye pattern which provides a qualitative performance assessment and later we investigated the BER performance under different channel conditions using the different detection techniques.

5.5.1 Eye Pattern Fluctuation characteristics due to delay spread using a static Two-Ray model

In frequency-selective fading channels, there arises an annoying problem of timing fluctuation of eye pattern caused by the multipath induced envelope delay distortion which has been found in most of the conventional demodulation techniques to be intolerable and becomes a source of abnormally burst errors. In this subsection we will take a closer look at this eye pattern fluctuations. In Figure 5.26 through 5.29, a number of simulated eye patterns are shown for various instantaneous conditions of a frequency-selective fading using the two-ray channel model. The signaling scheme used is a GMSK with $h=0.5$, a pre-modulation filter bandwidth $B_f T=0.35$ and a Gaussian pre-detection filter with bandwidth $BT=1$. We first investigated the effect of the carrier phase difference, ψ , between the main ray C and the delayed ray D at a fixed value of delay spread selectable from the range $T/4 \leq T_{ds} \leq T$ at an increment of $T/4$, and a two equal power rays i.e. $C/D=0$ dB. The increment of the phase difference ψ is $\pi/4$. As it can be seen from the figures, depending on the amount of delay spread, as the phase difference progresses towards π the deterioration of the eye pattern gets worse. For any value of delay spread T_{ds} the eye pattern is completely distorted at a phase difference of π which, in general, represents

1. The higher the number of channel realization, the more accurate the results and the longer the simulation time. For our case, we have found that this number provides a good compromise between simulation time and accuracy.

Figure 5-26: Eye pattern distortion in a static two-ray channel model with $T_d = T/4$, $C/D = 0$ dB and different values of phase difference $\Delta\Psi$ between the two rays. (a) $\Delta\Psi = \pi/4$ (b) $\Delta\Psi = \pi/2$ (c) $\Delta\Psi = 3\pi/4$ (d) $\Delta\Psi = \pi$.

Figure 5-27: Eye pattern distortion in a static two-ray channel model with $T_{\Delta} = T/2$, $C/D = 0$ dB and different values of phase difference $\Delta\Psi$ between the two rays. (a) $\Delta\Psi = \pi/4$ (b) $\Delta\Psi = \pi/2$ (c) $\Delta\Psi = 3\pi/4$ (d) $\Delta\Psi = \pi$.

(a)

(b)

(c)

(d)

Figure 5-28: Eye pattern distortion in a static two-ray channel model with $T_{d1}=3T/4$, $C/D=0$ dB and different values of phase difference $\Delta\Psi$ between the two rays. (a) $\Delta\Psi=\pi/4$ (b) $\Delta\Psi=\pi/2$ (c) $\Delta\Psi=3\pi/4$ (d) $\Delta\Psi=\pi$.

Figure 5-29: Eye pattern distortion in a static two-ray channel model with $T_{ds}=T$, $C/D=0$ dB and different values of phase difference $\Delta\Psi$ between the two rays. (a) $\Delta\Psi=\pi/4$ (b) $\Delta\Psi=\pi/2$ (c) $\Delta\Psi=3\pi/4$ (d) $\Delta\Psi=\pi$.

the worst case scenario for $C/D = 0$ dB at any value of T_{ds} . The reason for this severe distortion is that the two received signals are out of phase by π and therefore they add up in a very destructive manner, moreover because of the delay spread (time dispersion) the transmitted pulses will no longer have the same original shape.

Table 5-4: Summary of the Eye pattern behavior of a GMSK signal using Two-Ray channel model

Relative Phase Difference	Delay Spread	Observation	
		Eye Opening	Remarks
$\pi/4$ to π (in steps of $\pi/4$)	$T/4$	<ul style="list-style-type: none"> - Sampling Offset $+0.09 T$ - Max. eye opening remains the same for all cases except π - The eye is completely distorted at π 	<ul style="list-style-type: none"> For the case of π, At 10 dB power ratio between the two rays the eye is - 0.7 dB wider -The sampling offset is $-0.09 T$
	$T/2$	<ul style="list-style-type: none"> - Sampling Offset $+0.23 T$ - Smaller eye opening - The eye is completely distorted at π 	<ul style="list-style-type: none"> For the case of π, At 10 dB power ratio between the two rays the eye is - 1.67 dB wider -The sampling offset is $-0.13 T$
	$3T/4$	<ul style="list-style-type: none"> - Sampling Offset $+0.34 T$ - Small eye opening up to $3\pi/4$ - The eye is completely distorted from $3\pi/4$ and up 	<ul style="list-style-type: none"> For the case of π, At 10 dB power ratio between the two rays the eye is -2.5 dB wider -The sampling offset is $-0.09 T$
	T	<ul style="list-style-type: none"> - The eye has a 3-level opening at the sampling instant up to $3\pi/4$ - The eye is completely distorted from $3\pi/4$ and up 	<ul style="list-style-type: none"> - same as the above case except that the sampling offset is $0.13 T$

In Figure 5.30 the ratio C/D is made equal to 10 dB and the eye pattern is shown for different delay spreads T_{ds} with the phase difference ψ set to it's worst case value, i.e. $\psi = \pi$. These eye pattern plots are intended to show the effect of the relative power ratio between the two rays. The improvement in the eye opening is evident and that is because the destructive add up is not as severe due to the high relative power difference between the two rays and in such scenario the BER performance of the system is expected to be improved compared to the equal power case. In other words as the ratio C/D increases the channel becomes less selective. Table 5-4 provides a summary of the eye pattern behavior

Figure 5-30: Eye pattern distortion in a static two-ray channel model with $\Delta\Psi=\pi$, $C/D=10$ dB and different values of time delay T_{d3} between the two rays. (a) $T_{d3}=T/4$ (b) $T_{d3}=T/2$ (c) $T_{d3}=3T/4$ (d) $T_{d3}=T$.

of the GMSK signal at different instantaneous conditions of a static dispersive channel. Note that the measurements are all relative to a dispersion-free eye pattern.

5.5.2 Error rate performance

The Limiter-discriminator receiver of a GMSK signal was tested over a frequency-selective fading channel. The BER curves using the different detection techniques are shown in Figures 5.31 through 5.33 for different channel conditions. From the figures, the BER performance of each receiver arrangement increases as the channel delay spread increases. In any of the cases investigated, there is an error rate floor due to saturation of the bit errors caused by severe channel dispersion. The influence of the ratio C/D on the different receiver arrangement was also investigated. By comparing Figure 5.31 and 5.33 we can see that the BER performance has been improved as the ratio C/D is increased from 0 to 10 dB. From the BER curves it can be seen that both HD and SD detection techniques exhibit almost the same BER performance. The justification of the above situation is that due to the time-varying nature of the channel, the 4-state Viterbi equalizer requires knowledge of the channel coefficients in order to update the equalizer's reference points. Therefore, the equalizer is not expected to work properly with fixed reference points under these time-varying conditions.

The updating of the reference points can be achieved by including a channel estimator in parallel with the detection algorithm. Due to the nonlinear nature of the FM discriminator receiver, channel estimation is very difficult to be implemented. In addition, the estimator might not converge due to the occurrence of the clicks or if the channel is rapidly changing. For the purpose of examining how the receiver behaves with an adaptive Viterbi algorithm, we assumed that we have an ideal channel estimator. This was done by using the channel coefficients, which are generated in the simulation program, for updating the algorithm. The updating of the reference points of the Viterbi equalizer is carried out by

first running the simulation without the noise, the reference points of the algorithm are then calculated. The resulting reference points are then stored in an input file and the simulation is run with the noise over the same channel for estimating the BER performance. Figure 5.34 shows the BER performance of GMSK signalling under the same conditions as in Figure 5.31. As it is evident, the performance with the adaptive algorithm has been improved compared to figure 5.35. The error floor has been reduced an order of magnitude and the GMSK with the 2-bit I&D filter followed by the 4-state equalizer provides the best performance compared to the other receivers arrangements. The performance under slowly varying flat fading channel is also shown in the figure for comparison, where the performance of GMSK under frequency-selective fading has been brought to 5 dB from the flat fading case at an average BER performance of 10^{-3} . In Figure 5.34 it is also noticed that the error rate curves of the GMSK with the adaptive algorithms start leveling off at high SNR which is probably due to a residual ISI which is not being taken care of by the equalizer due to the shorter ISI memory used.

Figure 5-31: BER performance of GMSK signalling ($h=0.5$, $B_1T=0.35$), using different detection techniques, over a frequency-selective fading channel. The normalized delay spread, $T_{ds}=0.25$, $C/D=0$ dB and $SIR=\infty$.

Figure 5-32: BER performance of GMSK signalling ($h=0.5$, $B_fT=0.35$), using different detection techniques, over a frequency-selective fading channel. The normalized delay spread, $T_{ds}=0.5$, $C/D=0$ dB and $SIR=\infty$.

Figure 5-33: BER performance of GMSK signalling ($h=0.5$, $B_1T=0.35$), using different detection techniques, over a frequency-selective fading channel. The normalized delay spread, $T_{ds}=0.25$, $C/D=10$ dB and $SIR=\infty$.

Figure 5-34: BER performance of GMSK signalling ($h=0.5$, $B_f T=0.35$), using different detection techniques, over a frequency-selective fading channel. The normalized delay spread, $T_{ds}=0.25$, $C/D=0$ dB and $SIR=\infty$. The Viterbi equalizers in this case are adaptive.

6

Conclusions and Recommendations

6.1 Thesis Summary

The intent of this thesis was to investigate post-detection processing techniques for improving the performance of the non-coherent limiter-discriminator receiver of narrow-band digital FM signal. Most of the results presented in this thesis were obtained using computer simulation. A computer program was developed to investigate the BER performance improvement of different receiver arrangements under different channel conditions. Among the broad class of digital FM signals, three were dealt with the most; these are: MSK, Duobinary coded MSK and GMSK.

Based on the findings of this thesis, the following conclusions can be made:

- (1) The IF filter plays an important role in determining the BER performance of the limiter-discriminator receiver. In the design of such filter, two conflicting requirements have to be taken into consideration; the filter bandwidth should be narrow for noise reduction, but it should be wide enough for minimizing the signal distortion. An optimum design is usually possible. The filter approximation used is also important. For the cases considered (i.e. $h < 1$), a bandwidth equals to the data rate is about optimum from a BER performance point of view. For the two filter types examined, the 8th order Butterworth filter provides a superior performance

over the Gaussian filter. That is because the Butterworth filter exhibits a flat pass-band response characteristics and produces less ISI compared to the Gaussian filter.

- (2) The post-detection I&D filter is more effective for the 2-bit integration period, since this provides more smoothing of the noise at the output of the discriminator. By examining the eye pattern at the output of the 2-bit I&D filter, the eye opening at the sampling instant is wider than that of 1-bit I&D filter. For the case of MSK the improvement in noise margin when using the 2-bit I&D filter is approximately 1.25 dB and for GMSK it is approximately 1.5 dB. This suggests that the 2-bit I&D filter provides some equalization for in-band ISI distortion suffered in the transmission.
- (3) 2-state and 4-state Viterbi equalizers, using the two and three-level eye opening, have been successfully applied to the output of the I&D post-detection filter. The performance of this equalizer has been simulated for various schemes. In the simulation the equalizer has been implemented based on the exact noise distribution at the output of the receiver. The implementation based on the approximation that the noise is Gaussian has also been simulated for comparison. In the low SNR region, the performance of the equalizer based on the Gaussian assumption provides a poorer performance (approximately 0.8 dB inferior to the non-Gaussian implementation at an error rate of 10^{-2}). At the high SNR region the difference in performance between the two implementations is very small with the non-Gaussian implementation being slightly better (approximately 0.1 dB at 10^{-5}). The reason behind that is the fact that the noise at low SNR has an impulsive component or clicks, therefore the equalizer based on the Gaussian noise approximation is not expected to perform well in this region. At high SNR the

clicks vanish and the exact noise distribution becomes a Gaussian like function, and therefore both implementation provide similar performance. This result shows that the implementation of the Viterbi equalizer based on the Gaussian assumption is very robust in improving the performance of the discriminator receiver, at the same time it is simple for hardware implementation.

- (4) For digital FM schemes with little ISI memory (e.g. MSK), the BER improvement when using the sequence estimation is not high when compared to hard decision detection performance. For schemes with more ISI memory (e.g. GMSK) the BER improvement is high compared to hard decision detection. In other words, the Viterbi equalizer seems to be superior when the response of the channel filters is longer than the symbol period. For example, for MSK with Butterworth IF filter and 4-state equalizer following a 1-bit I&D filter the difference is approximately 1.2 dB at an error rate of 10^{-5} , whereas for GMSK with the same receiver arrangements the improvement is approximately 3.7 dB. It is expected that the equalizer will provide more improvements if it utilizes the full ISI memory.
- (5) It has been shown that the use of Viterbi equalizer following a 2-bit I&D filter, has brought the BER performance of MSK and GMSK signals using the limiter-discriminator detector to approximately 1.5 dB and 2.5 dB loss in SNR, respectively, compared to coherent MSK reception in the case of no fading.
- (6) It has been demonstrated that the use of soft decision detection following the post-detection I&D filter has improved the BER performance of GMSK signal, with B_1T product of 0.35 and $h=0.5$, in the presence of cochannel interference. A minimum improvements of 6 dB has been achieved, which shows that with the soft decision detection the discriminator receiver is very tolerable to cochannel interference. The BER performance in a channel that suffers from flat Rayleigh fading has also been

improved. For the slowly varying channel the improvement, compared to hard decision detection, is approximately 5 dB with the 1-bit I&D filter and 6 dB with the 2-bit I&D filter, at an average error rate of 10^{-3} . For a channel with rapidly changing rate the irreducible error floor has been reduced an order of magnitude compared to hard decision detection (for the cases considered it has been reduced from 10^{-3} to 10^{-4}).

- (7) The eye pattern fluctuation due to channel dispersion has been closely examined using a nonfading two-ray channel model. The combined effect of fading and dispersion on the BER performance of the limiter-discriminator receiver has been studied for different receiver arrangements. The resulting BER curves show that the use of a fixed Viterbi equalizer provides the same BER performance as the simple bit-by-bit detection. An adaptive Viterbi equalizer with perfect adaptation can greatly improve the receiver performance and reduces the error floor an order of magnitude.

6.2 Recommendations for future work

Much more work remains to be done in the area of non-coherent detection of digital FM signalling. It would appear that the following areas would be worth investigation

- (1) When we combine the effects of channel and the band-limiting filters the ISI memory is expected to span several symbols, especially if the channel has a longer delay spread. Since the number of states of Viterbi algorithm depends on the amount of ISI memory to be taken care of, it would be possible to improve the performance of the discriminator receiver over a dispersive channel with Viterbi equalizer that utilizes the full memory of the channel (i.e, by using a larger number of trellis states in the Viterbi algorithm).

- 2) For the Viterbi equalizer to work properly over a multipath dispersive channel, knowledge of the channel characteristics are required in order to adapt the reference points of the equalizer. In this study we have used an ideal channel estimator for adapting the equalizer. It would be worth looking at a more realistic channel estimation techniques for the nonlinear discriminator receiver.
- 3) Due to multipath delay spread, the timing characteristics of the eye pattern are severely deteriorated. This problem has a serious consequences on the clock recovery circuits and it may result in a long burst of errors. It would also be worth looking at the possibility of estimating the amount of delay spread and to use that as a feedback for an adaptive clock recovery circuit, which keeps track of the optimum sampling instant at the detector output.
- 4) The use of other noncoherent detection techniques such as differential detection combined with soft decision detection has to be studied for various digital FM signalling schemes.

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