İSTANBUL TECHNICAL UNIVERSITY ★ INSTITUTE OF SCIENCE AND TECHNOLOGY

DESIGN AND REALIZATION OF GMSK MODEM

M.Sc. Thesis by
Arif Kürşad KAVAS, B.Sc.
(504031202)

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Supervisor (Chairman): **Prof. Dr. Osman PALAMUTÇUOĞULLARI**
Members of the Examining Committee **Prof.Dr. Mehmet Sait TÜRKÖZ**
Prof.Dr. Bülent ÖRENCİK

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GMSK MODEM TASARIMI VE GERÇEKLENMESİ

YÜKSEK LİSANS TEZİ
Müh. Arif Kürşad KAVAS
(504031202)

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Tez Danışmanı : **Prof.Dr. Osman PALAMUTÇUOĞULLARI**
Diğer Jüri Üyeleri **Prof.Dr. Mehmet Sait TÜRKÖZ**
Prof.Dr. Bülent ÖRENCİK

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FOREWORD

In todays' communication systems, the channel separations and the channel bandwidths are restricting tasks. In order to use the frequency spectrum efficiently, standards have been developed for different communication systems. To use the communication channel efficiently, without drawback from the information capacity, new modulation schemes have been developed and used. GMSK modulation is one of these and used commonly on cellular land mobile channels as well as in satellite communication systems.

This thesis consists of design and realization of a GMSK MODEM on a DSP chip. First of all, I explained what the bandwidth means for a wireless radio channel and importance of constant envelope scheme modulation. Then I defined the GMSK modulation with respect to its counterparts. I have stated the reasons why GMSK is popularly used by taking into interest the bandwidth considerations. I gave the modulator and the demodulator structures in the GMSK and I modelled and realized the GMSK modulator and the demodulator on MATLAB. Using one bit differential detection in the demodulator, I made symbol synchronizations in order to compensate for timing errors, for the extraction of the clock. I simulated the modulator and the demodulator for noisy channel by using an appropriate model for AWGN. I made simulations to observe the multipath fading effects and the co-channel interference effects on the performance of the MODEM.

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May, 2006 **Arif Kürşad KAVAS**

CONTENTS

FOREWORD **iii**

CONTENTS **iv**

LIST OF ABBREVIATIONS **vi**

LIST OF TABLES **vii**

LIST OF FIGURES **viii**

ÖZET **ix**

SUMMARY **xi**

1.INTRODUCTION **1**

2.DIGITAL CARRIER SYSTEMS **4**

2.1. Modulation Schemes **5**

2.1.1. Amplitude Shift Keying (ASK) 5
2.1.2. Frequency Shift Keying (FSK) 7
2.1.3. Phase Shift Keying 8

2.2. Spectral Density **10**

2.2.1. Energy Spectral Density 10
2.2.2. Power Spectral Density 11

2.3. Eye Diagram **13**

2.4. Performance Degrading Factors in Digital Communications **15**

2.4.1. Noise 15
2.4.1.1. Additive White Gaussian Noise (AWGN) 16
2.4.1.2. AWGN Channel & Channel Capacity 17
2.4.2. Multipath Fading 19
2.4.3. Co-Channel Interference 21
2.4.4. Intersymbol Interference 22

3.GMSK MODULATION **25**

3.1. Continuous Phase Modulation **25**

3.2. Continuous-Phase Frequency-Shift Keying **27**

3.3. Minimum-Shift-Keying **27**

3.4. Gaussian Minimum Shift Keying 29
3.4.1. Modulation 31
3.4.1.1. FM Modulator 31
3.4.1.2. I/Q Modulator 32
3.4.2. Demodulation 36
3.4.2.1. Detection Using Composite Signal 37
3.4.2.2. Detection Using Decomposed I/Q Signals 39
4.SYNCHRONIZATION 42
4.1. Squaring Algorithm 42
4.2. Joint Symbol Timing Error and Frequency Offset Estimations 46
5.SIMULATION RESULTS AND PERFORMANCE EVALUATION 50
5.1. Modulator 50
5.2. Demodulator 54
5.3. AWGN Channel Performance 55
5.4. Multipath Fading Performance 56
5.5. Co-Channel Interference Performance 57
6. CONCLUSION 58
REFERENCES 59
APPENDIX A: MATLAB CODE 60
APPENDIX B: MODEM SCHEMATIC 62
APPENDIX C: PCB LAYOUT TOP VIEW 63
APPENDIX D: PCB LAYOUT BOTTOM VIEW 64
AUTOBIOGRAPHY 65

LIST OF ABBREVIATIONS

MODEM	: Modulator/Demodulator
GMSK	: Gaussian Minimum Shift Keying
ISI	: Intersymbol Interference
GSM	: Global System for Mobile - Communications
MSK	: Minimum Shift Keying
DSP	: Digital Signal Processing/Processor
MIPS	: Million Instructions Per Second
PCM	: Pulse Code Modulation
RF	: Radio Frequency
ASK	: Amplitude Shift Keying
OOK	: On-Off Keying
FSK	: Frequency Shift Keying
PSK	: Phase Shift Keying
BPSK	: Binary Phase Shift Keying
QPSK	: Quadrature Phase Shift Keying
ESD	: Energy Spectral Density
PSD	: Power Spectral Density
RZ	: Return-to-Zero
NRZ	: Non Return-to-Zero
AWGN	: Additive White Gaussian Noise
SNR	: Signal to Noise Ratio
BER	: Bit Error Rate
CPM	: Continuous Phase Modulation
LRC	: Raised Cosine with Pulse Length L
TFM	: Tamed Frequency Modulation
LSRC	: Spectral Raised Cosine with Pulse Length L
LREC	: Rectangular Frequency Pulse with Length L
CPFSK	: Continuous Phase Frequency Shift Keying
OQPSK	: Offset Quadrature Phase Shift Keying
BT	: Bandwidth-Time Product
FM	: Frequency Modulation
FFT	: Fast Fourier Transform
VCO	: Voltage Controlled Oscillator

LIST OF TABLES

	Page No
Table 3.1 The Frequency Pulse Defined for Some CPM Schemes	26
Table 3.2 Occupied Bandwidth for Specified Percentage Power	30

LIST OF FIGURES

		Page No
Figure 2.1	: Basic Transmitter	4
Figure 2.2	: Generation of the ASK Signal	7
Figure 2.3	: Generation of the FSK Signal	8
Figure 2.4	: BPSK Signal	9
Figure 2.5	: QPSK Waveform	10
Figure 2.6	: Eye Diagram	14
Figure 2.7	: Eye Pattern After Filtering and at the Receiver Output	15
Figure 2.8	: Noise Addition	17
Figure 2.9	: Original and AWGN Channel Output Signals at SNR=10dB	19
Figure 2.10	: Multipath Fading Channel Model	21
Figure 2.11	: Co-Channel Interference Model	22
Figure 2.12	: Filtering Effect of the Channel	22
Figure 2.13	: Filtered Pulses Interfering Neighbouring Symbols	23
Figure 2.14	: Filtered NRZ Data	23
Figure 3.1	: MSK Signal Formation	28
Figure 3.2	: Comparison of Power Spectral Densities	30
Figure 3.3	: GMSK FM Modulator	31
Figure 3.4	: I/Q Representation of a Signal	32
Figure 3.5	: I and Q Forming Composite Output	33
Figure 3.6	: GMSK I/Q Modulator	33
Figure 3.7	: NRZ Mapping	34
Figure 3.8	: Gaussian Impulse Response	35
Figure 3.9	: One Bit Differential Detector	37
Figure 3.10	: I/Q Decomposition	39
Figure 3.11	: I/Q Detector	41
Figure 4.1	: Block Diagram of the Synchronization Method	46
Figure 4.2	: Discriminator Output Periodic Signal	48
Figure 5.1	: Oversampling Binary Data	50
Figure 5.2	: Unipolar to Bipolar Conversion	51
Figure 5.3	: Impulse Response of the Gaussian Low-Pass Filter	51
Figure 5.4	: Smoothed Data	52
Figure 5.5	: Accumulated Continuous Phase	52
Figure 5.6	: I/Q Signals	53
Figure 5.7	: GMSK Signal	53
Figure 5.8	: Demodulator Output	54
Figure 5.9	: Hard Decision Output Received Data	55
Figure 5.10	: Performance in the Presence of Noise	55
Figure 5.11	: Performance at BT = 0.3 for Multipath Fading Channel	56
Figure 5.12	: Performance at BT = 0.5 for Multipath Fading Channel	56
Figure 5.13	: Performance at BT = 0.3 for Co-Channel Interference	57
Figure 5.14	: Performance at BT = 0.5 for Co-Channel Interference	57

GMSK MODEM TASARIMI VE GERÇEKLENMESİ

ÖZET

Geride bıraktığımız yüzyıl, telsiz haberleşme açısından birçok yeniliklere sahne olmuştur. Yaşandığı dönemde devrim niteliğinde olan telsiz olarak basit bir verinin havadan iletiminden günümüze çok fazla yol kat edilmiş ve bugün karmaşık modülasyon sistemleri kullanılarak, çok miktarda bilginin verimli bir şekilde iletimi olanaklı hale gelmiştir.Sesin ve daha sonra da görüntünün analog modülasyon türleri kullanılarak iletimi yerini daha kompleks fakat daha verimli olan sayısal modülasyon çeşitlerine bırakmıştır.

Sayısal modülasyon, birçok yönden analog modülasyon çeşitlerine üstünlük sağlamıştır. Bu üstünlükteki en önemli kıyaslama kriterleri veri kapasitesi ve bant genişliği kavramlarıdır.Aslında veri kapasitesi ve bant genişliği kavramları birbiri ile yakından alakalıdır. Bir iletişim kanalında, iletilen verinin miktarının artması, birim zamanda iletilen verinin miktarının da artmasına ve dolayısıyla iletimin daha geniş bantta yapılmasına sebep olur. Malesef sonsuz bir banda sahip değiliz. İletmek istediğimiz bilgi ise sürekli olarak artmaktadır.

Bugün çok yüksek hızda gerçek zamanlı çalışan, sayısal işaret işleyen işlemciler sayesinde işaret işleme tekniklerinin uygulamaları artmıştır. Bu işlemcilerin sayısal haberleşme sistemlerinde de kullanımı yaygınlaşmıştır. Birçok sayısal modulasyon türleri bu işlemcilerle gerçekleştirilmektedir.Cep telefonlarının da dahil olduğu hücresel mobil kanallardan, uydu haberleşmesine birçok sistem yüksek performanslı işaret işleyici işlemciler kullanarak, haberleşme sistemindeki analog kısımların yükünü hafifletip, sayısal olarak iletişime olanak sağlamaktadrılar.

Bu çalışmada, son zamanlarda popüler olarak kullanılan GMSK modulasyonu, modülator ve demodülatörünün tasarımı ile anlatılmıştır. İlk olarak günümüzde kullanılan bazı sayısal modulasyon türleri tanıtılmıştır. Ardından GMSK modulasyonu detaylı bir şekilde ele alınmıştır. Modulator ve demodulator tasarlanmasının ardından, gürültü, kanallara arası girişim ve çok yollu sönme için performans simulasyonları yapılmıştır.Sonuç olarak bit-hata oranları çıkartılmıştır. Sembol senkronizasyonu sağlanarak, demodule edilen işaretin zamanlama hataları düzeltilmiştir. Son olarak, tasarlanan MODEM, TMS320C5509A sayısal işaret işlemcisinde gerçeklenmiştir.

DESIGN AND REALIZATION OF GMSK MODEM

SUMMARY

The past century has witnessed lots of innovations in the field of wireless communication. From the times, where it was a revolution to transmit simple information by means of wireless techniques, till today, many steps have been taken, and today by the use of complex modulation techniques, it has become possible to transmit large amounts of information, in an efficient way. Transmission of audio and then the video by using analog modulation techniques are replaced with digital modulation techniques which are much more complex but more efficient.

Digital modulation gained superiority with respect to analog modulation in a number of ways. The most important comparison criterions in this superiority are the band efficiency and transmitted information capacity. Actually the band efficiency and the transmitted information capacity concepts are closely related. In a communication channel, increase in the amount of information transmitted, increases the amount of information transmitted in unit time and therefore causes the communication to be made in a wider band. Unfortunately we do not have infinite bandwidths. Though, the information that we want to transmit goes on increasing.

Today by the use of digital signal processors, which work at high speeds and in real time, applications of digital signal processing techniques have been increased. Usage of these processors in digital communication systems has spread as well. Most of the digital modulation schemes are realized using these processors. Most of the systems varying from cellular mobile channels, including mobile phones, to satellite communication systems use high performance digital signal processors and decrease the burden on analog side, allowing digital communication.

In this work, GMSK modulation, which is popularly used lately is described, by design of the modulator and the demodulator. First of all, some digital modulation methods, which are used today are described. Then the GMSK modulation is described in detail. After design of the modulator and the demodulator simulations are made to test the performance of the overall system, with respect to noise, co-channel interference and multipath fading. As a result, bit-error rates are obtained.
By symbol synchronization, timing errors of the demodulated signal are corrected. Finally, designed MODEM is realized on the TMS320C5509A digital signal processor.

1. INTRODUCTION

The increasing demand in the amounts of data to be transferred through the wireless media arise the need for different modulation methods, capable of handling high data rates using lower bandwidths, in the digital communication systems. On the other hand standards have been developed in order to partition the available spectrum to variety of communication systems. The necessity for higher data rates in a wireless system with a predetermined bandwidth is a challenging task. The modulation schemes utilized in today's wireless communication systems focus on the problem of transmission of data rates with taking into consideration the usage of the spectrum and the problem of ISI.

The major purpose of the wireless digital communication system is the transmission of the information in digital form, from one location to another. Evolution of the digital media devices increased the importance of wireless transmission of data. For example, through the development of the cellular systems, the first generation cellular devices were limited to transmission of speech, the second generation allowed the speech and the limited data and the third generation came up with higher supported data rates. The satellite communication systems where massive amount of data is transferred use more complex methods in order to compress and transfer the data in a limited bandwidth.

There are a number of factors that determine the choice of the modulation scheme. Gaussian Minimum Shift Keying (GMSK) is a digital modulation scheme, which is a linear and constant envelope scheme, crucial in the second and the third generation cellular systems achieving high capacity. GMSK is vastly popular in Europe's GSM cellular standard. Its' narrow bandwidth and ability to use coherent detection, characterizes the GMSK as a constant envelope modulation technique.

GMSK is a member of the Minimum Shift Keying (MSK) modulation family. It differs from the MSK in the aspect of the usage of the filter. GMSK is a result of the attempts to improve the MSK power spectrum. However, the advantage of MSK which is that it does not produce ISI (Intersymbol Interference) is not seen in GMSK. The transmitted pulse in the MSK is confined within its bit duration resulting in no adjacent channel interference. But the GMSK possesses a more compact spectrum, with the application of the low pass filter, helping to reduce its spectral side lobes.

In the GMSK, the phase of the carrier signal is continuously varied by the antipodal signal, which has been shaped by a Gaussian filter. Since it is a type of MSK, it has a modulation index of 0.5 and may be demodulated using differential detection. The Gaussian filter concentrates the energy, allowing for the lower out of band power. The constant envelope allows GMSK to be less susceptible to a fading environment.

The development of real time digital signal processor chips made the realization of complex modulation schemes possible on a single chip rather than employing discrete realizations, where sensitivity of the system to the environmental effects is a stringent task. Today the DSP chips reach the clock speeds of 1 GHz and the processing speeds up to 8000 MIPS. This is important because in order for a system to work in real time, meaning that giving fast responses to incoming inputs, it has to implement the necessary tasks in a fast manner as the delays are critical and not accepted. The DSP chips fulfill complex digital processes in real time, which makes them a candidate for realization of complex modulation methods.

GMSK modulation and demodulation consists of a number of both simple and complex mathematical operations, such as integration and convolution. In the digital domain, these operations have to be carried out using numerical methods. These methods consist of lots of calculation steps to obtain the solution. Due to the reasonable number of steps for realization of the mathematical procedures, the modulation and demodulation processes take time. But the system must respond to the incoming signal in no time. This is the point where it is get to known why the

DSP is widely used in the realization of digital communication schemes. Its superior performance allows the designers to realize complex communication algorithms as well as simple communication functions. In addition to the computing power, the designer eliminates the tolerances and sensitivities that come with discrete realization. The DSP also offers built in functions, which fasten the coding procedure. The reasons stated above, make the DSP chipsets of considerable interest in the realization of variety of wireless infrastructure.

2. DIGITAL CARRIER SYSTEMS

Modulation means to vary or change. The information bearing signals modulate a signal called carrier in a manner that we call modulation, in order to transmit the information through a media either wireless or wired. The modulation process can be done according to any reliable detectable change in signal characteristics. It can be either amplitude or frequency or the phase of the carrier signal that changes according to the information signal. At the receiver, demodulation occurs by detection of changes in the carrier signal.

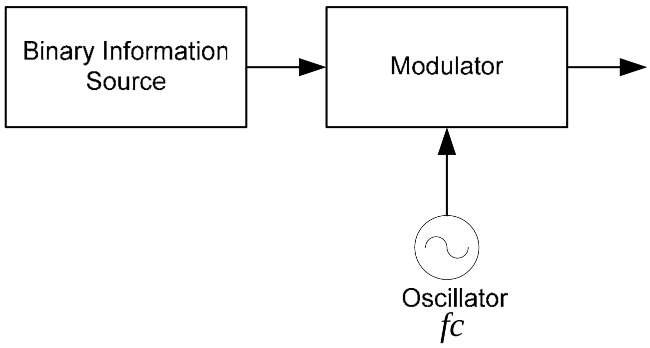

Figure 2.1: Basic Transmitter

In digital communication systems, the baseband signals have sizable power at low frequencies and they are suitable for transmission over a pair of wires or coaxial cables. Local telephone communications as well as short-haul PCM are examples of this. Baseband signals cannot be transmitted over a radio link because this would necessitate impracticably large antennas to efficiently radiate the low-frequency spectrum of the signal. In order to reach this aim, the signal spectrum must be shifted to a high-frequency range. A spectrum shift is also required to transmit several messages simultaneously by sharing the large bandwidth of the transmission medium. The spectrum of a signal can be shifted to a higher frequency by modulating a high-frequency sinusoid by the baseband signal.

Throughout years analog modulation techniques were used and are still being used to transmit voice, image and other kinds of information signals. The development of digital techniques, the need for efficiency and demand for higher data rates, made the digital modulation much more popular.

One of the main advantages in the digital communications when compared to analog is the ease of regeneration of the digital signals. The digital circuitry is less susceptible to distortion and interference. The digital circuits operate on two distinct states and therefore there must be a large amount of disturbance to change the operation state from one to another.

The move to digital modulation provides more information capacity, compatibility with digital data services, higher data security, and better quality in communication. The main constraints in the communication systems are the available bandwidth, permissible power and the inherent noise level of the system. The RF spectrum must be shared, yet every day there are more users for that spectrum, as demand for communication systems increases. Digital modulation schemes have greater capacity to convey large amounts of information than analog modulation schemes.

2.1 Modulation Schemes

2.1.1 Amplitude Shift Keying (ASK)

An ASK signal can be defined by;

$$s(t) = A \cdot m(t) \cdot \cos 2\pi f_c t \qquad 0 \le t \le T \qquad \textbf{(2.1)}$$

where A is a constant, m(t) = 1 or 0 , f_c is the carrier frequency, and T is the bit duration. Taking $V_c(t)$ as the carrier signal and $V_d(t)$ as the data signal yields the mathematical notation for the ASK signal as below.

$$V_c(t) = \cos\omega_c t \tag{2.2}$$

$$V_d(t) = \frac{1}{2} + \frac{2}{\pi}\left(\cos\omega_0 t - \frac{1}{3}\cos\omega_0 t + \frac{1}{5}\cos\omega_0 t - ...\right) \tag{2.3}$$

$$V_{ASK}(t) = V_c(t) \cdot V_d(t) \tag{2.4}$$

$$V_{ASK}(t) = \frac{1}{2}\cos\omega_c t + \frac{2}{\pi}\left\{\cos\omega_c t \cdot \cos\omega_0 t - \frac{1}{3}\cos\omega_c t \cdot \cos\omega_0 t + ...\right\} \tag{2.5}$$

$$\begin{aligned} V_{ASK}(t) = {} & \frac{1}{2}\cos\omega_c t + \frac{1}{\pi}\cos(\omega_c - \omega_0)t + \cos(\omega_c + \omega_0)t \\ & - \frac{1}{3\pi}\left[\cos(\omega_c - 3\omega_0)t + \cos(\omega_c + 3\omega_0)t\right] \\ & + \frac{1}{5\pi}\left[\cos(\omega_c - 5\omega_0)t + \cos(\omega_c + 5\omega_0)t\right] - .. \end{aligned} \tag{2.6}$$

In the ASK modulation, according to the incident bit on a serial signal path, the carrier is either transmitted, that is the case when the bit is 1, or no transmission occurs, the case when the bit is 0. It is also referred as On-Off Keying(OOK).

Figure 2.2 shows the ASK signal according to the occurring bits. The bit sequence is shown in the first signal. The second signal is the modulating carrier signal. According to the incident ones and zeros, the output OOK signal is formed as in the third signal.

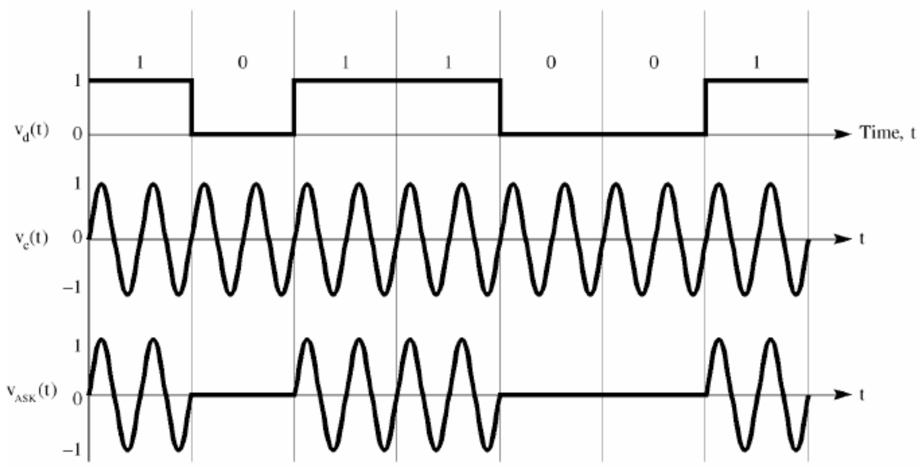

Figure 2.2: Generation of the ASK Signal

2.1.2 Frequency Shift Keying(FSK)

In the Frequency Shift Keying, frequency of the carrier is switched between two distinct frequencies such as,

$$\phi_1(t) = A\sin\omega_1 t \quad \textbf{(2.7)}$$

$$\phi_2(t) = A\sin\omega_2 t \quad \textbf{(2.8)}$$

As we see that the magnitude of the transmitted signal remains unchanged while the frequency of the signal changes. Each carrier can be assigned to an occurring symbol, as, if a 1 occurs first signal is transmitted, and if a zero occurs the second signal is transmitted. The vice-versa is valid too. The FSK signal can be written as below, taking $V_d(t)$ as the data signal.

$$V_{FSK}(t) = \cos\omega_1 t \cdot V_d(t) + \cos\omega_2 t \cdot V_{dc}(t) \quad \textbf{(2.9)}$$

The two carriers are ω_1 and ω_2 ,

$$V_{dc}(\mathrm{t}) = 1 - V_d(\mathrm{t}) \quad \textbf{(2.10)}$$

$$\begin{aligned} V_{FSK}(t) = {} & \cos\omega_1 t\left\{\frac{1}{2}+\frac{2}{\pi}\left(\cos\omega_0 t-\frac{1}{3}\cos\omega_0 t+\ldots\right)\right\} \\ & +\cos\omega_2 t\left\{\frac{1}{2}-\frac{2}{\pi}\left(\cos\omega_0 t-\frac{1}{3}\cos\omega_0 t+\ldots\right)\right\} \end{aligned} \quad \textbf{(2.11)}$$

$$
\begin{aligned}
V_{FSK} &= \frac{1}{2}\cos\omega_1 t + \frac{1}{\pi}\left\{\cos\left(\omega_1-\omega_0\right)t+\cos\left(\omega_1+\omega_0\right)t\right\} \\
&-\frac{1}{3\pi}\left\{\cos\left(\omega_1-3\omega_0\right)t+\cos\left(\omega_1+3\omega_0\right)t+\ldots\right\} \\
&+\frac{1}{2}\cos\omega_2 t + \frac{1}{\pi}\left\{\cos\left(\omega_2-\omega_0\right)t+\cos\left(\omega_2+\omega_0\right)t\right\} \\
&-\frac{1}{3\pi}\left\{\cos\left(\omega_2-3\omega_0\right)t+\cos\left(\omega_2+3\omega_0\right)t+\ldots\right\}
\end{aligned} \qquad \textbf{(2.12)}
$$

In Figure 2.3, generation of the FSK signal is shown. The first signal is the binary information, the second and the third are the two distinct carrier frequencies. The fourth signal is the FSK signal obtained by switching between the two carrier signals according to the bit sequence, in a predetermined carrier allocation for the symbols.

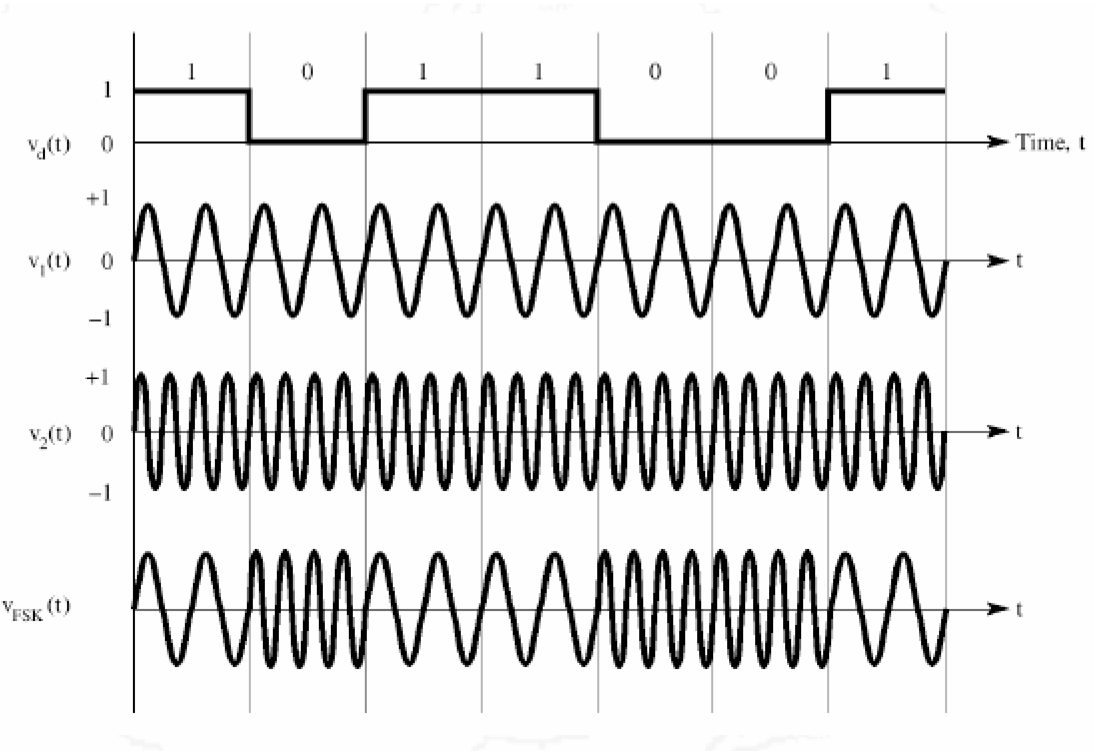

Figure 2.3: Generation of the FSK Signal

2.1.3 Phase Shift Keying (PSK)

Phase Shift Keying technique buries the information in the phase of the carrier signal. In PSK, the phase of a single carrier signal is varied between two different phases, for binary signals. For the binary values 0 and 1, the following carrier with two different phase angles can be chosen.

$$\phi_1(t) = A\sin\left(\omega t + \varphi_1\right) \quad \textbf{(2.13)}$$

$$\phi_2(t) = A\sin\left(\omega t + \varphi_2\right) \quad \textbf{(2.14)}$$

The carrier and the bipolar data signal are given to generate PSK signal.

$$V_c(t) = \cos\omega_c t \quad \textbf{(2.15)}$$

$$V_d = \frac{4}{\pi}\left\{\cos\omega_0 t - \frac{1}{3}\cos 3\omega_0 t + \frac{1}{5}\cos 5\omega_0 t - ...\right\} \quad \textbf{(2.16)}$$

$$V_{PSK} = V_c(t)\cdot V_d(t) \quad \textbf{(2.17)}$$

$$= \frac{4}{\pi}\left\{\cos\omega_c t\cdot\cos\omega_0 t - \frac{1}{3}\cos\omega_c t\cdot\cos 3\omega_0 t + ...\right\} \quad \textbf{(2.18)}$$

$$V_{PSK} = \frac{1}{\pi}\{\cos(\omega_c - \omega_0)t + \cos(\omega_c + \omega_0)t\} - \frac{1}{3\pi}\{\cos(\omega_c - 3\omega_0)t + \cos(\omega_c + 3\omega_0)t\} + .. \quad \textbf{(2.19)}$$

The binary signal, alternate the phase of the carrier signal between two different phases. This is called binary PSK or BPSK. The symbols and the BPSK signal are shown in Figure 2.4.

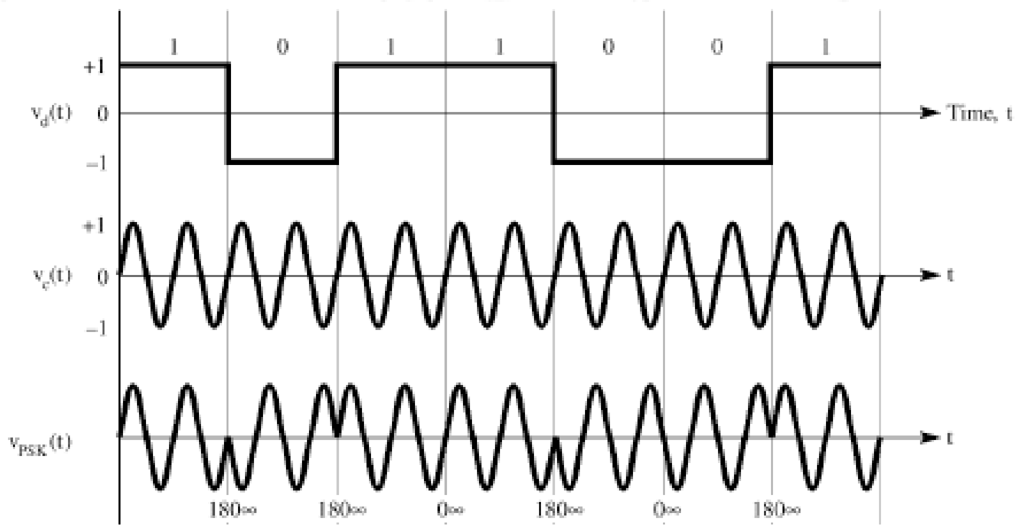

Figure 2.4: BPSK Signal

In the BPSK, bandwidth efficiency is 1bps/Hz. This means that the amount of bits sent per second per Hertz is one. The Quadrature Phase Shift Keying QPSK, is a type of Phase Shift Keying, with an improved bandwidth efficiency of 2bps/Hz. In QPSK the symbols consist of two bits. This makes a total of four different symbols and four different signals with same carrier frequency but with different phase (e.g. 0^o, 90^o, 180^o, 270^o). This way, the amount of information transmitted per second increases by two, when compared to BPSK. Typical QPSK waveform is given in Figure 2.5.

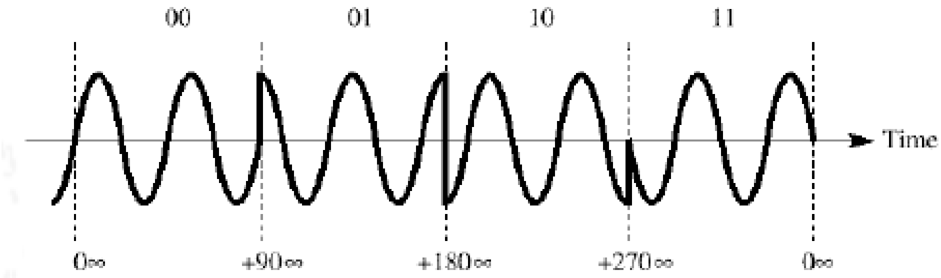

Figure 2.5: QPSK Waveform

2.2 Spectral Density

Spectral density of a signal is the characterization of the signals power or energy distribution on the frequency domain. It is important in communication systems due to distinction between channels. The filtering operations are done using this fact. Evaluation of signal and noise at the filter output uses the energy spectral density (ESD) and power spectral density (PSD).

2.2.1 Energy Spectral Density

For a real signal g(t) the energy E_g is defined as

$$E_g = \int_{-\infty}^{\infty} g^2(t)dt \qquad \textbf{(2.20)}$$

Parseval's relation states that the total energy in the signal may be determined either by computing the energy per unit time ($|g(t)|^2$) and integrating over all time or by computing the energy per unit frequency ($|G(w)|^2/2\pi$) and integrating over all frequencies.

$$E_g = \int_{-\infty}^{\infty} g^2(t)dt = \frac{1}{2\pi}\int_{-\infty}^{\infty} |G(\omega)|^2\, d\omega \tag{2.21}$$

$|G(w)|^2$ is referred to as energy spectrum density of the signal g(t), taking into account that g(t) is a finite energy signal.

$$E_g = \int_{-\infty}^{\infty} |G(f)|^2\, df \tag{2.22}$$

Energy spectral density describes the signal energy per unit bandwith measured in joules/Hertz. Since g(t) is a real signal, G(f) is an even function of frequency and there are equal energy contributions from both the positive and the negative frequency components. Therefore the total signal energy of g(t) can be written as

$$E_g = 2\int_{0}^{\infty} |G(f)|^2\, df \tag{2.23}$$

2.2.2 Power Spectral Density

If a signal g(t) exists over the entire interval $(-\infty,\infty)$, the power P_g of a real signal g(t) can be defined as the average power dissipated in a 1 ohm resistor when a voltage g(t) is applied across it. P_g can be written as

$$P_g = \lim_{T\to\infty}\frac{1}{T}\int_{-T/2}^{T/2} g^2(t)dt \tag{2.24}$$

If the energy of g(t), E_g, is finite, then the signals power is zero and if P_g is finite then E_g is infinite. Signals for which E_g is finite are said to be energy signals and the signals with nonzero and finite P_g are known as power signals.

There are some signals that cannot be classified as either energy or power signals because both E_g and P_g are infinite, such as,

$$g(t) = e^{-at} \qquad -\infty < t < \infty \tag{2.25}$$

In order to find the frequency-domain expression for the power P_g , it is observed that power signals have infinite energy and therefore may not have Fourier transforms. In this case we can consider the truncated signal $g_T(t)$ and define it as

$$g_T(t) = \begin{cases} g(t) & |t| \leq \dfrac{T}{2} \\ 0 & |t| > \dfrac{T}{2} \end{cases} \tag{2.26}$$

As T is finite, $g_T(t)$ has finite energy, and its Fourier transform can be taken. Letting

$$g_T(t) \leftrightarrow G_T(\omega) \tag{2.27}$$

The energy E_T of $g_T(t)$ is given by

$$E_T = \int_{-\infty}^{\infty} g_T{}^2(t)dt = \frac{1}{2\pi}\int_{-\infty}^{\infty} |G_T(\omega)|^2 d\omega \tag{2.28}$$

Hence the power P_g is given by

$$P_g = \lim_{T\to\infty}\frac{E_T}{T} = \lim_{T\to\infty}\frac{1}{T}\left[\frac{1}{2\pi}\int_{-\infty}^{\infty} |G_T(\omega)|^2 \, d\omega\right] \tag{2.29}$$

As T increases, E_T the energy of $g_T(t)$ also increases. Thus $|G_T(\omega)|^2$ increases with T, and as $T\to\infty$, $|G_T(\omega)|^2$ also approaches infinity. However, $|G_T(\omega)|^2$ must approach the infinity at the same rate as T, because for a power signal, the integral on the right hand side of the equation (2.29) must converge. This convergence lets us to interchange the order of the limiting process and integration, yielding

$$P_g = \frac{1}{2\pi}\int_{-\infty}^{\infty} \lim_{T\to\infty} \frac{|G_T(\omega)|^2}{T} d\omega \qquad \textbf{(2.30)}$$

Then the power spectral density is defined as

$$S_g(\omega) = \lim_{T\to\infty} \frac{|G_T(\omega)|^2}{T} \qquad \textbf{(2.31)}$$

Power of the signal can be re-written as

$$P_g = \frac{1}{2\pi}\int_{-\infty}^{\infty} S_g(\omega) d\omega \qquad \textbf{(2.32)}$$

Power spectral density (PSD) is a positive, even and real function of ω. The unit of power spectral density is watt/Hz. It describes the amount of power per unit of frequency.

2.3 Eye Diagram

The eye diagram is an oscilloscope display of the signal, repetitively sampled to get a good representation of its behavior. It is a useful tool for the qualitative analysis of signal used in digital transmission. It provides evaluation of the system performance and can offer insight into the nature of channel imperfections. The analysis of eye diagram can give approximations of signal to noise, clock timing jitter and skew.

In the radio communications, digital data consists of train of logical ones and zeros either referenced to zero volts return-to-zero (RZ), or with no voltage reference non-return-to-zero (NRZ). In either case these pulses contain considerable amount of energy in their harmonics. In order to reduce interference in the radio channels, the bandwidth is limited. Otherwise the harmonic energy in the data signal would create corresponding modulation sidebands that extend well beyond the intended bandwidth of the allocated communication channel. In order to reduce the unwanted sidebands, the data signal must be filtered in a manner that reduces harmonic energy while maintaining the integrity of the transmitted data. The eye diagram can be used after filtering to assure that the filter is behaving properly. General use of the eye diagram is at the receiver side to evaluate the received signal quality. Impairments to the signal can occur in the transmitter, through the frequency conversion and amplifier chain, propagation path, receiver front end, IF circuits and baseband signal processing. The timing errors on the receiver or the transmitter can be isolated with tests taken on each equipment.

The basic information contained in the eye diagrams is the size of the eye openings, which give information about the, the magnitude of the amplitude and timing errors.

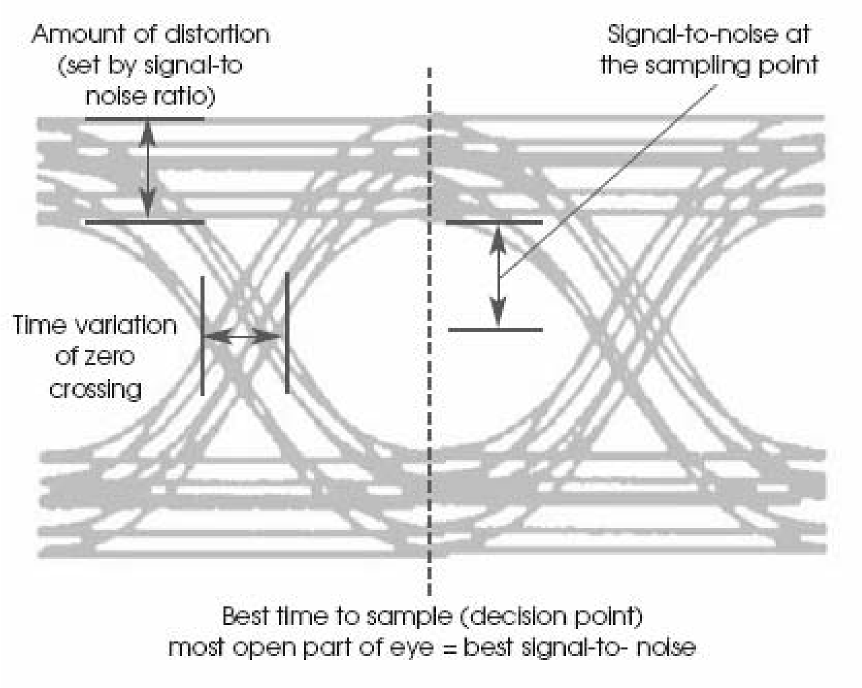

Figure 2.6: Eye Diagram

The eye diagram of a non-filtered signal is nearly square. The filtered signal has an eye diagram, which has smooth transitions. After the addition of timing errors and noise in the transmitter and receiver, as well as the channel imperfections, the eye diagram takes the form as in the Figure 2.7.

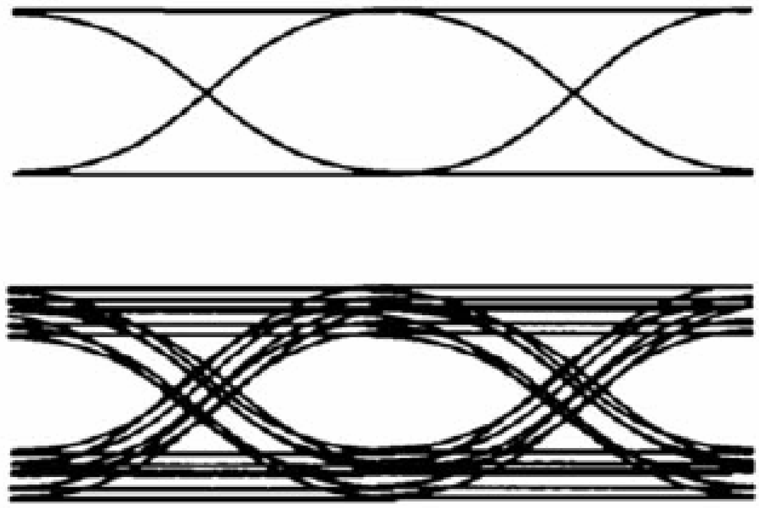

Figure 2.7: Eye Pattern After Filtering and at the Receiver Output

2.4 Performance Degrading Factors in Digital Communications

2.4.1 Noise

Noise is the unwanted electrical signals that are always present in an electrical system. The noise on a signal tends to mask the original signal. In mobile communications, the receiver part of the system is highly related with the noise concept, as it limits the ability of the receivers correct symbol decisions. This causes the limitation on the rate of information transmission. The noise can be either man made or caused by natural facts.

Filtering, shielding and the choice of the type of modulation can eliminate reasonable amount of noise. However there is natural source of noise called thermal noise that cannot be eliminated. Thermal noise is caused by the thermal motion of the electrons in dissipative components such as resistors, wires.

Thermal noise can be modeled as zero mean Gaussian random process. A Gaussian process n(t) is a random function whose value n at any arbitrary time t is characterized by the Gaussian probability density function

$$p(n) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{n^2}{2\sigma^2}} \qquad \textbf{(2.33)}$$

where σ^2 is the variance of n. The normalized Gaussian density function is obtained by assuming that $\sigma = 1$.

Gaussian distribution is often used as the system noise model, as the central limit theorem states that under general conditions, the probability distribution of the sum of i statistically dependent random variables approach the Gaussian distribution as $i \rightarrow \infty$. Therefore, even though the individual noise mechanisms might have distributions other than Gaussian distribution, the aggregate of many mechanisms will tend toward the Gaussian distribution.

2.4.1.1 Additive White Gaussian Noise (AWGN)

The primary spectral characteristic of thermal noise is that the power spectral density is same for all frequencies. Thermal noise sources contribute equal amount of noise power per unit bandwidth for all frequencies. A simple model for thermal noise can be represented as

$$G_n(f) = N_0 / 2 \qquad \text{(Watts/Hertz)} \qquad \textbf{(2.34)}$$

The power spectral density is constant for all frequencies. As the noise power has such a uniform spectral density, it is called "White Noise" in the same sense as the white light, which contains equal amounts of all frequencies that are present in the visible region of the electromagnetic spectrum.

The average power P_n of white noise is infinite as it has infinite bandwidth.

$$P_n = \int_{-\infty}^{\infty} \frac{N_0}{2} \, df = \infty \qquad \textbf{(2.35)}$$

Although white noise is a useful abstraction, the noise process cannot be truly white. However the noise encountered in many systems can be assumed to be approximately white. If the noise is observed on a real system, it can be seen that bandwidth of the noise is larger than that of the system. Therefore the noise can be considered to have infinite bandwidth.

The equation (2.35) states that two different samples of a white noise process are uncorrelated. Since thermal noise is a Gaussian process and the samples are uncorrelated, the noise samples are also independent. Thus the effects on detection process of a channel with additive white Gaussian noise (AWGN) are independent. This means that the channel is a memoryless channel. The term additive stands for, noise is simply added to the signal and there are no multiple mechanisms of noise.

2.4.4.2 AWGN Channel & Channel Capacity

The AWGN channel is a random channel, whose output is a real random process

$$Y(t) = X(t) + N(t) \tag{2.36}$$

where X(t) is the input waveform, regarded as a real random process, and N(t) is a real white Gaussian noise process with single-sided noise power density N_0 which is independent of X(t).

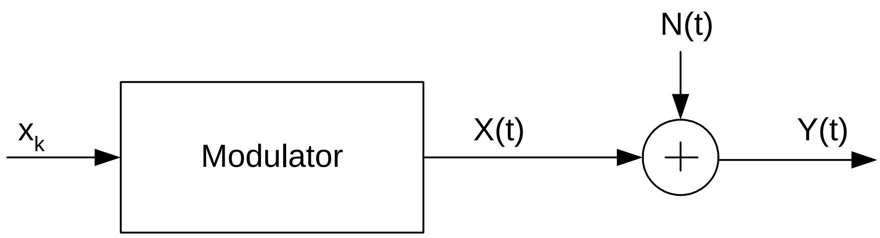

Figure 2.8: Noise Addition

The input X(t) is assumed to be both power-limited and band-limited. The average input power of the input waveform X(t) is limited to some constant P. The channel band B is a positive frequency interval with bandwidth W Hertz. The channel is said to be baseband if B=[0,W], and passband otherwise. The Fourier transform of any sample function x(t) of the input process X(t) is limited to B.

The signal to noise ratio, SNR, of the channel is then

$$SNR = \frac{P}{N_0 \cdot W} \qquad \textbf{(2.37)}$$

where N_0W is the total noise power in the band B. The parameter N_0 is the noise power per positive frequency Hz. Therefore the double sided power spectral density of N(t) is $S_{nn}(f)=N_0/2$ at least over the bands ±B.

The two parameters W and SNR turn out to characterize the channel completely for digital communication purposes. The capacity of any such channel in bits per second is

$$C = W \cdot \log_2(1 + SNR) \qquad \text{b/s} \qquad \textbf{(2.38)}$$

If a particular digital communication scheme transmits a continuous bit stream over such a channel at a rate R b/s, then the spectral efficiency of the scheme is said to be ρ= R/W (b/s)/Hz .The Shannon limit on spectral efficiency is therefore

$$C = \log_2(1 + SNR) \qquad \text{(b/s)/Hz} \qquad \textbf{(2.39)}$$

Therefore a reliable transmission is possible when $\rho < C_{[(b/s)/Hz]}$, but not when $\rho > C_{[(b/s)/Hz]}$.

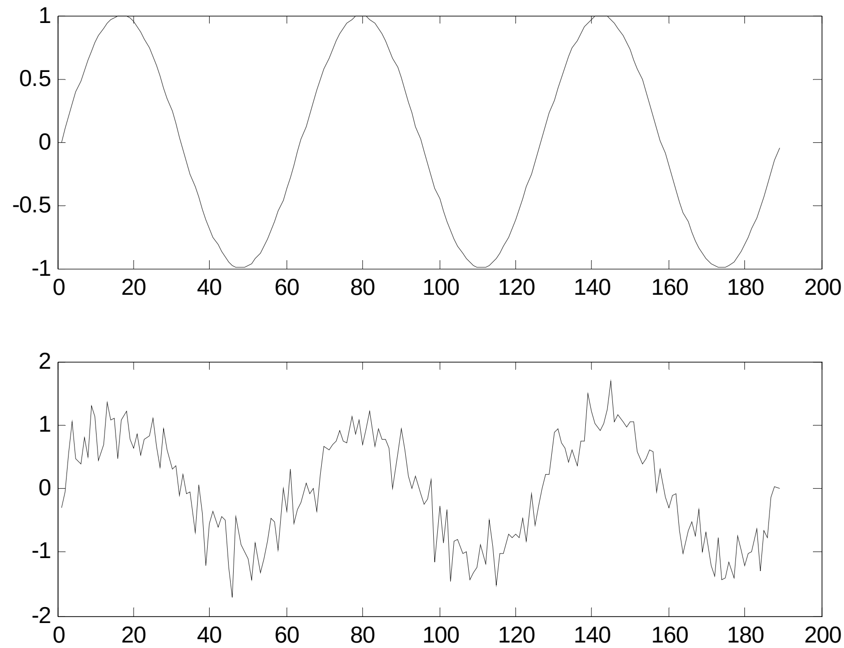

Figure 2.9: Original and AWGN Channel Output Signals at SNR = 10dB

2.4.2 Multipath Fading

The transmitted signal follows many paths before arriving at the receiving antenna, and it is the aggregate of these paths that constitutes the multipath radio propagation channel. The resulting signal strength will undergo large fluctuations, which, when the signal is small, results in a fade. This situation is referred as multipath fading. The multipath fading can be of two different categories.

1.The multipath signal paths are made up of a relatively small and identifiable number of components reflected by small hills, houses and other structures. This results in a channel model with a finite number of multipath components. This kind of channel is referred to as discrete multipath channel.

2.The multipath signal paths are generated by a large number of unresolvable reflections as might occur in mountainous areas or dense urban environment. This signal is composed of a continuum of unresolvable multipath components. This channel model is referred to as a diffuse multipath channel.

The real measured channels may contain both discrete and diffuse components. For modeling, these channels can be separated into the discrete and diffuse components.

A model for discrete multipath channels has the form

$$y(t) = \sum_n a_n(t) \cdot s(t - \tau_n(t)) \quad \textbf{(2.40)}$$

where s(t) is the bandpass input signal, $a_n(t)$ is the attenuation factor for the signal received from the nth path, and $\tau_n(t)$ is the corresponding propagation delay. If we express s(t) as

$$s(t) = \mathrm{Re}\left\{ \tilde{s}(t) \cdot e^{j2\pi f_c t} \right\} \quad \textbf{(2.41)}$$

then we can express the channel output as

$$y(t) = \mathrm{Re}\left\{ \left[\sum_n a_n(t) e^{-j2\pi f_c \tau_n(t)} \cdot \tilde{s}(t - \tau_n(t)) \right] e^{j2\pi f_c t} \right\} \quad \textbf{(2.42)}$$

and the complex envelope of the output is

$$\begin{aligned} \tilde{y}(t) &= \sum_n a_n(t) e^{-j2\pi f_c \tau_n(t)} \cdot \tilde{s}(t - \tau_n(t)) \\ &= \sum_n \tilde{a}_n(\tau_n, t) \cdot \tilde{s}(t - \tau_n(t)) \end{aligned} \quad \textbf{(2.43)}$$

From equation (2.43) we can define the multipath channel by a time varying, complex, low-pass equivalent impulse response

$$\tilde{c}(\tau_n(t), t) = \sum_n \tilde{a}_n(\tau_n(t), t) \cdot \delta(\tau - \tau_n(t)) \quad \textbf{(2.44)}$$

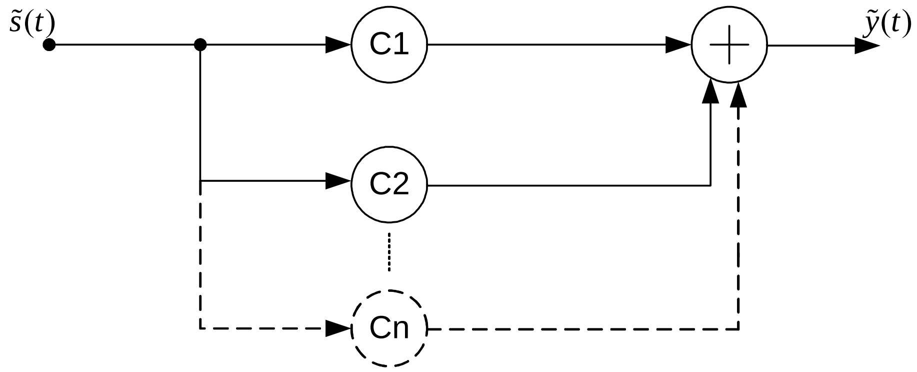

Figure 2.10: Multipath Fading Channel Model

2.4.3 Co-Channel Interference

In general the source of noise is a source, which primarily does not intend to produce electromagnetic disturbance patterns, for example microwave ovens or other electrical or electronic equipment. Another source of noise is given by the thermal effects existing for example in any electric circuit as in amplifiers.

Besides these sources of signal distortion, other communication systems might be active in the environment. Such sources, which have the primary goal to produce electromagnetic radiation for communication purposes are not represented by noise, instead they are referred to as interference. Like noise, the interference has an additive distorting impact on the signal. Interference occur in the radio systems due to the fact that bandwidth is limited and system users have to reuse certain spectra of the overall bandwidth.

If the two transmission devices operating within the same radio frequency band are active and a receiver, originally trying to receive the signal from one of the transmitters, also receives a weak signal from the second transmitter, this situation is referred to as Co-Channel Interference. If more than two or three interference sources are active, interference may be modeled as a white Gaussian process.

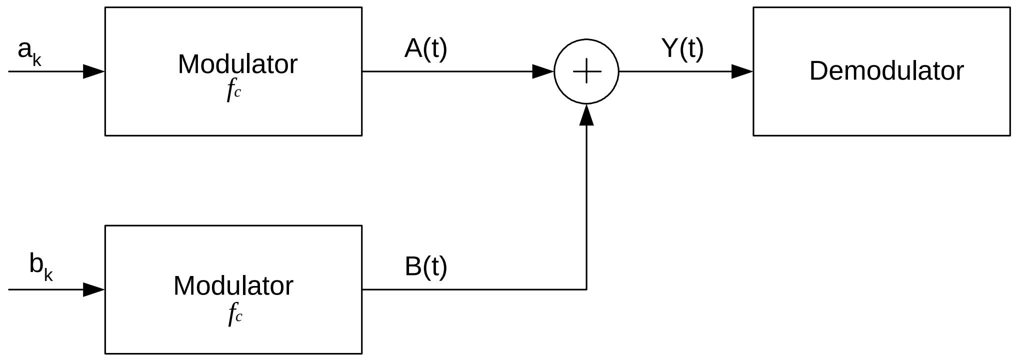

Figure 2.11: Co-Channel Interference Model

2.4.4 Intersymbol Interference

Intersymbol Interference (ISI) is a situation in which the energy from one symbol slot is spread out over neighbouring symbol slots. ISI may be introduced either by the channel, when the RMS delay spread becomes an appreciable fraction of the bit period, or by the filtering of the data pulses, in order to reduce the out of band power, before the modulation process.

Any practical channel has the inevitable filtering effect, which causes spreading of individual data symbols passing through the channel.

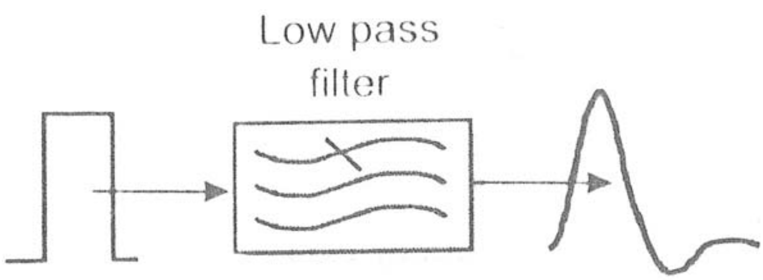

Figure 2.12: Filtering Effect of the Channel

When a sequence of signals spread, parts of the symbol energy overlap with neighbouring symbols causing intersymbol interference (ISI).

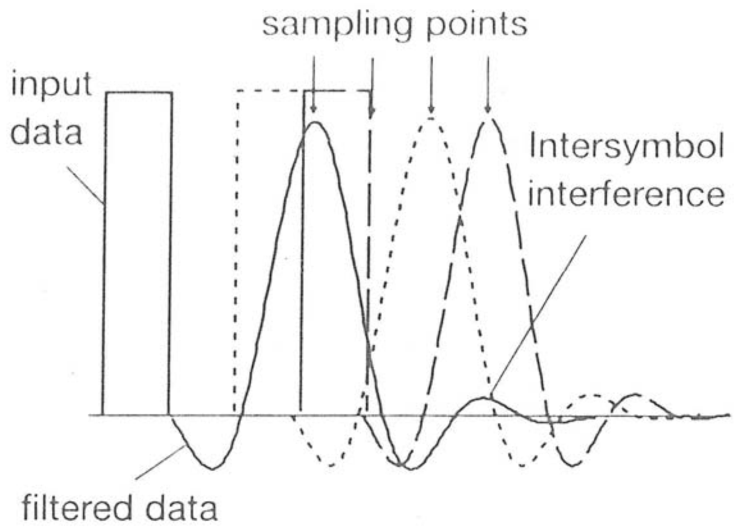

Figure 2.13: Filtered Pulses Interfering Neighboring Symbols

In some modulation techniques, data pulses are filtered, in order to suppress harmonic energy of the pulse and decrease bandwidth occupancy. This filtering process causes the symbols to spread out of the normal bit duration.

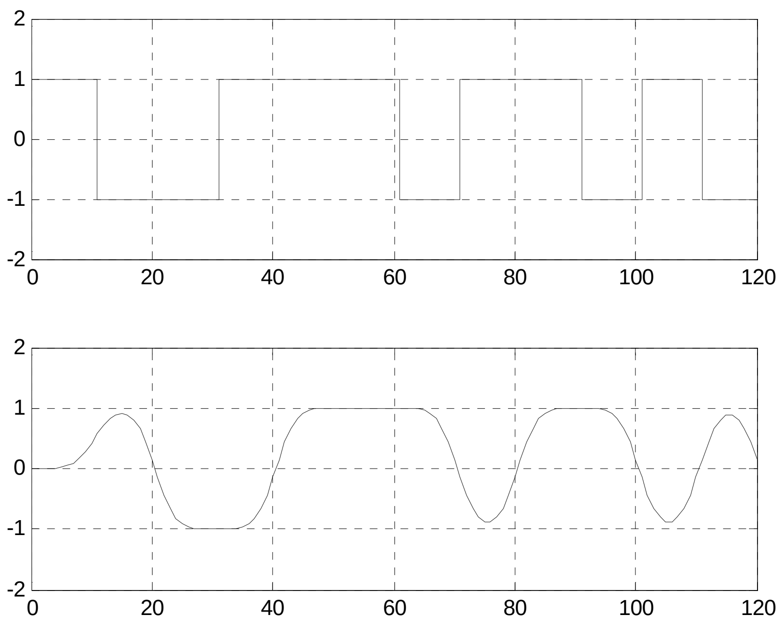

Figure 2.14: Filtered NRZ Data

ISI can significantly degrade the ability of the receiver to differentiate a current symbol from the diffused energy of the adjacent channels, therefore decreases the bit-error rate (BER) performance.

3. GMSK Modulation

3.1 Continuous Phase Modulation

Continuous phase modulation (CPM) is a form of digital phase modulation where the phase is constrained to remain continuous; that is, the phase cannot jump discontinuously between symbols, as it can in QPSK. A general CPM signal is given by

$$Y(t) = A\cos[\omega_c t + \phi(t) + \phi_0] \tag{3.1}$$

$$\tilde{Y}(t) = A\exp[j\phi(t) + j\phi_0] \tag{3.2}$$

where ϕ(t) is given as

$$\begin{aligned} \phi(t) &= 2\pi \int_{-\infty}^{t} \sum_{k=-\infty}^{n} d_k h_k g(\tau - kT)\, d\tau \\ &= 2\pi \sum_{k=-\infty}^{n} d_k h_k q(t - kT) \qquad nT \le t \le (n+1)T \end{aligned} \tag{3.3}$$

and g(t) is the frequency pulse, where as

$$q(t) = \int_{0}^{t} g(t)\, dt \tag{3.4}$$

The constraint imposed on (3.3) establishes the continuity of the phase. The parameter T is the symbol duration; $\{d_k\}$ is the data sequence where $d_k \in \{\pm 1, \pm 3, \ldots\ldots, (M-1)\}$; and h_k is called the modulation index. Usually h_k is constant but in some instances h_k varies with k in a cyclic manner where this situation is referred to as multi-h CPM.

In practice, g(t) is finite in extent,

$$g(t) = 0, \qquad t < 0, \qquad t > LT \tag{3.5}$$

and the normalization

$$\int_0^{LT} g(\tau)d\tau = \frac{1}{2} \tag{3.6}$$

is used.

When L=1, it is referred to as full response CPM and when L≥2 it is called partial response CPM.

Table 3.1: The Frequency Pulse Defined for Some CPM Schemes

LRC	$g(t) = \begin{cases} \frac{1}{2LT}\left[1-\cos\left(\frac{2\pi t}{LT}\right)\right], & 0 \le t \le LT \\ 0, & otherwise \end{cases}$
TFM	$g(t) = \frac{1}{8}\left[ag_0(t-T) + bg_0(t) + ag_0(t+T)\right]; \quad a=1,\ b=2$ $g_0(t) \approx \sin\left(\frac{\pi t}{T}\right)\left[\frac{1}{\pi t} - \frac{2-(2\pi t/T)\cot(\pi t/T) - \pi^2 t^2/T^2}{24\pi t^3/T^2}\right]$
LSRC	$g(t) = \frac{1}{LT}\frac{\sin(2\pi t/LT)\cos(\beta \cdot 2\pi t/LT)}{(2\pi t/LT)(1-[(4\beta/LT)t]^2)}; \quad 0 \le \beta \le 1$
GMSK	$g(t) = \frac{1}{2T}\left[Q\left(2\pi B_b \frac{t-T/2}{\sqrt{\ln 2}}\right) - Q\left(2\pi B_b \frac{t+T/2}{\sqrt{\ln 2}}\right)\right]; \quad 0 \le B_b T \le \infty$ $Q(t) = \int_t^{\infty} \frac{1}{\sqrt{2\pi}} \exp(-\tau^2/2)d\tau$
LREC	$g(t) = \begin{cases} 1/(2LT), & 0 \le t \le LT \\ 0, & otherwise \end{cases}$

3.2 Continuous – Phase Frequency-Shift-Keying

When the instantaneous frequency in each signalling interval is fixed and chosen from a set of M values, the method is called continuous-phase frequency-shift-keying, or CPFSK. In order to obtain a fixed instantaneous frequency in each signaling interval g(t) is set as

$$g(t) = \frac{1}{2T} p_T(t) \quad \textbf{(3.7)}$$

and $\phi(t)$ is obtained as

$$\phi(t) = \frac{1}{T}\pi h d_n(t - nT) + \pi h \sum_{k=-\infty}^{n-1} d_k, \qquad nT \le t \le (n+1)T \quad \textbf{(3.8)}$$

Generally h_k=h, a fixed value for modulation index is used, but it is not necessary in order to create a FSK signal.

The expression for the instantaneous frequency in CPFSK is given by

$$f_i(t) = \frac{1}{2\pi}\frac{d\phi(t)}{dt} = \frac{hd_n}{2T}, \qquad nT \le t \le (n+1)T \quad \textbf{(3.9)}$$

3.3 Minimum-Shift-Keying

Minimum-Shift-Keying (MSK) is a form of CPFSK, for which M=2 and h= 0.5.Using this specifications the phase function is obtained as

$$\phi(t) = \frac{\pi}{2T} d_n(t - nT) + \frac{\pi}{2}\sum_{k=-\infty}^{n-1} d_k, \qquad nT \le t \le (n+1)T \quad \textbf{(3.10)}$$

MSK can be thought of as a special case of OQPSK with sinusoidal pulse weighting.

The MSK signal can be defined as

$$s(t) = a_I(t)\cos\left(\frac{\pi t}{2T}\right)\cos 2\pi f_c t + a_Q(t)\sin\left(\frac{\pi t}{2T}\right)\sin 2\pi f_c t \qquad \textbf{(3.11)}$$

Figure 3.1 shows the various components of the MSK signal.

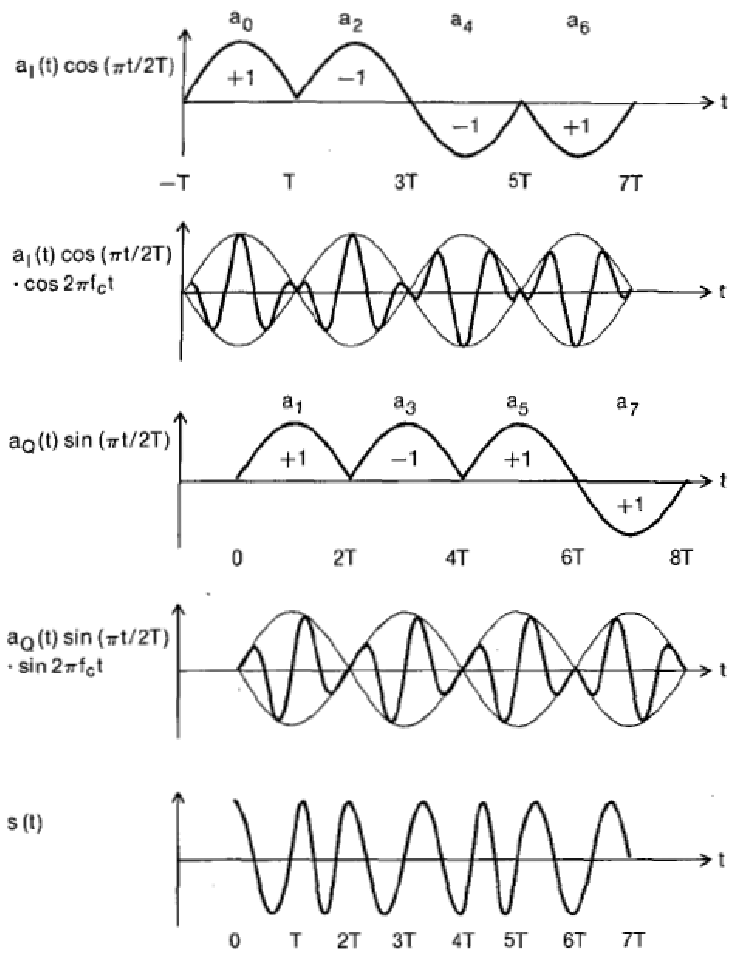

Figure 3.1: MSK Signal Formation

3.4 Gaussian Minimum Shift Keying

In mobile radio communications, the out of band radiation power in the adjacent channel should generally be suppressed about 70~80 dB than that of the concerning channel. In order to compact the bandwidth, manipulation is needed in the spectrum of the output signal. Smoothness of the signal is proportional with the amount of harmonics and their impact on the signal. Continuous phase change in a signal is related with the smoothness of the signal.

Gaussian minimum shift keying is a standard modulation used in global system for mobile (GSM) communication. It is a type of minimum shift keying which differs in usage of a premodulation filter, that is a low-pass filter, which has a Gaussian shaped impuls response.

In GMSK, data signals are passed through a low-pass filter before entering the phase modulation. The integral of the impulse response of a Gaussian filter output is quite smooth which causes the phase of the modulated signal to vary in a continuous manner.

GMSK has a low out of band power characteristic and a constant envelope which makes it a desirable choice for usage in the wireless mobile communications. The effect of the filter brings out the suppression of the out of band power by its sharp cut-off property.

GMSK finds a wide range of usage due to its spectral efficiency. As GMSK is a type of MSK scheme, it has a modulation index of 0.5. The Gaussian filter concentrates the energy on a desired band allowing for low out of band power characteristic. Widely known advantages of GMSK, that are narrow bandwidth and constant envelope modulation, make the GMSK suitable for both coherent and incoherent detection. Due to the constant envelope scheme, that is a property of GMSK, makes it less susceptible to fading environments than amplitude modulation and it requires inexpensive class-C amplifiers to be utilized for this scheme.

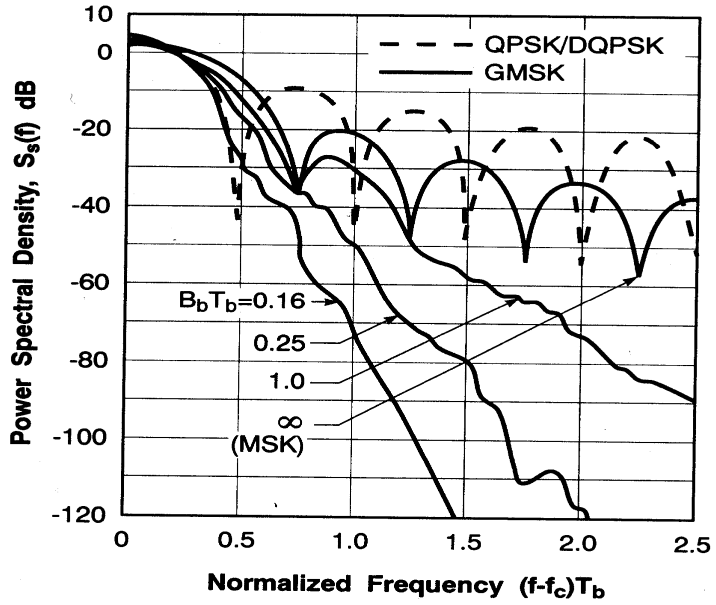

Figure 3.2: Comparison of Power Spectral Densities

Table 3.2: Occupied Bandwidth for Specified Percentage Power

B_bT \ Power%	90	99	99.9	99.99
0.2	0.52	0.79	0.99	1.22
0.25	0.57	0.86	1.09	1.37
0.5	0.69	1.04	1.33	2.08
MSK	0.78	1.20	2.76	6
TFM	0.52	0.79	1.02	1.37

3.4.1 Modulation

3.4.1.1 FM Modulator

GMSK can be considered as an FM modulation. Block diagram representation of GMSK modulation, using an FM modulator is given in Figure 3.3.

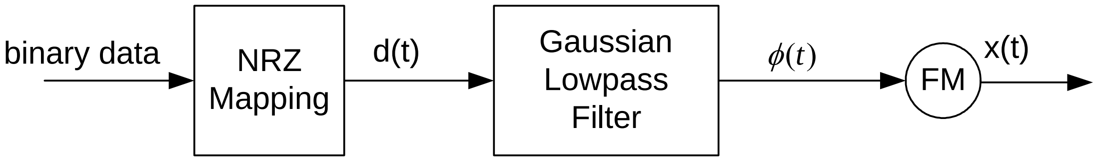

Figure 3.3: GMSK FM Modulator

The output of the FM modulator can be written as

$$x(t) = A_0 \cos(\omega_c t + \phi(t)) \quad \textbf{(3.12)}$$

where A_0 is the signal amplitude, w_c is the carrier frequency in radians per second and $\phi(t)$ is the transmit filtered data phase.

Binary data usually is in the unipolar form. Data may be represented as a voltage value for symbol "1" and zero voltage for the symbol "0". In this case the data has a DC value. In order to remove this DC before modulation, NRZ mapping technique is used. It is a simple technique, which removes the DC offset from the binary data and converts the data in the bipolar form. It simply assigns a voltage for the symbol "1" and the negative of that voltage for the symbol "0". The sequence is then spread into pulses and passed through the low-pass Gaussian filter. In case of two bit differential detection, which is used to reconstruct the symbols at the receiver side, the symbols must be encoded before modulation. However in one bit differential detection, no encoding is necessary. At the output of the Gaussian filter the phase function is obtained which modulates a carrier in the FM modulator block.

3.4.1.2 I/Q Modulator

In digital communications, modulation is often expressed in terms of I and Q. This is a rectangular representation of the polar diagram. On a polar diagram, the I axis lies on the zero degree phase reference and the Q axis is rotated by 90 degrees. The signal vector projection onto the I axis is its “I” component and the projection onto the Q axis is its “Q” component. I stands for the “In-phase” and Q for the “Quadrature” component of the signal.

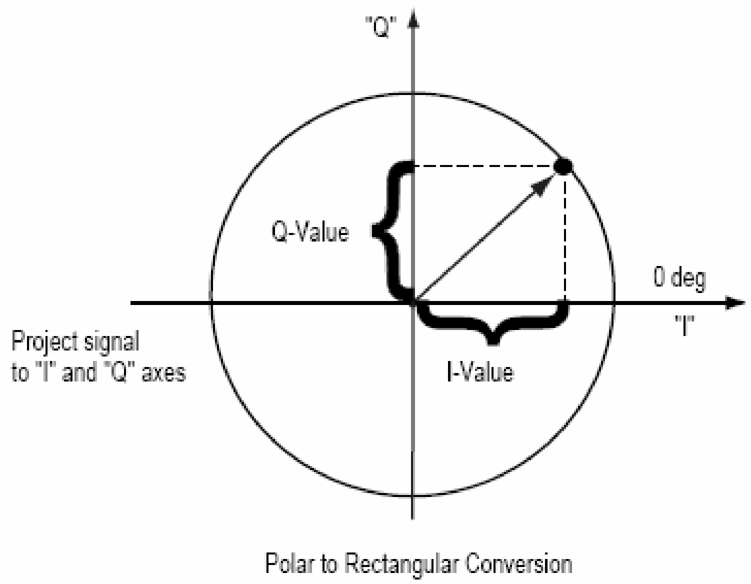

Figure 3.4: I/Q Representation of a Signal

I/Q diagrams are particularly useful because they mirror the way most digital communication signals are created using I/Q modulator. In the transmitter, I and Q signals are mixed with the same local oscillator. A 90 degree phase shifter is placed in one of the local oscillator paths. Signals that are separated with a 90 degree phase are also known as being orthogonal or quadrature to each other. Orthogonal signals do not interfere with each other. They are two independent components of the signal. When recombined, they are summed to form a composite output signal.

The two independent signals can be sent and received with simple circuits. This simplifies the design of a digital radio. The main advantage of the I/Q modulation is the symmetric ease of combining independent signal components into a single composite signal and later splitting such a composite signal into its independent component parts.

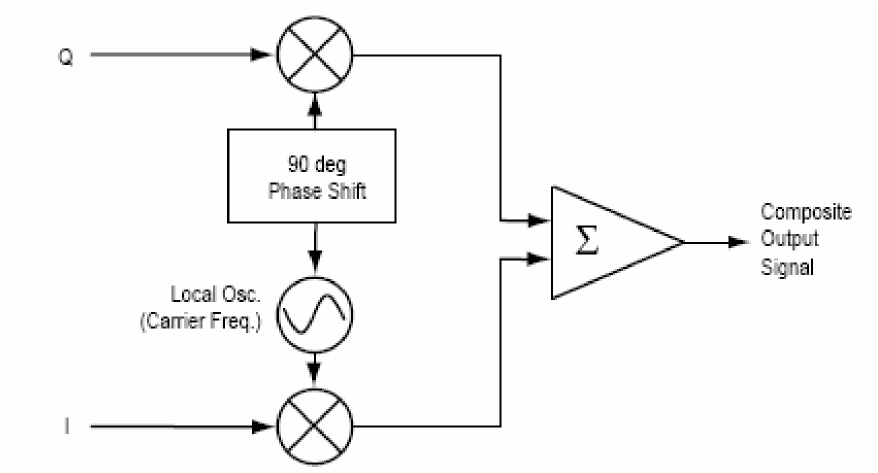

Figure 3.5: I and Q Forming Composite Output

GMSK modulator can also be implemented as an I/Q modulator. A block diagram of the modulator is given in Figure 3.6.

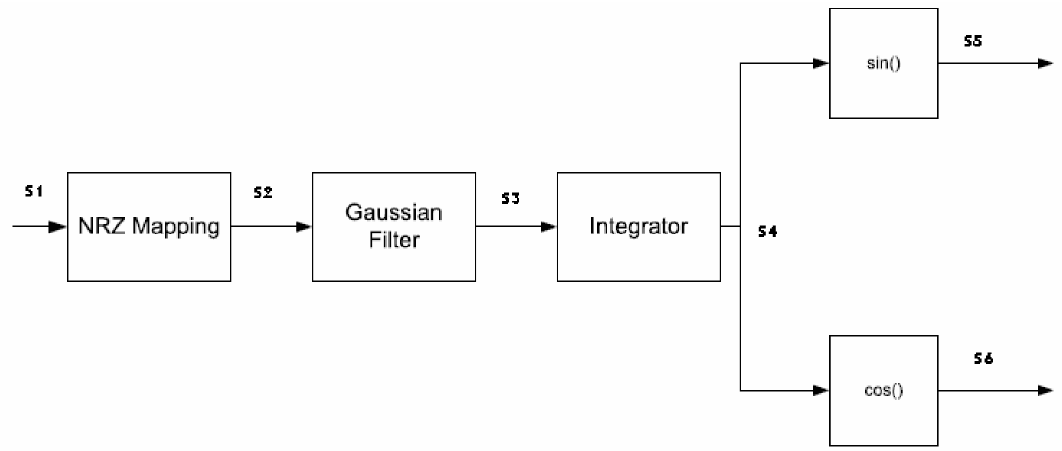

Figure 3.6: GMSK I/Q Modulator

The binary signals, which are in unipolar form arrive in the NRZ mapping block in order to be converted into bipolar form. The unipolar to bipolar conversion is shown in Figure 3.7.

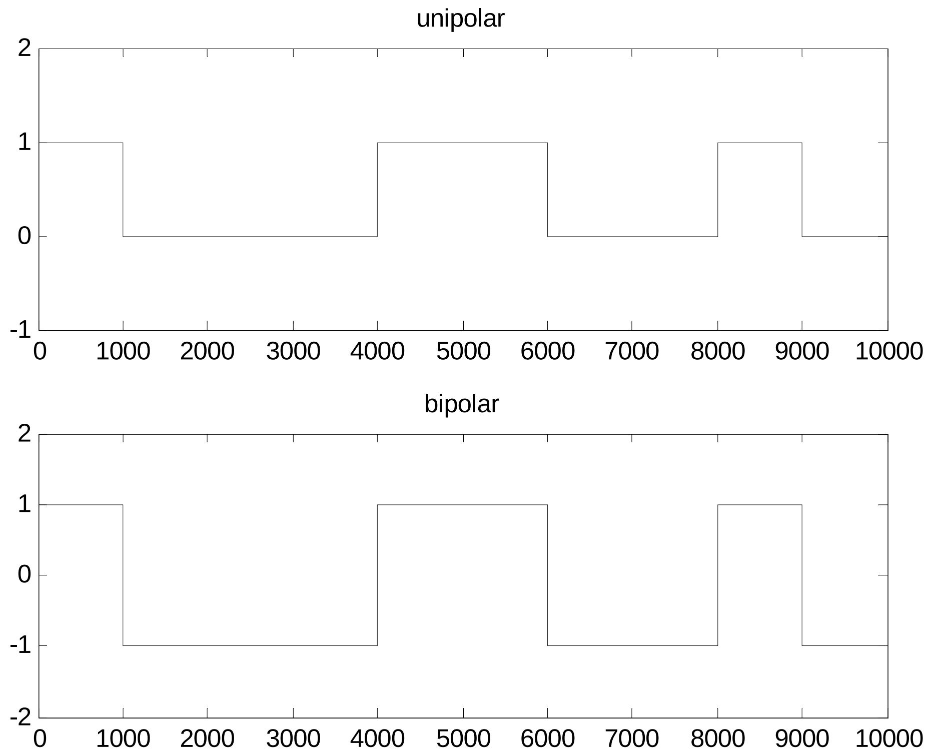

Figure 3.7: NRZ Mapping

NRZ process can be implemented in either discrete circuitry or in the processor.

The NRZ mapped zero DC valued signal is passed through the low-pass filter, which has a Gaussian shape impulse response. The impulse response of the filter function realizing the Gaussian pulse shape is given as

$$g(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left(-\frac{x^2}{\left(2\sigma^2\right)} \right) \tag{3.13}$$

$$\sigma = \sqrt{\frac{\ln(2)}{2\pi BT}} \tag{3.14}$$

BT is the bandwidth-time product related to the specified bandwidth of the designed low-pass filter and the time duration for a unit bit interval. BT product has a great influence on the ISI on the modulator. As the filtering process spreads the signal in time domain, the choice of BT is of importance due to the fact that it also influences the detection performance of the receiver. The truncated and scaled impulse response of the Gaussian filter is given in Figure 3.8.

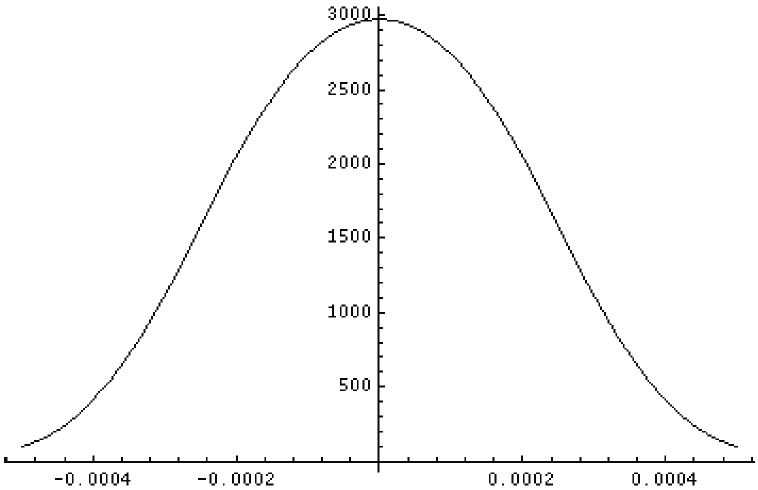

Figure 3.8: Gaussian Impulse Response

The output of the Gaussian filter is summed in order to obtain the accumulated phase. The summing process is utilized by the use of an integrator.

$$\phi(t) = \pi h \sum_{i=-\infty}^{n-L} \alpha_i \quad \textbf{(3.15)}$$

In this expression h stands for the modulation index and α_I stands for the symbols obtained from filtering. Time domain expression of the accumulated phase can be given as

$$\phi(t) = \frac{\pi}{2T_b} \int_{-\infty}^{T} \left[d(t) * h(t) \right] dt \quad \textbf{(3.16)}$$

where * denotes convolution of two functions. The $\pi/2$ factor in the equation (3.16) scales the phase such that the modulation scheme is minimum shift keying. In other words the modulation index is 0.5, which indicates that the maximum frequency deviation about the carrier frequency is half of the signaling rate.

The accumulated phase function $\phi(t)$ is used to obtain the inphase and quadrature components of the complex baseband signal. This is accomplished by passing the accumulated phase as arguments, to the sine and cosine functions.

$$I(t) = \cos(wt + \phi(t)) \quad \textbf{(3.17)}$$

$$Q(t) = \sin(wt + \phi(t)) \quad \textbf{(3.18)}$$

In order to obtain a complex baseband signal, I and Q components are mixed with the same oscillator, with a 90 degree phase shift between the mixing oscillator signals. This can be represented as,

$$x(t) = I(t)\cos(\omega_c t) - Q(t)\sin(\omega_c t) \quad \textbf{(3.19)}$$

3.4.2 Demodulation

In GMSK modulation, each symbol brings a change in the transmission signal phase. In order to obtain the transmitted symbols, the change in the phase of the transmission signals have to be found over each bit period. The phase difference can be written as

$$\Delta\phi_b(t) = \phi(t) - \phi(t - T_b) = \frac{\pi}{2T_b}\int_{t-T_b}^{t}\left[d(t) * h_t(t)\right]dt \quad \textbf{(3.20)}$$

It can be noted that the value of $\int[d(t)*h_t(t)]dt$ does not exceed T_b. This means that the maximum possible change in the phase is one bit period. The direction of the phase change either negative or positive corresponds with the symbols transmitted.

$\phi(t)$ has to be obtained in order to determine the transmitted symbols. Received RF signal is multiplied with the same carrier frequency that was used to upconvert the complex baseband signal. This downconversion yields the complex baseband signal.

In order to obtain the I and Q components of the composite signal, the signal is multiplied with the local oscillator with two components differing 90 degree in phase.

$$I(t) = \cos(2\pi f_c \mathrm{t}) \times \mathrm{R} \quad \textbf{(3.21)}$$

$$Q(t) = \sin(2\pi f_c \mathrm{t}) \times \mathrm{R} \quad \textbf{(3.22)}$$

The f_c stands for the oscillator frequency and R for the received composite signal. Demodulation of the signal can be accomplished by one-bit differential detection. In order to utilize one bit differential detection, it is not necessary to obtain the inphase and quadrature components of the signal. Detection can either be done using composite signal or the I/Q components of the signal.

3.4.2.1 Detection Using Composite Signal

Demodulation of the baseband composite GMSK signal can be accomplished by one bit differential detection technique as illustrated in the Figure 3.9.

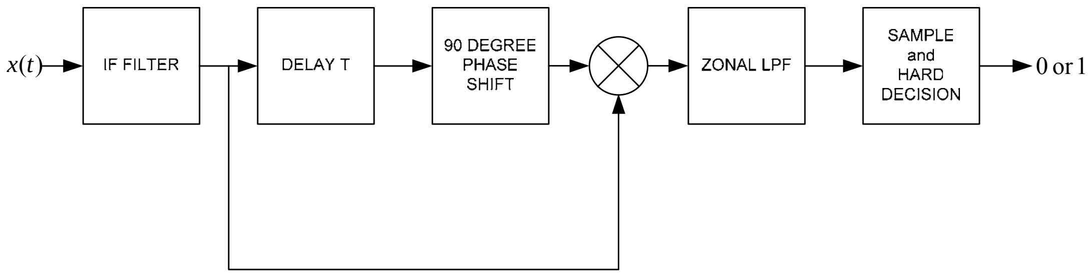

Figure 3.9: One Bit Differential Detector

The input x(t) is the arrived signal that can be represented as

$$x(t) = \sqrt{2S}\cos[\omega_0 t + \theta(t)] \quad \textbf{(3.23)}$$

where S is the signal power, $\omega_0 = 2\pi f_0$ is the center IF filter radian frequency, and $\theta(t)$ is the transmit-filtered data phase after modulation.

$$\theta(t) = \frac{\pi}{2T}\int [d(t) * h_t(t)]dt \tag{3.24}$$

The IF filter in the detector is used to bandlimit the noise in the original signal x(t) to obtain $x_{IF}(t)$.s

$$x_{IF}(t) = \sqrt{2S}a(t)\cos[\omega_0 t + \phi(t)] \tag{3.25}$$

where $[(2S)^{1/2}a(t)]$ is the time varying envelope and $\phi(t)$ is the distorted signal phase. $x_{IF}(t)$ can be rewritten in the polar form such as

$$x_{IF}(t) = R(t)\cos[\omega_0 t + \phi(t)] \tag{3.26}$$

In the delay block, a time delay is introduced at a period of one symbol duration, T. The phase shifter brings a 90 degree phase difference and $x_{IF}(t)$ is multiplied with its' delayed and phase shifted version, which results in

$$y(t) = \frac{R(t)R(t-T)}{2}\sin[\omega_0 T + \Delta\Phi(T)] \tag{3.27}$$

The phase difference is denoted as $\Delta\Phi(t)$, which is

$$\Delta\Phi(t) = \phi(t) - \phi(t-T) \tag{3.28}$$

and represents the change over a single symbol time.

It can be assumed that the carrier frequency be chosen as a multiple of 2π, this simplifies (3.27) to

$$y(t) = \frac{R(t)R(t-T)}{2} \sin \Delta\Phi(T) \qquad \textbf{(3.29)}$$

The receiver decides that a "1" was sent if y(t) > 0 and a "0" otherwise. Since the envelope R(t) is always positive, denoting the magnitude of the polar form, the decision rule is based on the sign of the sin[ΔΦ(t)] such that, if sin[ΔΦ(t)] > 0 , the decision is that a "1" is sent, and if sin[ΔΦ(t)] < 0, the decision is that "0" is sent.

3.4.2.2 Detection Using Decomposed I/Q Signals

In general, digital modulation schemes utilize decompositioning of the composite information bearing signals into their quadrature components before demodulation. The arriving composite signal is mixed with the local oscillator signal at the carrier frequency in two forms. One is at an arbitrary zero phase, and the other has a 90 degree phase shift. Thus the composite signal is broken into its two parts the in-phase and the quadrature part. These two components of the signal are independent and orthogonal. One can be changed without affecting the other. The Figure 3.10 illustrates the basic decomposition procedure.

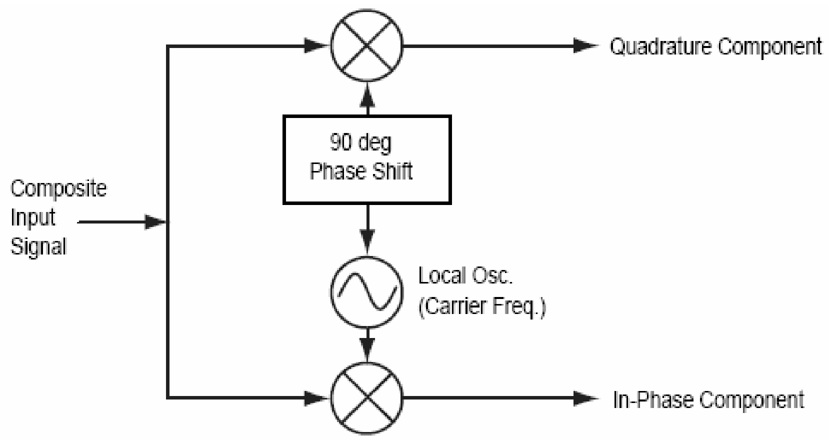

Figure 3.10: I/ Q Decomposition

In modulator, the time varying phase function $\phi(t)$ is used to generate the inphase and the quadrature components of the GMSK signal. The decomposed I/Q components in the demodulator is the same components that were generated in the modulator. The components can be written as

$$I = \cos[\phi(t)] \quad \textbf{(3.30)}$$

$$Q = \sin[\phi(t)] \quad \textbf{(3.31)}$$

The phase deviation over a single bit period gives the information of the sent symbol. Thus the the phase difference can be written as

$$\Delta\phi(T) = \phi(t) - \phi(t-T) \quad \textbf{(3.32)}$$

The decision is made through the sign of the phase difference which is obtained from

$$\sin\Delta\phi(T) = \sin[\phi(t) - \phi(t-T)] \quad \textbf{(3.33)}$$

The trigonometric identity of sine of the difference states that

$$\sin(a-b) = \sin a \cos b - \cos a \sin b \quad \textbf{(3.34)}$$

Using this identity, we can rewrite the (3.33) as

$$\sin[\phi(t) - \phi(t-T)] = \sin[\phi(t)]\cos[\phi(t-T)] - \cos[\phi(t)]\sin[\phi(t-T)] \quad \textbf{(3.35)}$$

The sine and cosine terms in the above expression can be replaced by the I/Q components of the signal such as

$$Q(t) = \sin[\phi(t)] \quad \textbf{(3.36)}$$

$$I(t-T) = \cos[\phi(t-T)] \quad \textbf{(3.37)}$$

I and Q signals and their delayed versions by a bit duration are used to estimate the phase difference in one bit interval. Then the phase difference is used to obtain sin[ΔΦ(t)], for the decision of the bit sent.

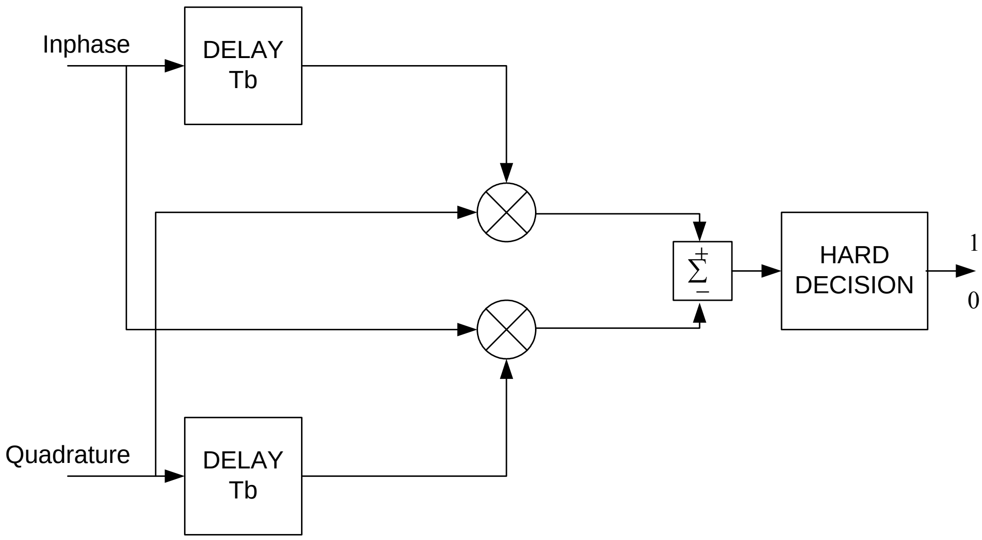

Figure 3.11: I/Q Detector

4. SYNCHRONIZATION

Symbol synchronization or timing recovery is a crucial part in detection of GMSK signals. Most of the problems rise due to the timing error between the transmitter and the receiver. The radio performance is usually degraded. Especially when the data is transmitted in burst mode, it is important to find fast and robust algorithms to estimate the timing offset and compensate for it.

In order to handle the timing recovery, a duration of bits called the preamble bits are sent by the transmitter, which degrade the transmission efficiency. This duration is used in order to synchronize for the incoming symbols and obtain the received data correctly. The symbol synchronization has to be accomplished in this duration. The data bits are obtained correctly if a successful synchronization of time is accomplished during the preamble bits.

4.1 Squaring Algorithm

The incoming sequence $\{r_k\}$ represents the signal obtained by sampling the complex envelope of a linear modulation at rate $1/T_s$ = N/T where T is the symbol transmission duration and N is the oversampling ratio. The symbol timing delay ε can be estimated by computing the complex Fourier coefficients at the symbol rate for every segment of $L_o N$ samples of $|r_k|^2$.The estimate ε_m is then given by

$$\varepsilon_m = \frac{1}{2\pi} \times \arg\left(\sum_{k=mL_0N}^{(m+1)L_0N-1} |\, r_k\,|^2 \; e^{-i2\pi k/N} \right) \tag{4.1}$$

where $\arg(x) \in \{-\pi, \pi\}$ denotes the phase of x and $\varepsilon_m \in \{-0.5, 0.5\}$.

If the original symbol timing delay ε is not within the range {-0.5, 0.5}, it actually corresponds to the case when the signal is shifted more than one symbol time. The estimate ε_m will then be given by the original timing delay ε added or subtracted by an integer such that the result falls in to the range {-0.5, 0.5}.

The squaring operation is equivalent to self convolution of the signal spectrum in the frequency domain. The spectrum at the output of the squaring operation contains spectral lines at $f = \pm 1/T$, which give the timing information. The strength of these spectral lines depends on the degree of spectral overlap when the signal spectrums are 1/T apart.

This symbol timing estimation is not necessarily effective for GMSK signals since GMSK is a nonlinear modulation. However the complex envelope of the GMSK signal can be approximated by a linear modulation with a pulse shape $C_o(t)$. That is, the GMSK signal s(t) can be approximated by

$$\begin{aligned} s(t) &\approx \sum_{N=0}^{\infty} \exp\left[j\pi h \sum_{n=0}^{N} a_n \right] C_0(t - NT - \varepsilon T) \\ &= \sum_{k=0}^{\infty} b_{2k+1} C_0(t - 2kT - T - \varepsilon'(2T)) + j \sum_{k=0}^{\infty} b_{2k} C_0(t - 2kT - \varepsilon'(2T)) \end{aligned} \qquad \textbf{(4.2)}$$

where

$$C_0(t) = \prod_{n=0}^{3} \frac{\sin[\Psi(t + nT)]}{\sin(h\pi)} \qquad \textbf{(4.3)}$$

with

$$\Psi(t) = \begin{Bmatrix} \pi \int_{-\infty}^{t} g(\tau) d\tau, & \mathrm{t < LT} \\ h\pi - \pi \int_{-\infty}^{t-LT} g(\tau) d\tau, & \mathrm{t \geq LT} \end{Bmatrix} \qquad \textbf{(4.4)}$$

and $b_{2k} = a_{2k}b_{2k} - 1$, $b_{2k+1} = -a_{2k+1}b_{2k}$ and $b_{-1} = 1$. In these expressions, a_n is the transmitted data of the GMSK signal, g(t) is the convolution between a Gaussian pulse and the rectangular pulse, and ε is the symbol timing delay of the GMSK signal with ε = 2ε`. Despite the linear approximation indicates the possibility of symbol timing estimation based on the squaring method, the squaring method may not work for GMSK signals. This is due to the small spectral density of $C_o(f)$ for $|f| > 0.5/T$. When the signal spectrums are 1/T apart, the spectral overlapping is not large enough to generate a stable estimate.

As the squaring method is commonly used in linear modulation with raised cosine pulse, one method to enhance frequency components of $C_o(f)$ for $|f| > 0.5/T$ is to design a matched filter such that

$$H_{MF}(f) = RC_{2T}(f)/C_0(f) \quad \textbf{(4.5)}$$

where $H_{MF}(f)$ is the frequency response of the matched filter and $RC_{2T}(f)$ is the spectrum of the raised cosine pulse with 3-dB cut-off frequency $f_c = 1/2T$.

For the case of GMSK with 0.3 BT product, the maximum value of the excess bandwidth of the raised cosine pulse is 0.2 due to the fact that if bandwidth is greater that 0.2, there are some frequencies in which $RC_{2T}(f)$ is finite but $C_o(f)$ is zero, making $H_{MF}(f)$ infinite at these frequencies. The match filter makes it possible to directly apply the squaring algorithm, but the performance is not satisfactory. In order to achieve the aim, a different form of this timing estimator can be defined.

The timing information can also be obtained by viewing a GMSK signal as a combination of two orthogonal linear modulations each with a symbol rate 1/2T and staggered with a time T. Timing delays can be estimated separately in inphase and quadrature channels and the two estimates can be subsequently combined to give the timing delay estimate for the GMSK signal, ε. The timing estimates from the inphase and quadrature components denoted as ε_I and ε_Q are

$$\varepsilon_I = 0.5 + \varepsilon' \quad \textbf{(4.6)}$$

$$\varepsilon_Q = \varepsilon' \quad \textbf{(4.7)}$$

In order to maintain generality $\varepsilon' \in \{-0.5, 0.5\}$ is assumed. If $0.5 \geq \varepsilon' > 0$, ε_I will be greater than 0.5, the estimate will be given by $0.5+\varepsilon'$ subtracted by 1 such that the result is in the range $\{-0.5, 0.5\}$ as

$$\varepsilon_I = 0.5 + \varepsilon' - 1 = \varepsilon' - 0.5 \quad \textbf{(4.8)}$$

$$\varepsilon_Q = \varepsilon' \quad \textbf{(4.9)}$$

and it follows that

$$|\varepsilon_I| + |\varepsilon_Q| = |\varepsilon' - 0.5| + |\varepsilon'| = 0.5 - \varepsilon' + \varepsilon' = 0.5 \quad \textbf{(4.10)}$$

If $-0.5 \leq \varepsilon' \leq 0$, both $0.5 + \varepsilon'$ and ε' will be in the range of $\{-0.5, 0.5\}$, and

$$|\varepsilon_I| + |\varepsilon_Q| = |\varepsilon' + 0.5| + |\varepsilon'| = 0.5 + \varepsilon' - \varepsilon' = 0.5 \quad \textbf{(4.11)}$$

It can be said that $|\varepsilon_I| + |\varepsilon_Q| = 0.5$. This means that when the value of $|\varepsilon_I|$ increases $|\varepsilon_Q|$ decreases and vice versa.

The proposed timing recovery can be given as

$$\varepsilon = \begin{cases} 2\varepsilon_I, & \text{if } |\varepsilon_I| \leq 0.25 \\ 2\varepsilon_Q, & \text{if } |\varepsilon_Q| < 0.25 \end{cases} \quad \textbf{(4.12)}$$

There are some cases when both $|\varepsilon_I|$ and $|\varepsilon_Q|$ are very close to 0.25 such as $|\varepsilon_I| = 0.24$ and $|\varepsilon_Q| = 0.26$, due to the varience of the estimation. In this case either $|\varepsilon_I|$ or $|\varepsilon_Q|$ can be chosen to give the estimation of the GMSK signal as they both give a value close to 0.5 or –0.5.

4.2 Joint Symbol Timing Error and Frequency Offset Estimation

The method achieves both timing error correction and frequency offset compensation. The converted baseband complex signal is first frequency discriminated and passed through a digital filter which performs FFT. The frequency offset can be estimated from the DC component of the FFT and the symbol timing error can be estimated from the phase angle of the FFT at a specified frequency, which is equal to an integral multiple of half of the bit rate. The two estimates are then used in frequency offset compensation and symbol timing recovery during the preamble period.

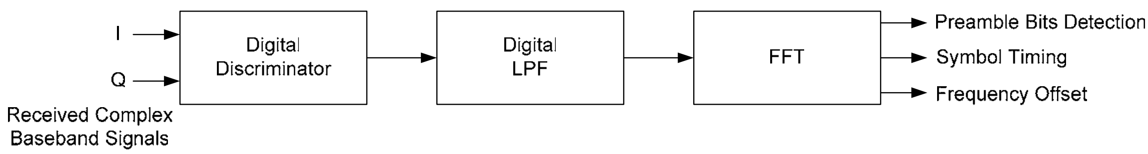

Figure 4.1: Block Diagram of the Synchronization Method

The RF signal is downconverted to baseband and decomposed into its real and imaginary parts I(t) and Q(t). These two signals are oversampled, digitally frequency discriminated and low-pass filtered in order to obtain the raw digital data. This data goes through a FFT for synchronization preamble bits detection. If the preamble is detected, both frequncy offset and sampling time error are estimated from the FFT results. Symbol timing synchronization is done in a feedforward manner and the frequency offset compensation is done in a hybrid manner. Frequency offset estimation is fed back to the VCO during the preamble period. This estimation can be used to change the decision threshold in a noncoherent detection mode or rotate the signal constellation in a coherent detection mode.

The received signal is assumed to be undistorted by the channel imperfections and predetection filter.

$$r(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t + \phi_s(t - \varepsilon T_s) + \phi_0) + n(t) \qquad \textbf{(4.13)}$$

where n(t) is an additive bandpass Gaussian noise with one sided power spectral density N_0, E_b and T_b are bit energy and bit period, φ_0 is an arbitrary phase, $\varphi_s(t)$ is the frequency modulated phase of the transmitted carrier, and εT_s is the time delay caused by channel filtering and corresponds to a time delay at which the signal should be sampled. ε is rounded off to an integer for implementing the synchronization algorithm digitally.

The noise term n(t) can be expressed as

$$n(t) = n_c(t)\cos(2\pi f_c t) - n_s(t)\sin(2\pi f_c t) \qquad \textbf{(4.14)}$$

and $n_c(t)+jn_s(t)$ is an equivalent base-band representation of n(t).
The term εT_s is limited to the interval $[-(T_b/2), T_b/2\ ]$. Ignoring the noise, the inphase and quadrature components of demodulated complex baseband signal are,

$$I(t) = \sqrt{\frac{E_b}{2T_b}} \cos(2\pi\Delta ft + \phi_s(t - \varepsilon T_s) + \theta) \qquad \textbf{(4.15)}$$

$$Q(t) = \sqrt{\frac{E_b}{2T_b}} \sin(2\pi\Delta ft + \phi_s(t - \varepsilon T_s) + \theta) \qquad \textbf{(4.16)}$$

where $\Delta f = f_c - f_c^{'}$ and θ are the frequency and the phase offsets produced by the receiver VCO at the centre frequency f_c. The inphase and quadrature signals are sent to the frequency discriminator and the discriminator output signal is

$$\psi(t) = \frac{I(t)\dot{Q}(t) - Q(t)\dot{I}(t)}{I(t)^2 + Q(t)^2} = 2\pi\Delta f + \dot{\phi}_s(t - \varepsilon T_s) \qquad \textbf{(4.17)}$$

When a training sequence 1010101010.... is received, the discriminator output signal will be periodic with a period of $2T_b$.

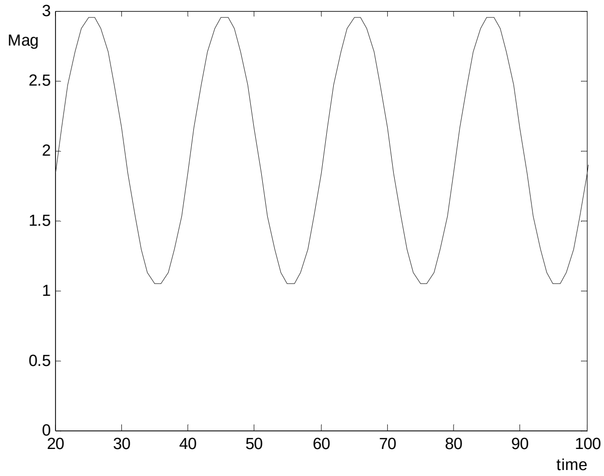

Figure 4.2: Discriminator Output Periodic Signal

This waveform consists of a DC term $2\pi\Delta f$ and a periodic component given as,

$$\dot{\phi}_s(t-\varepsilon T_s) = \sum_{l=-\infty}^{\infty} g_T(t-\varepsilon T_s - l\cdot 2T_b) \tag{4.18}$$

with a period of $2T_b$, where $g_T(t)$ is a DC free waveform of duration $T=2T_b$. If $\psi(t)$ is observed for L periods, it can be expressed as,

$$\psi(t) = 2\pi\Delta f + \sum_{l=0}^{L-1} g_T(t-\varepsilon T_s - l\cdot 2T_b) \tag{4.19}$$

$\psi(t)$ is sampled every T_s seconds and samples are denoted by $\psi[n]$,

$$\begin{aligned}\psi[n] = \psi(nT_s) &= 2\pi\Delta f + \dot{\phi}_s(nT_s - \varepsilon T_s) \\ &= 2\pi\Delta f + \sum_{l=0}^{L-1} g_T\left[n-\varepsilon - l\frac{2T_b}{T_s}\right]\end{aligned} \tag{4.20}$$

where $g_T[n]$ denotes the sample of $g_T(t)$ at $t = nT_s$. It is noted that $\psi[n]$ has a period of $M = 2T_b/T_s$.

N = LM is chosen, such that N-point discrete Fourier transform Ψ[k] of ψ[n] is obtained as,

$$\Psi[k] \equiv \sum_{n=0}^{N-1} \psi[n] \exp\left\{-j\frac{2\pi k}{N}\right\}, \qquad 0 \le k \le N-1 \tag{4.21}$$

If $g_T[n]$ is even and real with zero DC, then the frequency offset Δf and the sampling time error ε can be estimated by

$$\Delta f = \frac{1}{2\pi N}\sum_{n=0}^{N-1} \psi[n] \tag{4.22}$$

$$\varepsilon = -\frac{T_b / T_s}{\pi} \arg(\psi[L]) \tag{4.23}$$

The magnitude of Ψ[L] can be compared with a signal detection threshold to determine whether a training signal has arrived or not.

5. SIMULATION RESULTS AND PERFORMANCE EVALUATION

The proposed modem is simulated using MATLAB. The modulator is tested for two BT values, 0.3 and 0.5, the receiver performance is evaluated.

5.1 Modulator

Binary data is generated randomly and kept in a vector in order to calculate the BER at the receiver side. The generated data is spread over to samples by oversampling so that each symbol can be convolved by the filter.

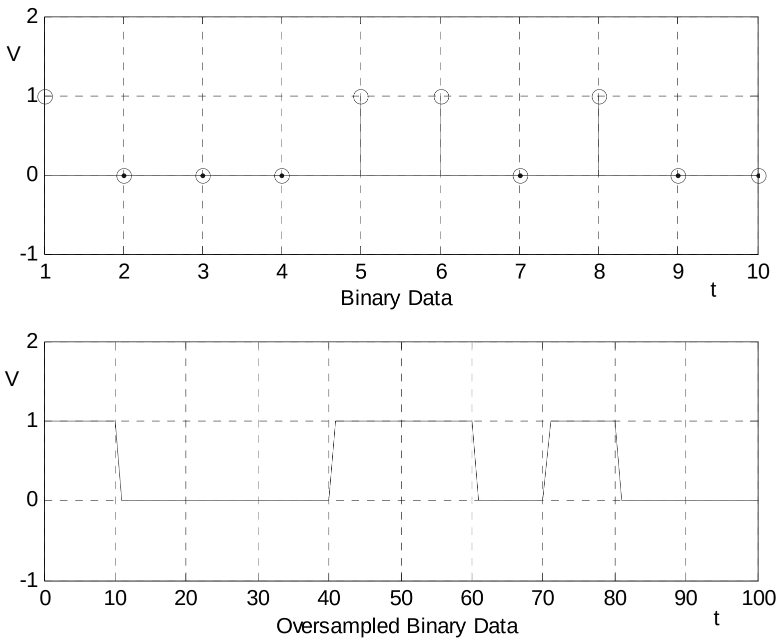

Figure 5.1: Oversampling Binary Data

The resulting waveform is NRZ mapped. This is done in order to remove the DC component from the signal. NRZ mapped data is convolved with the impulse response of the Gaussian low-pass filter.

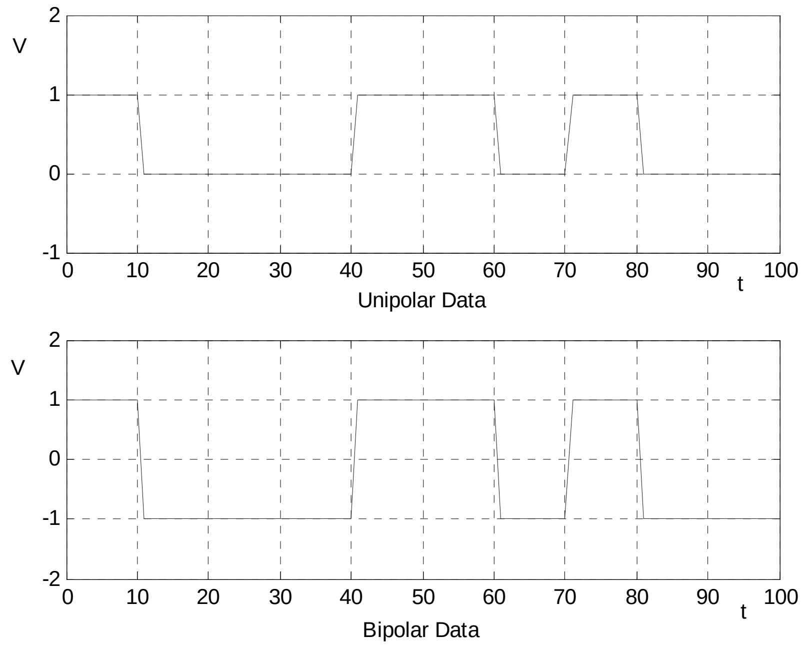

Figure 5.2: Unipolar to Bipolar Data Conversion

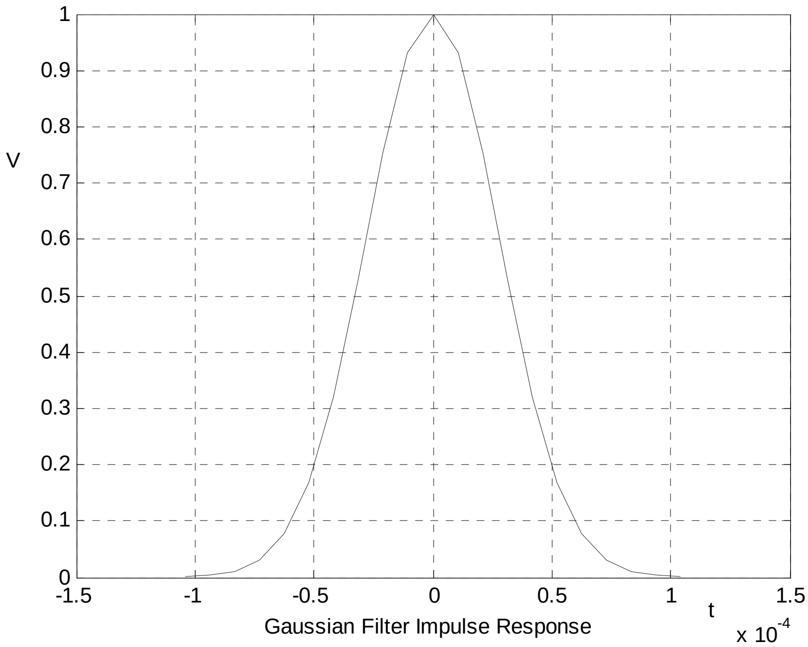

Figure 5.3: Impulse Response of the Gaussian Low-Pass Filter

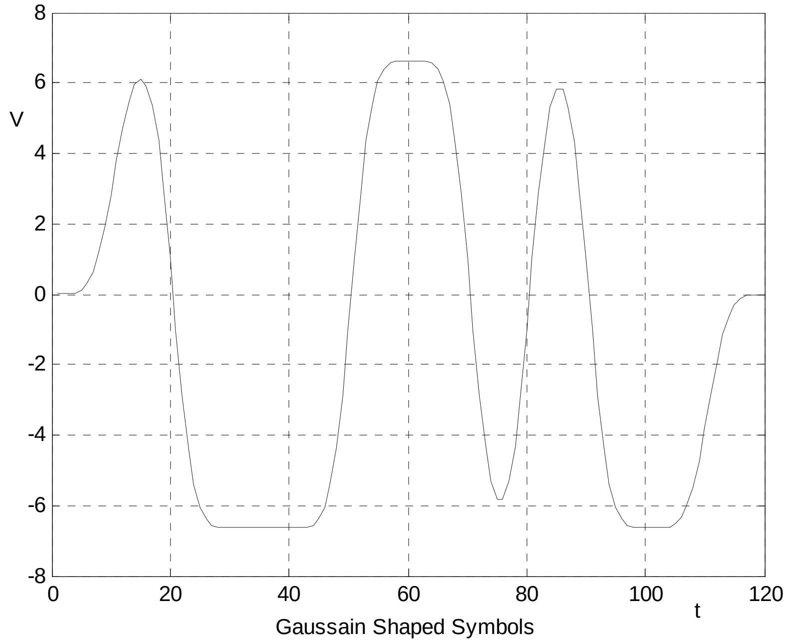

Figure 5.4: Smoothed Data

The filtered waveform is smoother as the harmonics of the square shaped signals are suppressed. This smooth waveform is then integrated in order to obtain the continuous phase function.

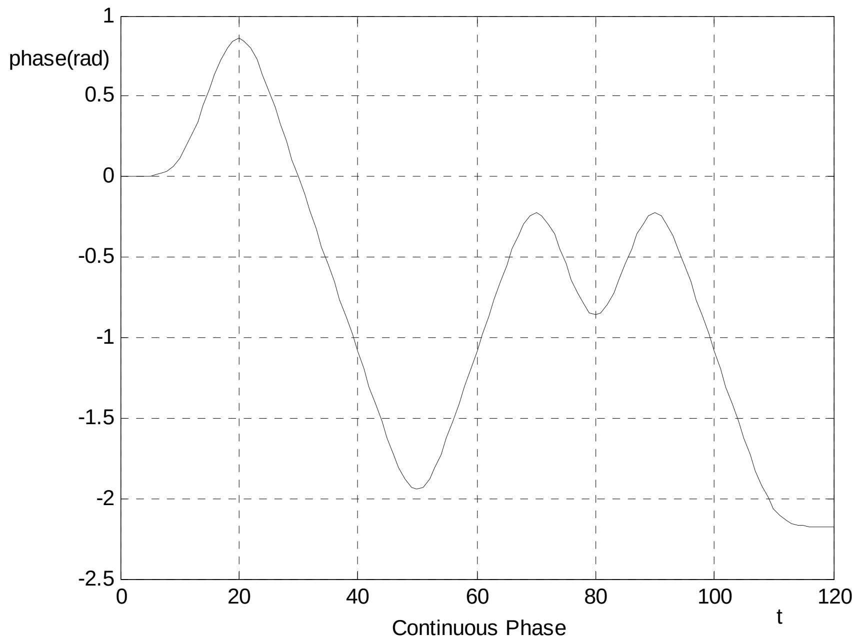

Figure 5.5: Accumulated Continuous Phase

Continuous phase is used to vary the phase of two identical oscillators, with 90 degree phase difference, in order to obtain the inphase and quadrature baseband signals, which make up the complex baseband GMSK signal.

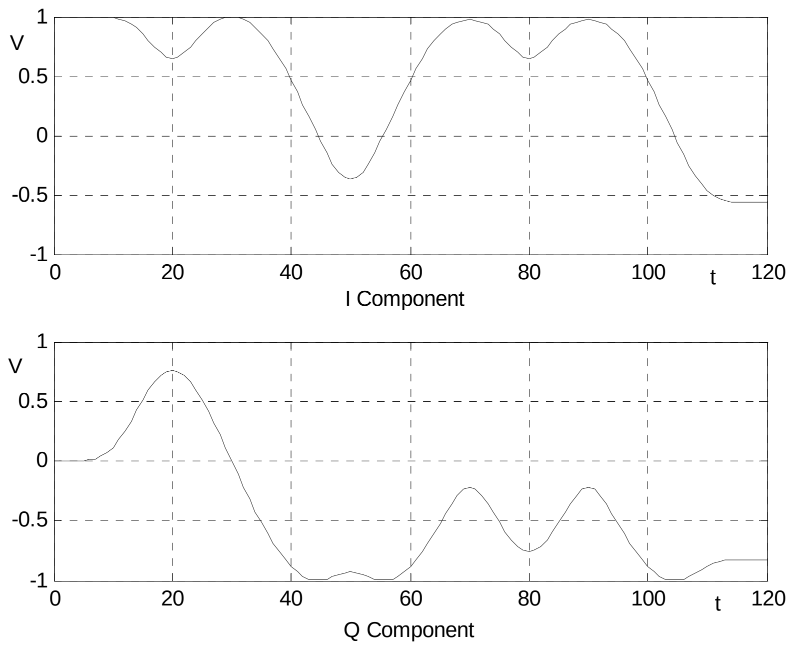

Figure 5.6: I/Q Signals

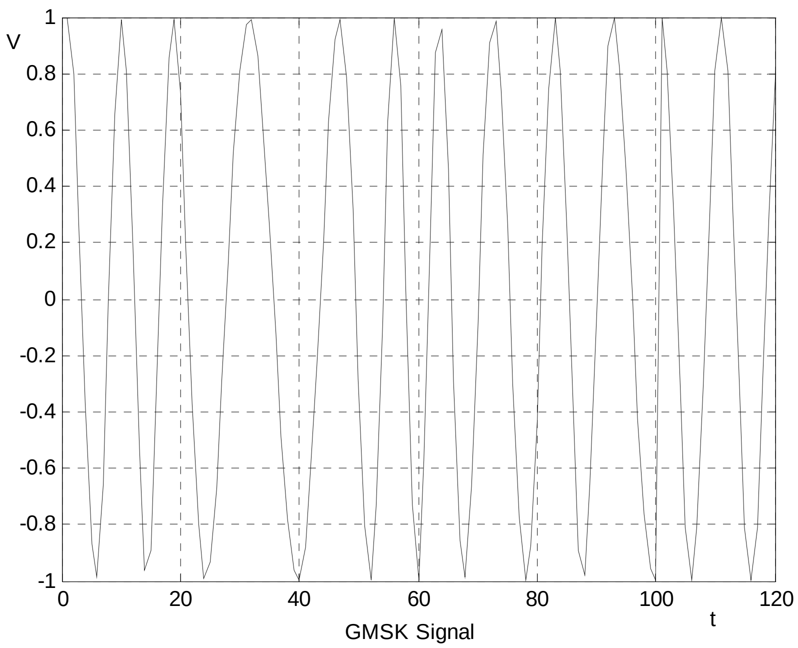

Figure 5.7: GMSK Signal

5.2 Demodulator

The complex baseband GMSK signal is delayed a bit period T_b, and 90 degree phase shifted. The obtained signal is multiplied with the original GMSK signal, in order to obtain the modulating symbols.

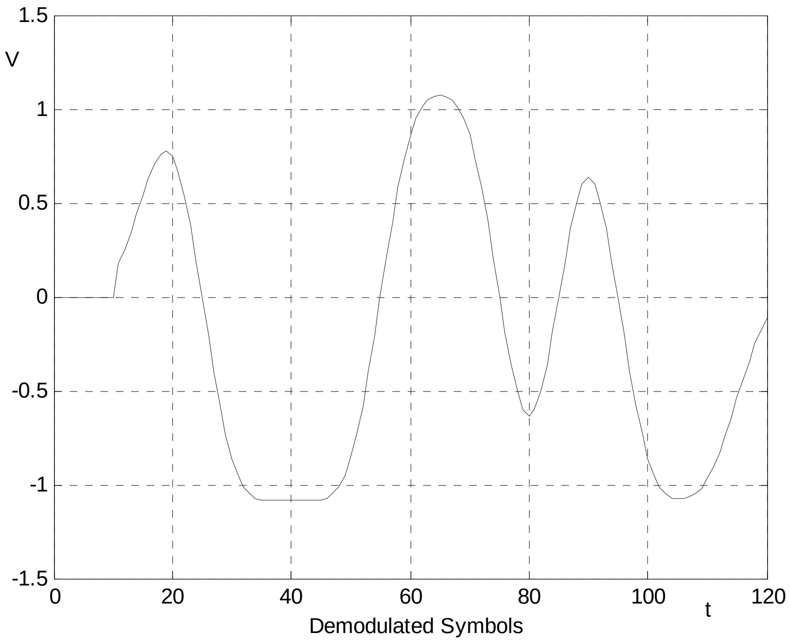

Figure 5.8: Demodulator Output

The resulting signals are sampled at a rate, which was used to oversample the binary data. Hard decision is made according to the sign of the sample taken. Obtained samples are the binary data that was sent by the modulator.

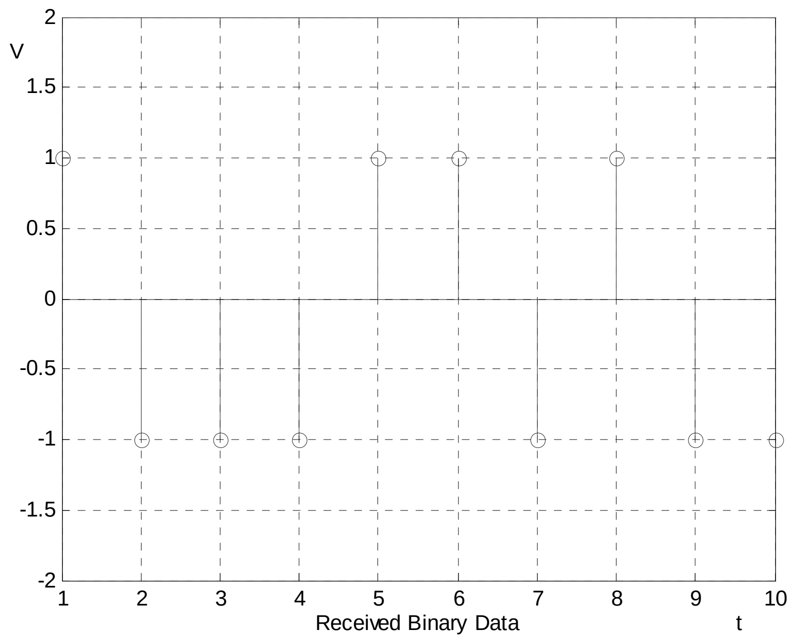

Figure 5.9: Hard Decision Output Received Data

5.3 AWGN Channel Performance

Random noise is generated and added to the original GMSK signal. The detection performance of the demodulator is tested for different levels of SNR.

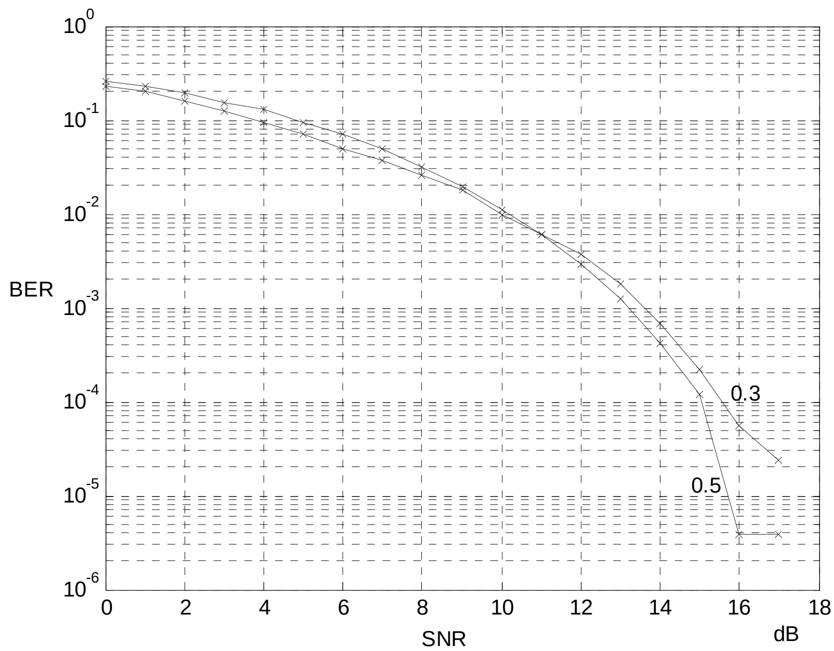

Figure 5.10: Performance in the Presence of Noise

5.4 Multipath Fading Performance

Three different channel models are used in order to simulate the system in the multipath fading environment. The simulations are made for BT=0.3 and BT=0.5.

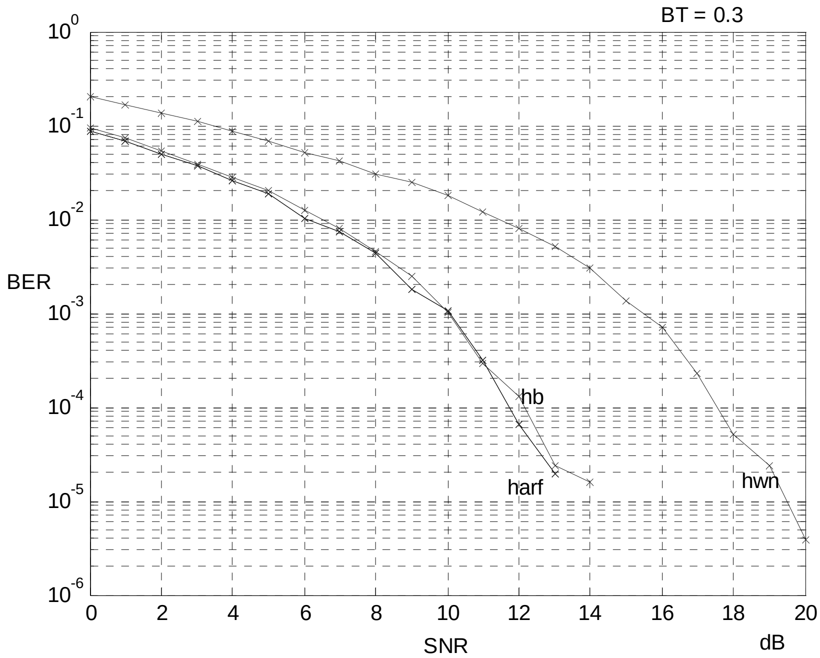

Figure 5.11: Performance at BT=0.3 for Multipath Fading Channel

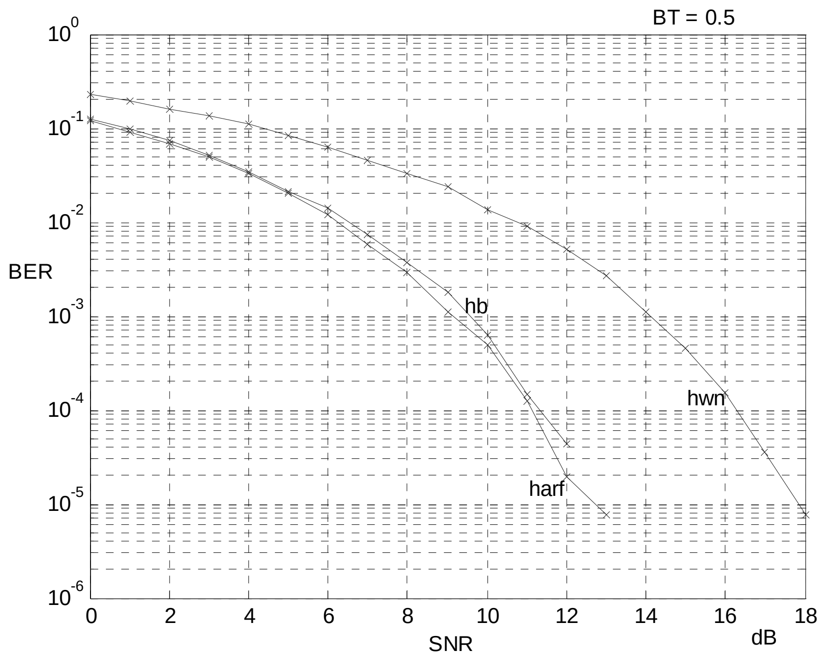

Figure 5.12: Performance at BT=0.5 for Multipath Fading Channel

5.5 Co-Channel Interference Performance

Different levels of interfering signals are added to the reception signal to obtain the demodulators detection performance.

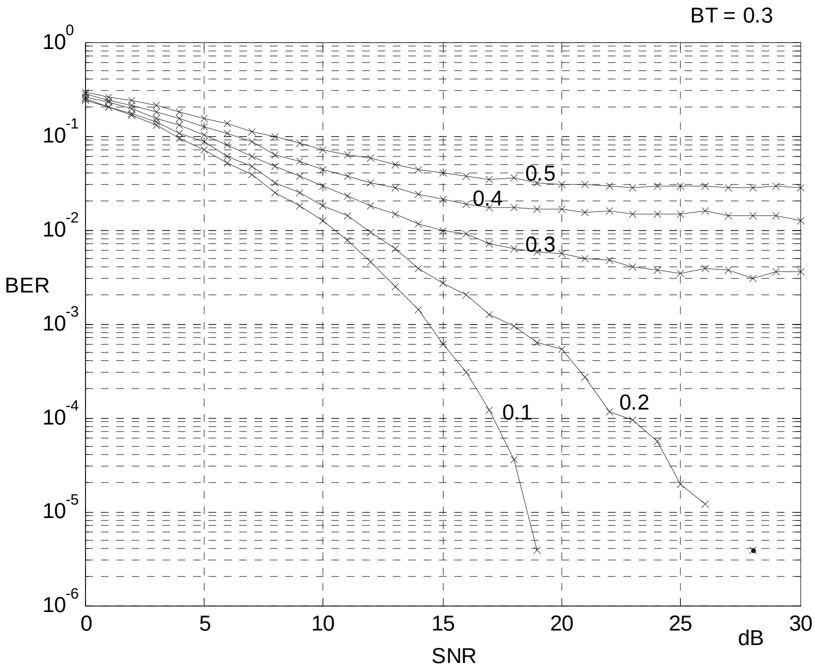

Figure 5.13: Performance at BT=0.3 for Co-Channel Interference

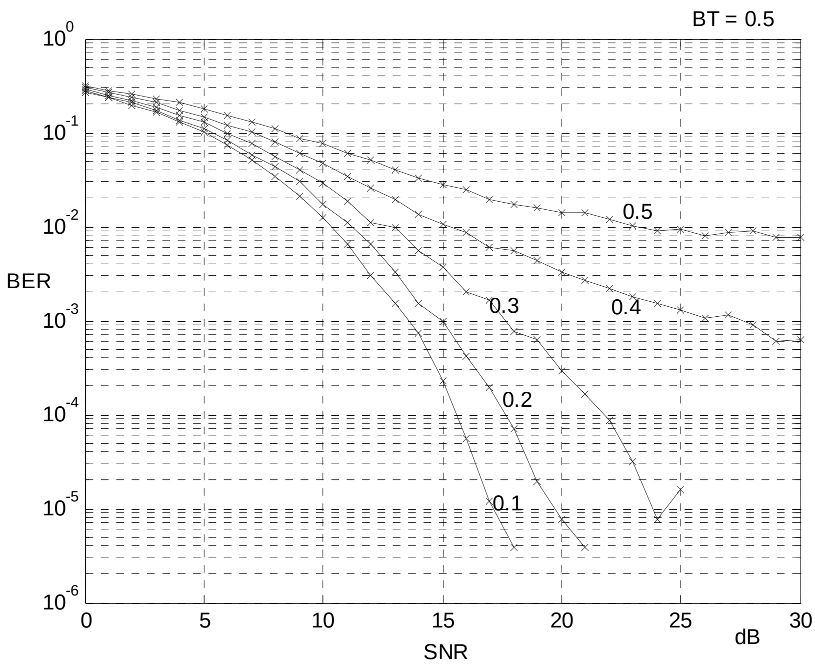

Figure 5.14: Performance at BT=0.5 for Co-Channel Interference

6. CONCLUSION

In this work, a GMSK modulator and demodulator are designed. The modulator and the demodulator are first modeled by MATLAB. Environmental facts such as noise multipath fading and co-channel interference is mathematically expressed and simulations are made in order to evaluate the overall system performance. Due to the asynchronous manner of the transmission, synchronization is done in order to compensate for the timing errors and frequency offsets. Finally, the system is coded on digital signal processor.

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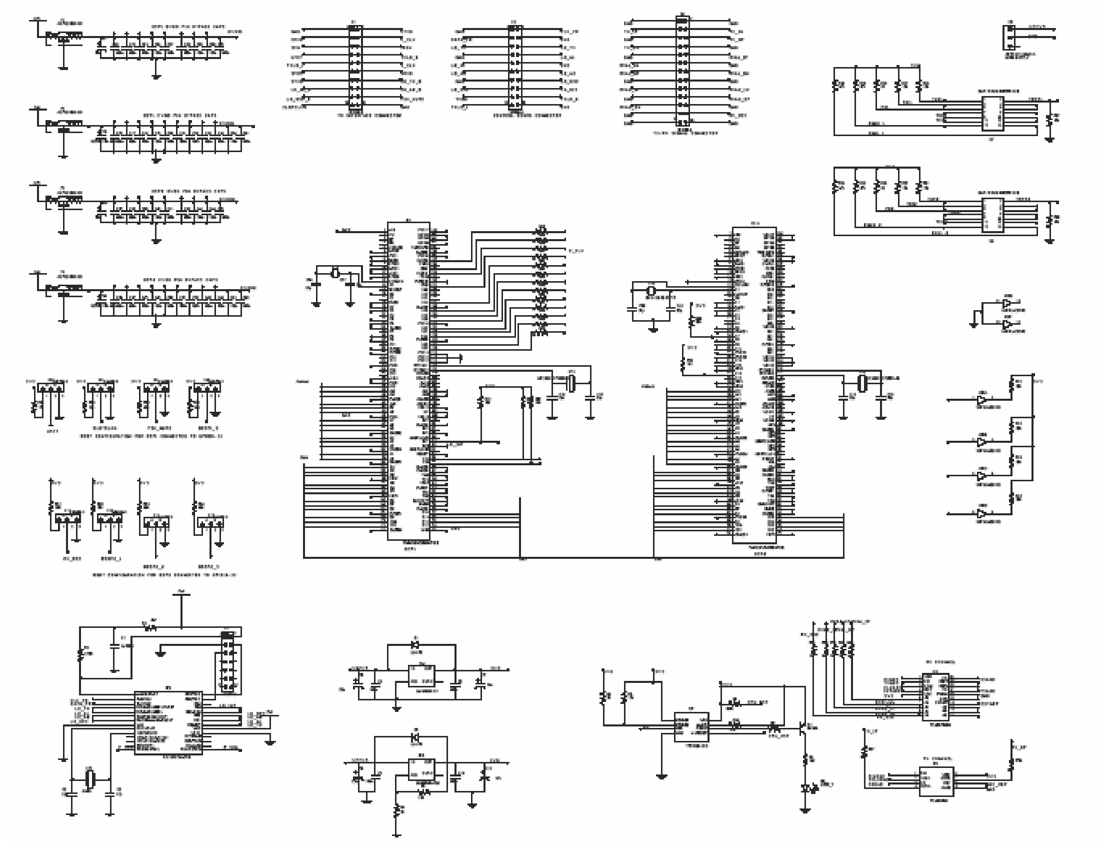
[10] **Marvin K. Simon, Charles C. Wang**, 1984. Differential Detection of Gaussian MSK in a Mobile Radio Environment

APPENDIX A: MATLAB CODE

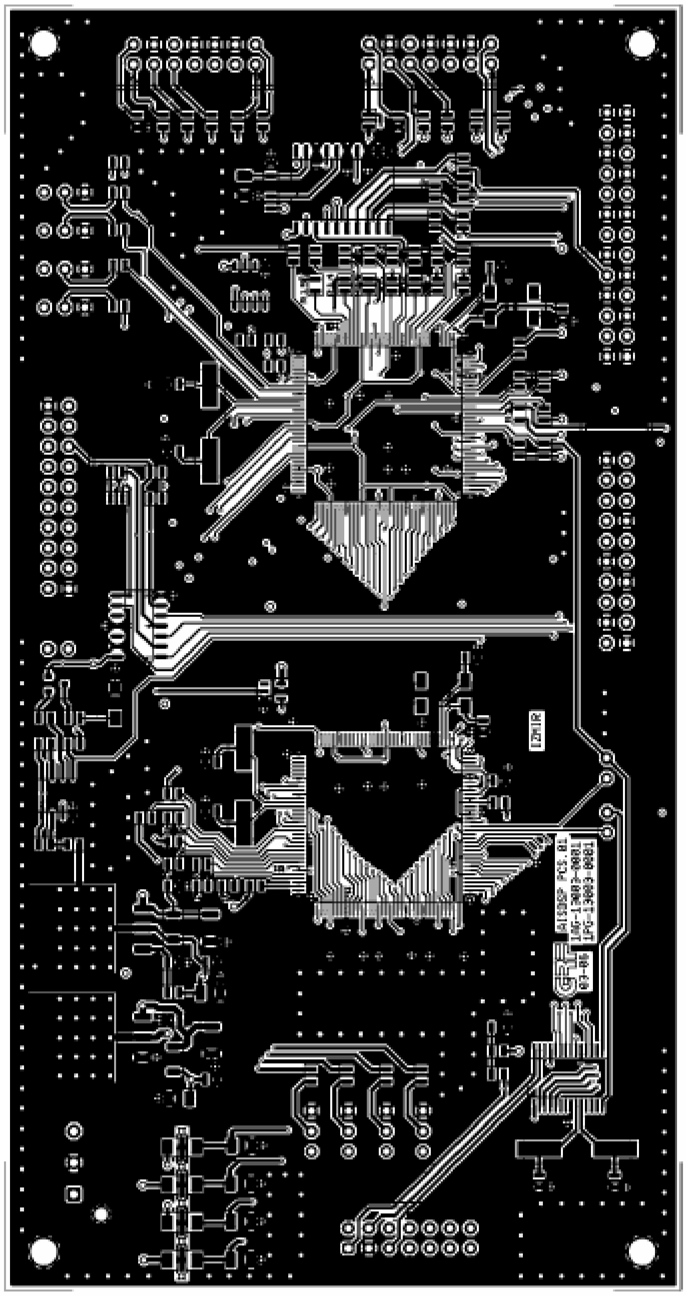
```
%****************************************************************%
%                    Implementation of the GMSK Modem            %
%****************************************************************%
clear all;
BR = 9600;
TB = 1/BR;
BT = 0.5;
N = 10;
SNR = 0:20;
BA = 256;
err_limit = 100;
packet_limit=1000;
err = zeros(1, length(SNR));
packet_error = err;
packet_counter =err;

for sn=1:length(SNR)
    packet_no=0;
    while (packet_error(sn)<err_limit) & (packet_no<packet_limit)

     packet_no=packet_no+1;
     En =10^(SNR(sn)/10);
     X = 2*round(rand(1,BA))-1;
     TXSEQ = X;
     X = kron(TXSEQ,ones(1,N));

 %**********************Gaussian    Filter ***************************
     t = -TB:TB/N:TB;
     sig = sqrt(log(2)) / (2*pi*BT);
     g = (1 / (sqrt(2*pi)*sig*TB)) * exp(-1*t.^2 / (2*sig^2*TB^2));
     g = g / (max(g));

%***********************Filtering Process *************************
     filtered = conv(X,g);
     %figure
     %subplot(211),stairs(X);
     %subplot(212),plot(filtered),grid;
     %filtered(length(filtered))

 %****************Extraction of the Phase Function*********************
```

```
        intfilt(1) = filtered(1);
        for k = 2:length(filtered)
           intfilt(k) = intfilt(k-1) + filtered(k);
        end
        intfilt = (pi/(2*N*N)) * intfilt;

%************** Computing I and Q components ************************
        I   = cos(intfilt);
       Q  = sin(intfilt);

%*****************Generating GMSK Signal**************************
       fc = 8000;
       T = 1 / (10*fc);
       z  = 0:T:T*(length(I)-1);
       Z = cos(2*pi*fc*z).*I - sin(2*pi*fc*z).*Q;

%*********************Decomposing I and Q**************************

       I = cos(2*pi*fc*t).*Z;
      Q = -(sin(2*pi*fc*t)).*Z;

%*********************Demodulation*******************************
        demod(1:length(I)) = Q .* [zeros(1,N), I(1:length(I)-N)] - I .* [zeros(1,N),
Q(1:length(Q)-N)];
        decision1=sign(demod);
        decision2=decision1(2*N:N:end-N);

%******************BER Calculation*******************************
          bit_error=length(find(TXSEQ(1:length(TXSEQ)-1)
        ~=  decision2(1:length(TXSEQ)-1)));

        if bit_error~=0
           packet_error(sn)=packet_error(sn)+1;
        end

        err(sn) =err(sn)+bit_error;

    end
    packet_counter(sn)=packet_no;
    [err' packet_counter' packet_error']
    save arif.mat err packet_counter;
end
ber=err./((BA-1)*packet_counter);
fer=packet_error./packet_counter;
figure;
semilogy(SNR, ber, '-x'); hold;grid;
semilogy(SNR, fer, '-r');

%**************************************************************
```

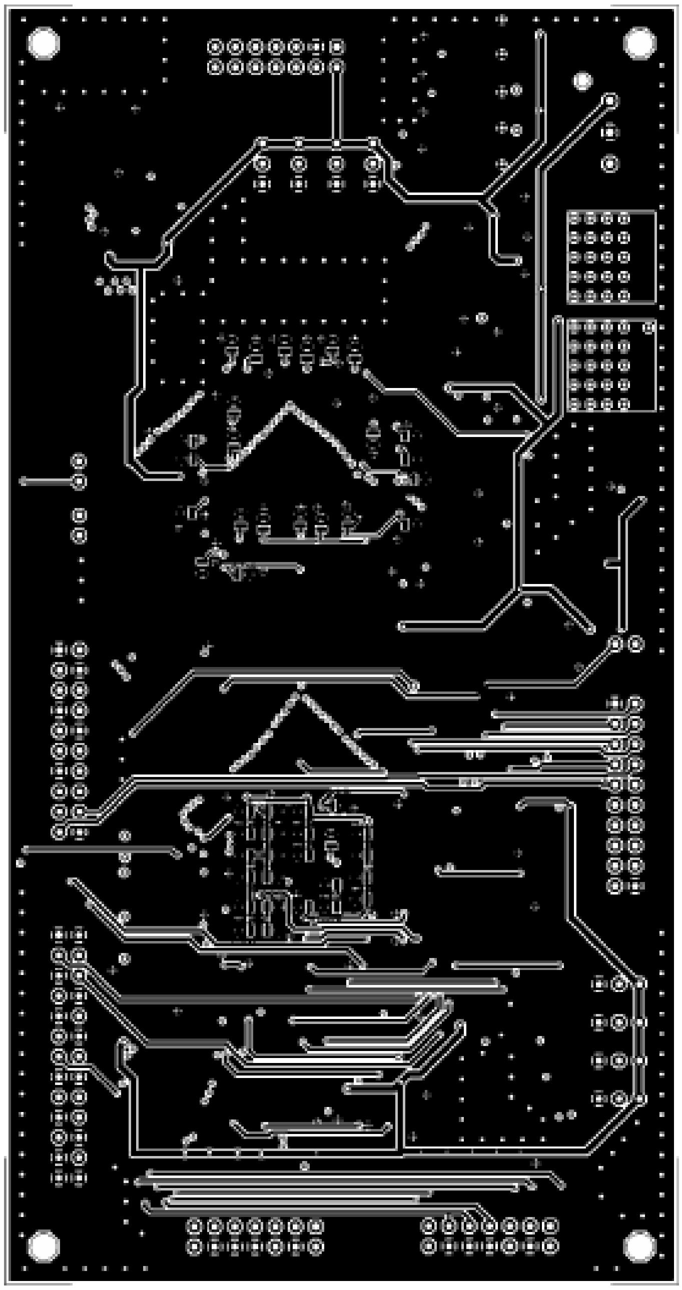
APPENDIX B: MODEM SCHEMATIC



APPENDIX C: PCB LAYOUT TOP VIEW



APPENDIX D: PCB LAYOUT BOTTOM VIEW



AUTOBIOGRAPHY

Arif Kürşad KAVAS was born in Afyon in 1981. He graduated from Bornova Anadolu Lisesi in 1999. He continued his education at Yeditepe University Electrical and Electronics Engineering with full scholarship and graduated ranking 2^{nd}. He started education for master degree at the Electronics Engineering Department of İstanbul Technical University. He is currently working as an R&D engineer in a telecom company. He is interested in RF design, wireless digital communications, embedded wireless control networks.