DESIGN OF ZIGBEE TRANSCEIVER FOR IEEE 802.15.4 USING MATLAB/SIMULINK

A THESIS SUBMITTED IN PARTIAL FULFILMENT OF THE REQUIRMENTS FOR THE DEGREE OF

MASTER OF TECHNOLOGY

IN

TELEMATICS AND SIGNAL PROCESSING

By

RAVIKANTH KANNA

Roll No: 209EC1106

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

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UNDER THE GUIDANCE OF

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CERTIFICATE

This is to certify that the thesis entitled, **"DESIGN OF ZIGBEE TRANSCEIVER FOR IEEE 802.15.4 USING MATLAB/SIMULINK"** submitted by **RAVIKANTH KANNA** in partial fulfilment of the requirements for the award of Master of Technology Degree in **Electronics & Communication Engineering** with specialization in **Telematics and Signal Processing** during 2010-2011 at the National Institute of Technology, Rourkela (Deemed University) is an authentic work carried out by him under my supervision and guidance. To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other University / Institute for the award of any Degree or Diploma.

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RAVIKANTH KANNA

Abstract

ZigBee technology was developed for a wireless personal area networks (PAN), aimed at control and military applications with low data rate and low power consumption. This thesis is mainly focusing on development of Matlab/Simulink model for ZigBee transceiver at physical layer using IEEE 802.15.4. ZigBee is a low-cost, low-power, wireless mesh networking standard. First, the low cost allows the technology to be widely deployed in wireless control and monitoring applications. Second, the low power-usage allows longer life with smaller batteries. Third, the mesh networking provides high reliability and more extensive range. The work presented here is to show how we can implement ZigBee transceiver with its specifications by using Matlab/simulink, without using complex mathematical blocks.

A ZigBee chip can tested and prepared by shifting the whole work from matlab environment to cadance environment. This can be done by HDL languages like Verilog HDL. Here, Minimum Shift Keying (MSK) modulation technique is described, an analysis of which shows that the theoretical maximum bandwidth efficiency of MSK is 2 bits/s/Hz which is same as for Quadrature Phase Shift Keying (QPSK) and Offset Quadrature Phase Shift Keying (Offset QPSK). The implementation clearly confirms the viability of theoritical approach. Results show that OQPSK modulation with half sine pulse shaping is perfectly employed ZigBee technology.

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List of Acronyms

ASK	Amplitude Shift Keying
AWGN	Additive White Guassian Noise
BPSK	Binary Phase Shift Keying
CPFSK	Continuous Phase Frequency Shift Keying
DSSS	Direct Sequence Spread spectrum
FHSS	Frequency Hop Spread Spectrum
FSK	Frequency Shift Keying
HVAC	Heating, Ventilation and Air Conditioning
IF	Intermediate Frequency
LNA	Low Noise Amplifier
MAC	Medium Access Layer
MSK	Minimum Shift Keying
OQPSK	Offset Quadrature Phase Shift Keying
OSI	Open System Interconnect
PHY	Physical Layer
PN	Pseudorandom Noise
PSD	Power Spectral Density
PSK	Phase Shift Keying
QPSK	Quadrature Phase Shift Keying
WLAN	Wireless Local Area Network
WPAN	Wireless Personal Area Network
TRF	Tuned Radio Frequency

CHAPTER 1

INTRODUCTION

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INTRODUCTION

1.1 INTRODUCTION:

Wireless personal area network (WPAN) and wireless local area network (WLAN) technologies are growing fast with the new emerging standards being developed. For sometime, Bluetooth was most widely used for short range communications. Now, ZigBee is becoming as an alternative to Bluetooth for devices with low power consumption and for low data rate applications.

The Bluetooth standard is a specification for WPAN. Although products based on the Bluetooth standards are often capable of operating at greater distances, the targeted operating area is the one around the individual i.e., within a 10m diameter. Bluetooth utilizes a short range radio link that operates in the 2.4GHz industrial scientific and medical(ISM) band similar to WLAN. However, the radio link in Bluetooth is based on the frequency hop spread spectrum. We know that Bluetooth occupies only 1MHz, the signal changes the centre frequency or hops at the rate of 1600Hz. Bluetooth hops over 79 centre frequencies, so over time the Bluetooth signal actually occupies 79MHz ([1], [2]).

ZigBee standard is developed by ZigBee Alliance, which has hundreds of member companies, from the semi-conductor industry and software developers to original equipment manufacturers and installers. The ZigBee alliance was formed in 2002 as a nonprofit organization open to everyone who wanted to join. The ZigBee standard has adopted IEEE 802.15.4 as its Physical Layer (PHY) and Medium Access Control (MAC) protocols. Therefore, a ZigBee compliant device is compliant with the IEEE 802.15.4 standard as well.

ZigBee is a low-cost, low-power, wireless mesh networking standard. First, the low cost allows the technology to be widely deployed in wireless control and monitoring applications. Second, the low power-usage allows longer life with smaller batteries. Third, the mesh networking provides high reliability and more extensive range. ZigBee is a standard that defines a set of communication protocols for low data rate short range wireless networking. ZigBee based wireless devices operate in 868MHz, 915MHz and 2.4GHz frequency bands. ZigBee is targeted mainly for battery power applications where low data rate, low cost and long battery life are main requirements. In many ZigBee applications, the total time the wireless device is engaged in any type of activity is very limited. The device spends most of its time in power saving mode, also known as sleep mode. As a result, ZigBee enabled devices are capable of being operational for several years before their batteries needs to be replaced [3].

ZigBee standard is specifically developed to address the need for very low cost implementation of low data rate wireless networks with ultra low power consumption. The ZigBee Standard reduced the implementation cost by simplifying the communication protocols and reducing the data rate. The minimum requirements to meet ZigBee and IEEE 802.15.4 specifications are relatively relaxed compared to other standards such as IEEE 802.11, which reduces the complexity and cost of implementing ZigBee compliant transceivers.

1.2 Objective:

The main focus of the work is to design a ZigBee Transceiver using Matlab/Simulink that uses OQPSK modulation with half sine pulse shaping. This work will helpful for designing a ZigBee chip.

1.3 Thesis Layout:

The first three chapters cover the theoretical and mathematical background of all modulation schemes. Fourth and fifth chapters covers the basics of commonly used radio transceiver architectures and the major functions to be done while implementing ZigBee transceiver. Sixth chapter contains all simulation results.

Chapter2 provides a brief description on IEEE 802.15.4 standard including operating frequencies, specifications and data rates. The relationship between IEEE 802.15.4 standard and ZigBee standard and comparison between ZigBee Vs Bluetooth and IEEE 802.11 also described.

Chapter3 provides a brief discussion on some of digital modulation and demodulation schemes are given, especially coherent demodulation. The modulation techniques like BPSK, QPSK, OQPSK and MSK schemes are discussed. For each of the above modulation schemes, constellation plot, modulator, coherent demodulator, power spectral density plots are discussed detailly.

In chapter4, commonly used radio transceiver architectures like superhetrodyne receiver, low IF receiver and direct conversion receiver are described. Advantages and disadvantages of each architecture were also explained. While designing, each and every function in the transceiver was explained briefly.

Chapter5 provides block diagram of ZigBee transmitter and receiver. A Step by step procedure to implement ZigBee transmitter and receiver using Simulink like bit to symbol mapping, symbol to chip mapping, serial to parallel conversion, halfsine pulse shaping and modulation are explained. While coming to the receiver part, demodulation, parallel to serial conversion and despreading are explained.

Chapter6 provides simulation results at each and every step of ZigBee transceiver, which are supporting the theory provided in the earlier chapters.

Finally the work is concluded in chapter7 and scope for the future work is explained.

CHAPTER 2

ABOUT ZIGBEE

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ABOUT ZIGBEE

2.1 About ZigBee:

ZigBee standard consists of a whole suite of specifications designed specifically for wireless networked sensors and controllers. The physical(PHY) and medium access control (MAC) layers are standardized by the IEEE 802.15 wireless personal area network (WPAN) working group under the designation of 802.15.4 [4]. The standard mainly aims at low cost, low data rate and low power wireless network. The higher layers are specified by the ZigBee Alliance [5], which is an industry alliance consisting of a full spectrum of companies, ranging from ZigBee chip providers to solution providers. Compared to other wireless communication technologies, ZigBee is designed specifically for providing wireless networking capability for battery-powered, low-cost, low capability sensor and controller nodes, typically powered only by an eight-bit microcontroller. The ZigBee technology is designed to provide a simple and low-cost wireless communication and networking solution for low-data rate and low power consumption applications, such as home monitoring and automation, environmental monitoring, industry controls, and emerging low-rate wireless sensor applications

2.1.1 HISTORY OF ZIGBEE

- ZigBee-style networks were conceived around 1998, when many application engineers realized that both Wi-Fi and Bluetooth were going to be unsuitable for many applications. In particular, many engineers saw a need for self-organizing ad-hoc digital radio networks.

- The IEEE 802.15.4-2003 standard was established in May 2003. This was superseded by the publication of IEEE 802.15.4-2006.

- The ZigBee Alliance membership had more than doubled in 2004, growing to more than 100 member companies, in 22 countries. By April 2005 membership had grown to more than 150 companies, and by December 2005 membership had passed 200 companies.

- The ZigBee specifications were ratified on 14 December 2004.
- The ZigBee Alliance announced public availability of Specification 1.0 on 13 June 2005, known as ZigBee 2004 Specification.
- The ZigBee Alliance announces the completion and immediate member availability of the enhanced version of the ZigBee Standard in September 2006, known as ZigBee 2006 Specification.
- During the last quarter of 2007, ZigBee PRO, the enhanced ZigBee specification was finalized.

The name of the brand was “ZigBee”, originated with reference to the behaviour of honey bees after their return to the bee hieve.

2.1.2 Applications:

The major applications of ZigBee are focused on sensor network and automatic control, such as personal medical assistance, industrial control, home automation, remote control and monitoring [6]. It is particularly suitable for biotelemetry applications because of low power consumption, e.g., the personal medical monitoring device for senior citizens. Rather than the traditional wired monitoring equipment, the biotelemetry techniques, allow electrical isolation from data processing devices and power lines.

One of the intended application of ZigBee is in-home patient monitoring. A patient’s vital body parameters, for example blood pressure and heart rate can be measured by wearable devices. The patient wears a ZigBee device that interfaces with a sensor that gathers health related information such as blood pressure on a periodic basis. Then the data is wirelessly transmitted to a local server, such as a personal computer inside the patient’s home, where initial analysis is performed. Finally the vital information is sent to the patient’s nurse or physician via the internet for further analysis.

Another example of a ZigBee application is monitoring the structural health of large scale building and structures. In this application, several ZigBee enabled wireless sensors like accelerometers can be installed in a building and all these sensors can form a single wireless network to gather the information that will be used to evaluate the building structural health and detects the signs of possible damage. After an earthquake, for example, a building could require before it reopens to the public. The data gathered by the sensors could help further and reduce the cost of inspection.

Home automation is one of the major application areas for ZigBee wireless networking. The typical data rate in home automation is only 10Kbps. Some of the possible ZigBee applications in a typical residential building are light control systems, security systems, meter reading systems, irrigation systems, multizone Heating, Ventilation, and Air Conditioning (HVAC) systems

2.2 Relationship between ZigBee and IEEE 802.15.4 standard

ZigBee wireless networking protocols are shown in Figure 2.1. ZigBee protocol layers are based on the Open System Interconnect (OSI) basic reference model. As shown in Figure 2.1Figure 2.1, the bottom two networking layers are defined by IEEE 802.15.4 standard. This standard is developed by IEEE 802 standards committee and was initially released in 2003. IEEE 802.15.4 defines the specifications for PHY and MAC layers of wireless networking, but it does not specify any requirements for higher networking layers. The ZigBee standard defines only the networking, applications and security layers of the protocol and adopts IEEE 802.15.4 PHY and MAC layers as a part of the ZigBee networking protocol. Therefore, ZigBee-compliant device conforms to IEEE 802.15.4 as well.

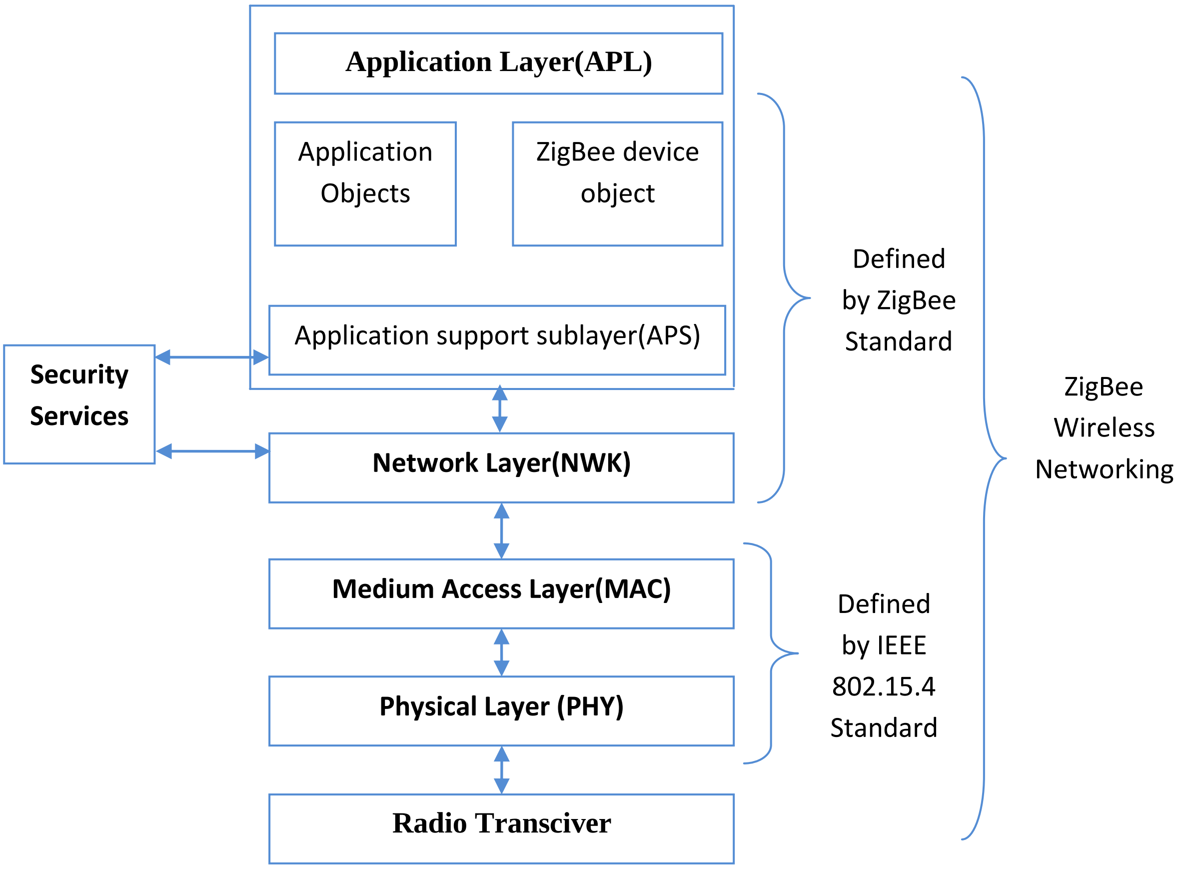

Figure 2.1: ZigBee wireless networking protocol Layers.

2.3 OPERATING FREQUENCIES and DATA RATES:

There are three frequency bands used for IEEE 802.15.4 [7]. These are

- 868 – 868.6 MHz (868MHz band)
- 902 – 928 MHz (915MHz)
- 2400 – 2483.5MHz (2.4GHz band)

The 868MHz band is used in Europe for a number of applications, including short range wireless networking.

The 915MHz and 2.4GHz bands are part of industrial, scientific and medical (ISM) frequency bands. The 915MHz frequency band is used mainly in North America, where as the 2.4GHz band is used worldwide.

The Table-1 given below provides additional details regarding the ways in which these three frequency bands are used in IEEE 802.15.4 standard [3]. IEEE 802.15.4 requires that if a transceiver supports 868MHz band, it must support 915MHz band as well, and vice versa. Therefore, the two bands are bundled together as the 868/915MHz frequency bands of operations. IEEE 802.15.4 has one mandatory and two optional specifications for the 868/915MHz bands. The mandatory requirements are simpler to implement but yield low data rates (20Kbps and 40Kbps, respectively). If a user chooses to implement the optional modes of operation, IEEE 802.15.4 still requires that it accommodate the low data rate mandatory operation in the 868/915MHz bands as well. Also, the transceiver must be able to switch dynamically between mandatory and optional modes of operation in 868/915MHz bands.

A 2.4GHz transceiver may support 868/915MHz bands, but it is not required by IEEE 802.15.4. There is room for only a single channel in the 868MHz band. The 915MHz band has 10 channels which are excluding the optional channels. The total number of channels in the 2.4GHz band is 16.

The 2.4GHz ISM band is accepted worldwide and has the maximum data rate and number of channels. For these reasons, developing transceivers for the 2.4GHz band is a popular choice for many manufacturers. However, IEEE 802.11b operates in the same 2.4GHz band and coexistence can be an issue in some applications. Also, the lower the frequency band is, the better since signal penetrates walls and various objects. Therefore, some users may find the 868/915MHz band a better choice for their applications.

There are three modulation types used in IEEE 802.15.4: binary phase shift keying (BPSK), amplitude shift keying (ASK), and Offset Quadrature Phase Shift Keying (O-QPSK). In BPSK and O-QPSK, the digital data is in the phase of the signal. In ASK, in contrast, the digital data is in the amplitude of the signal.

All wireless communication methods in IEEE 802.15.4 take advantage of either direct sequence spread spectrum (DSSS) or parallel sequence spread spectrum (PSSS) techniques. DSSS and PSSS helps to improve the performance of receivers in a multipath environment.

Table-1: IEEE 802.15.4 Data Rates and Frequencies of operation

Frequency (MHz)	Number of Channels	Modulation	Chip Rate (Kchip/s)	Bit Rate (Kb/s)	Symbol rate (Ksymbol/s)	Spreading Method
868 – 868.6	1	BPSK	300	20	20	Binary DSSS
902 – 928	10	BPSK	600	40	40	Binary DSSS
868 – 868.6	1	ASK	400	250	12.5	20 Bit PSSS
902- 928	10	ASK	1600	250	50	5 Bit PSSS
868 – 868.6	1	OQPSK	400	100	25	16- array Orthogonal
902-928	10	OQPSK	1000	250	62.5	16 – array Orthogonal
2400 – 2483.5	16	OQPSK	2000	250	62.5	16 array orthogonal

2.4 Interoperability

ZigBee has a wide range of applications; therefore, several manufacturers provide ZigBee -enabled solutions. It is important for these ZigBee based devices to be able to interact with each other regardless of the manufacturing origin. In other words, the devices should be interoperable, which is one of the key advantages of the ZigBee protocol stack. ZigBee-based devices are interoperable even when the messages are encrypted for security reasons.

2.5 Modulation and Spreading Methods for 2.4GHz Operation

We know that the graph of signal power versus frequency is referred to as the signal power spectral density (PSD). The Figure 2.2 represents a typical IEEE 802.15.4 signal PSD centered at 2450MHz [3]. This centre frequency is known as a carrier frequency. The signal bandwidth is normally considered the frequency band that contains the majority of the signal power. There are different definitions of signal bandwidth in the literature.

In the Figure 2.2, the power spectral density peak is at middle of the graph. The 3 dB corners of the PSD graph are the frequencies at which the PSD value drops 3 dB below its maximum. The 3 dB bandwidth is defined as the frequency interval between the 3 dB corners. Another way to define the signal bandwidth is to consider the nulls on the PSD graph as the signal borders. This is known as null-to-null bandwidth of the signal. The frequency band that contains 99% of the signal power is referred to as the bandwidth of the signal.

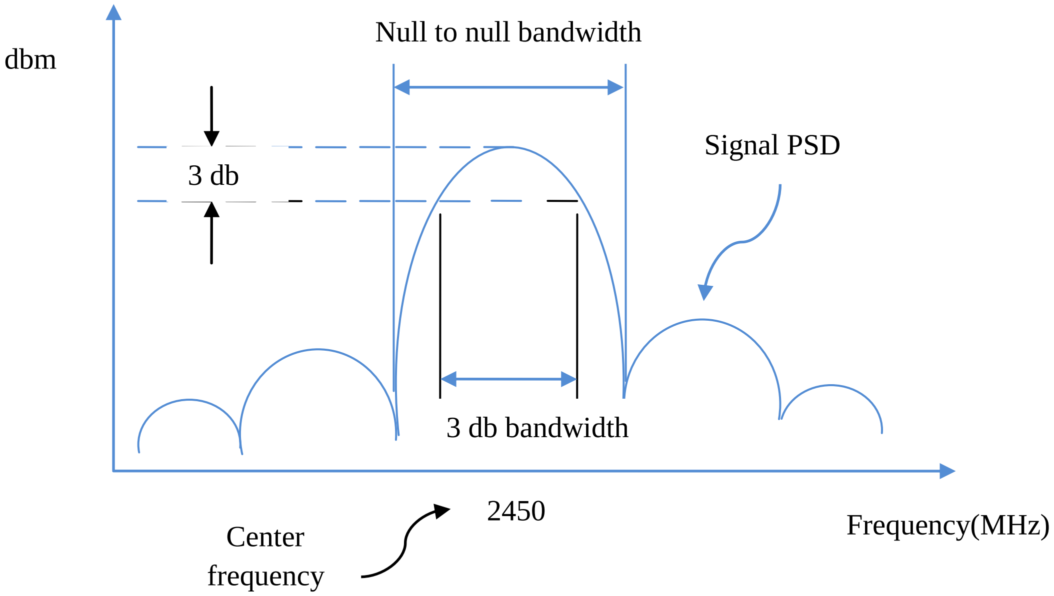

Figure 2.2: Signal Bandwidth definitions

When the destination receives the signal, the receiver circuit will use a combination of analog and digital filtering to remove the spectral contents outside the frequency band of interest. The noise is normally modelled as a signal with flat power density. When the signal is filtered, the only noise that matters is the noise within the frequency band of interest. The ratio of the total signal power to total noise power within the band of interest is called the signal to noise ratio (SNR). The SNR is an indication of the signal quality and increasing SNR will improve receiver PER if the receiver is not suffering from multipath issues.

Regardless of the definition of the signal bandwidth, the bit rate associated with a signal is proportional to the signal bandwidth in the frequency domain. That means for a given modulation/spreading method and signal SNR, to double the bit rate we need it double the signal bandwidth. The spectral efficiency is defined as bits/second/Hertz and determines the relationship between the signal bit rate and the associated bandwidth for a given modulation and spreading technique. For example, if a method provides 2 bits/second/Hertz spectral efficiency, it means that a signal of 1 MHz bandwidth can deliver a bit rate of up to 2Mbps.

IEEE 802.15.4 uses spreading methods to improve the receiver sensitivity level, increase the jamming resistance, and reduce the effect of multipath. The spreading method required by IEEE 802.15.4 for the 2.4GHz frequency band is the Direct Sequence Spread Spectrum (DSSS). In an IEEE 802.15.4-specific implementation of DSSS, at every 4 bits of each octet of a PHY protocol data unit (PPDU) are grouped together and referred to as symbol. Then a lookup table is used to map each symbol to a unique 32-bit sequence. This 32-bit sequence is also known as the chip sequence or the pseudorandom noise (PN) sequence. Here, every four bits are mapped to a unique chip sequence; the lookup table contains 16-chip sequences. Each such a chip sequence appears as a random sequence of zeroes and ones. But each sequence is selected by a procedure to minimize its similarity to the other 15 sequences. The similarity of the two sequences measured by calculating the cross-correlation function of two sequences. The cross-correlation is determined by multiplying the sequences together and then calculating the summation of result. A sequence containing 0 and 1 is replaced by its bipolar before calculation of cross-correlation. If x(n) and y(n) are two sequences, the cross-correlation of these two sequences is given below:

$$r_{xy}(0) = \sum_{n=-\infty}^{n=+\infty} x(n)y(n) \qquad (2.1)$$

The $r_{xy}(0)$ is the calculated cross-correlation of x(n) and y(n) when neither of the sequences is shifted. The higher the absolute value of $r_{xy}(0)$, the higher the similarity of two sequences. If the cross-correlation $r_{xy}(0)$ is equal to zero, it indicates that the sequences x(n) and y(n) are as dissimilar as possible. In this case, the sequences x(n) and y(n) are known as orthogonal sequences. The 16 sequences used in IEEE 802.15.4 are not completely orthogonal and are referred to as near-orthogonal or quasi-orthogonal sequences.

The cross correlation can be calculated for x(n) and y(n-k) as well, where y(n-k) is the sequence y(n) shifted by k:

$$r_{xy}(k) = \sum_{n=-\infty}^{n=+\infty} x(n)y(n-k) \qquad (2.2)$$

In this case, $r_{xy}(k)$ is an indication of the similarity of the x(n) sequence and y(n-k) sequence. Since each four bits of the actual data are mapped to a 32-bit chip sequence, the effective over-the-air bit rate is increased by a factor of eight. If the original signal before spreading has a bandwidth of 250KHz, after spreading the bandwidth will be increased to 2MHz. Since the signal spreading does not add any physical energy to the signal, the peak of the PSD of the signal will be less after the spreading because the same signal energy is distributed over a larger bandwidth.

When the signal with 2MHz bandwidth is travelling over the air, undesired noise and interferers can be added to it. The receiver of the signal will use the dispreading procedure to recover the original signal. In dispreading, every 32 bits of the received signal are compared to 16 possible chip sequences, and the chip sequence that has the most similarity to the received signal will be chosen as the received chip sequence. From the same lookup table used by the transmitter, the receiver can recover the original four bits packet.

As a result of despreading, the signal energy will be concentrated back to the original bandwidth of 250 KHz, but the despreading will not affect noise level in the 250KHz band of interest. Since the signal energy is increased in the desired band without increasing the noise level, the effective SNR is increased by signal despreading. This increase in signal SNR will directly improve the receiver sensitivity. This improvement in SNR is referred to as the processing gain. The value of the processing gain is equal to the ratio of the signal bit rate after spreading to the signal bit rate before spreading. For example, the processing gain for the 2.4GHz mode of operation in IEEE 802.15.4 is equal to 9dB:

$$Processing\ gain = 10\ X\ log_{10}\left(\frac{2Mbps}{250Kbps}\right) \sim 9dB.$$

2.6 Specifications of ZigBee operating on 2.4 GHz:

The specifications for the ZigBee technology operating in the 2.4GHz frequency band are given in Table-2([3], [5], [8]).

Table-2: Specifications of ZigBee Technology

Parameter	Specification
Data rate	250 kbps
Number of channels	16
Operating frequency	2.4GHz
Channel spacing	5 MHz
Spread spectrum	Direct Sequence Spread Spectrum
Chip rate	2 Mega chips per second.
Modulation	OQPSK with Half sine pulse shaping

2.7 ZIGBEE Vs BLUETOOTH and IEEE 802.11

The comparison between ZigBee with Bluetooth and IEEE 802.11 WLAN helps for understanding how ZigBee differentiates itself from existing established standards. Table-3 shows the basic characteristics of these three standards [3].

Table-3: Characteristics of ZigBee, Bluetooth and IEEE 802.11b

Technology	Data rate	Typical Range	Application
ZigBee	20 – 250 Kbps.	10 – 100 m	Wireless sensor networks
Bluetooth	1 – 3 Mbps.	2 – 20 m	Wireless mouse
IEEE 802.11b	1 – 11 Mbps.	30 – 100 m	Wireless Internet Connection

IEEE 802.11 is a family of standards; IEEE 802.11b is selected here because it operates in 2.4 GHz band, which is common with Bluetooth and ZigBee. IEEE 802.11b has a high data rate up to 11 Mbps, and providing a wireless Internet connection is one of its typical applications. The indoor range of IEEE 802.11b is typically between 30 and 100 meters. Bluetooth, on the other hand, has a lower data rate (less than 3 Mbps) and its indoor range is typically 2–10 meters. One popular application of Bluetooth is in wireless headsets, where Bluetooth provides the means for communication between a mobile phone and a

hands-free headset. ZigBee has the lowest data rate and complexity among these three standards and provides significantly longer battery life.

ZigBee's very low data rate means that it is not the best choice for implementing a wireless Internet connection or a CD-quality wireless headset where more than 1Mbps is desired. However, if the goal of wireless communication is to transmit and receive simple commands and/or gather information from sensors such as temperature or humidity sensors, ZigBee provides the most power and the most cost-efficient solution compared to Bluetooth and IEEE 802.11b.

CHAPTER 3

REVIEW OF DIGITAL MODULATION SCHEMES

CHAPTER 3

REVIEW OF DIGITALMODULATION SCHEMES

3.1 INTRODUCTION:

Here we are analyzing various modulation and demodulation schemes used in digital communication system. For each modulation technique, we analyze the mathematical representation of the waveforms, the signal space, the generation and detection, the spectral content and the performance over the AWGN channel. We can define pass-band signaling as a form of transmission that uses modulation, i.e. the information signal to be transmitted is modulated onto a carrier. Moreover, a pass-band waveform has its frequency spectrum concentrated around the carrier frequency f_c. When the modulating signal is digital, we call the process of generating the pass-band signal as digital modulation. In theory, any baseband digital signal can be used to modulate a carrier, directly or indirectly, and generate a pass-band signal. However, as in the case of baseband waveforms, there are specific digital modulations that are more suitable to a given channel or more adequate to satisfy design constraints. Pass band transmission is adequate to communication channels whose frequency response typically have a band-pass response effect, like wireless channels. Some baseband channels can also carry pass band waveforms, usually having moderate-to low frequency carriers. One typical example is the twisted pair cable when used to carry signals from voice-band modems [7].

Similarly to the case of analog modulation, in a digital modulation scheme the digital modulating signal (data bits) alters the amplitude, phase or frequency of a sinusoidal carrier. A combination of these parameters can be affected by the modulating signal to form variants of basic digital modulations. There are mainly three types of modulation schemes.

1. Amplitude Shift Keying (ASK).
2. Phase Shift Keying (PSK).
3. Frequency Shift Keying (FSK).

In the binary ASK modulation, the carrier amplitude is switched between two levels according to the modulating data bits. In the binary PSK modulation, the carrier phase is switched between two values according to the data bits. In the binary FSK modulation, the carrier frequency is switched between two values according to the modulating data bits. These modulations are the digital counterparts are the analog AM (amplitude modulation), PM (phase modulation) and FM (frequency modulation), respectively. Figure 3.1 shows the waveforms of ASK, PSK and FSK schemes.

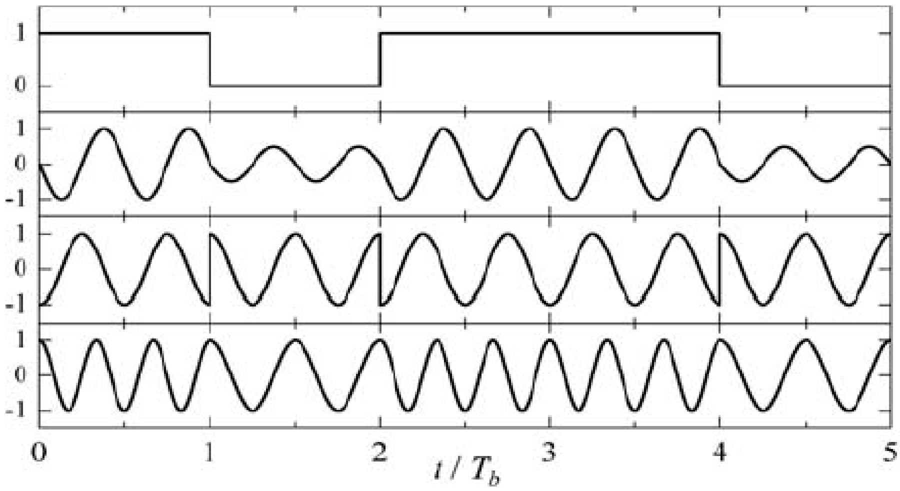

Figure 3.1 : Basic digital modulation schemes

3.1.1 Coherent and Non-coherent Detection

Due to propagation delay and fading in the communication channel, the transmitted signal can undergo phase changes when traveling through it. This means that if we use a base function with a given initial phase at the transmitter, the receiver must be fed by a base-function with a phase that reflects any phase change caused by the channel. In other words, carrier synchronization between the transmitter and the receiver must be performed. Any modulation scheme that makes use of this synchronization is said to be coherently detected. Non-coherent detected modulations do not need carrier synchronization.

Coherent detection is performed at a cost, that is, the receiver must be equipped with a carrier recovery circuitry, which increases system complexity, and can increase size and power consumption. Additionally, there is no ideal carrier recovery circuit. Then, strictly speaking, no practical digital communication system works under perfect phase coherence. Non-coherent detection is simpler, but it suffers from performance degradation as compared to coherent detection, but this difference can be small in practice for some modulation schemes due to the specifics of the modulation and also due to the penalty caused by imperfections in the carrier recovering process. So there is no need of carrier recovery circuit in case of non-coherent detection [7].

3.1.2 Spectral efficiency and power efficiency

The spectral efficiency of a digital modulation is the ratio between the transmitted bit rate *Rb* and the occupied bandwidth B [9].

$$\rho = \frac{R}{B}\text{bits/s/Hz} \tag{3.1}$$

The result, expressed in bits per second per hertz (bit/s/Hz), is a measure of the data packing capability of the modulation. Note that to compute the spectral efficiency, first we have to define the occupied bandwidth *B*.

Power efficiency is a measure of the $\frac{E_b}{N_0}$ ratio needed for a target symbol or bit error probability. Unfortunately, for most of the systems, spectral efficiency and power efficiency are conflicting parameters, which means that increasing one will often cause a reduction in the other.

3.2 Various Digital modulation techniques:

3.2.1 Binary Phase Shift Keying

In the binary ASK (amplitude-shift keying) modulation, the carrier amplitude is switched between two levels according to the modulating data bits. The BPSK modulated symbol can be represented by the equation given below.

$$S_i(t) = \sqrt{\frac{2E_b}{T_b}}\cos(2\pi f_c t + (i-1)\pi), \quad \begin{cases} 0 \le t \le T_b \\ i = 1,2 \\ f_c = \dfrac{1}{T_b} \end{cases} \tag{3.2}$$

where E_b is the average bit energy, T_b is the bit duration and f_c is the carrier frequency. If it is desired to have phase transitions occurring in the same points of the carrier signal, f_c must be an integer multiple of $\frac{1}{T_b}$. Nevertheless, it is not mandatory to satisfy this condition.

In the above equation, if $S_1(t) = -S_2(t)$ then this BPSK modulation scheme is known as anti-podal signaling. Then, the base-function can be determined by normalizing $s_1(t)$ or $s_2(t)$ to unit energy. The base function for the BPSK modulation system is given by

$$\phi_1(t) = \sqrt{\frac{2}{T_b}}\cos(2\pi f_c t) \qquad (3.3)$$

3.2.2 BPSK CONSTELLATION:

The BPSK signal-vector coefficients are given by

$$S_{11} = \int_0^{T_b} S_1(t)\Phi_1(t)dt = \sqrt{E_b}$$

$$S_{21} = \int_0^{T_b} S_2(t)\Phi_1(t)dt = -\sqrt{E_b} \qquad (3.4)$$

Resulting in constellation shown in the Figure 3.2 below ([7], [9]) .

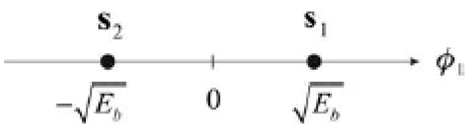

Figure 3.2 : BPSK constellation plot

3.2.3 BPSK MODULATOR & DEMODULATOR

It is easy to derive the structure of the BPSK modulator, which is shown in Figure 3.3 below. Data bits are level-converted using, for example, the mapping rule: bit "0"$\rightarrow -\sqrt{E_b}$ and bit "1"$\rightarrow +\sqrt{E_b}$. The resultant bipolar NRZ signal multiplies the base-function, generating the BPSK signal.

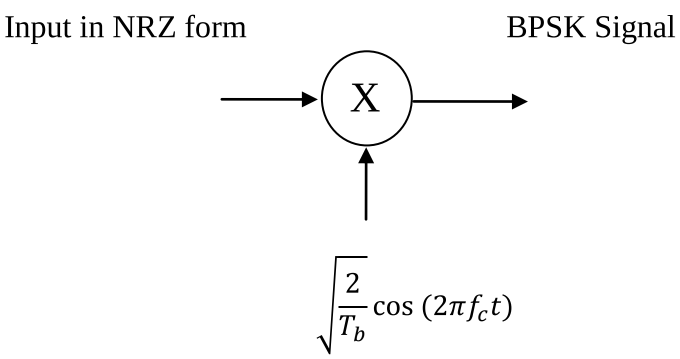

Figure 3.3: BPSK Modulator

The Figure 3.4 shown below explains about the BPSK receiver using coherent detection ([7], [10]). The receiver configuration we are using is correlator receiver. In this the received BPSK signal is multiplied with the base function which results baseband data and second order harmonics of carrier frequency. This higher harmonics are eliminated by using a low pass filter or an integrator and making it to pass only baseband signal. These signals is sampled using sample and hold circuit and pass it to the comparator. Comparator will compare the incoming data with threshold. If the incoming data value is greater than threshold comparator detects the bit as '1' otherwise it detects as '0'.

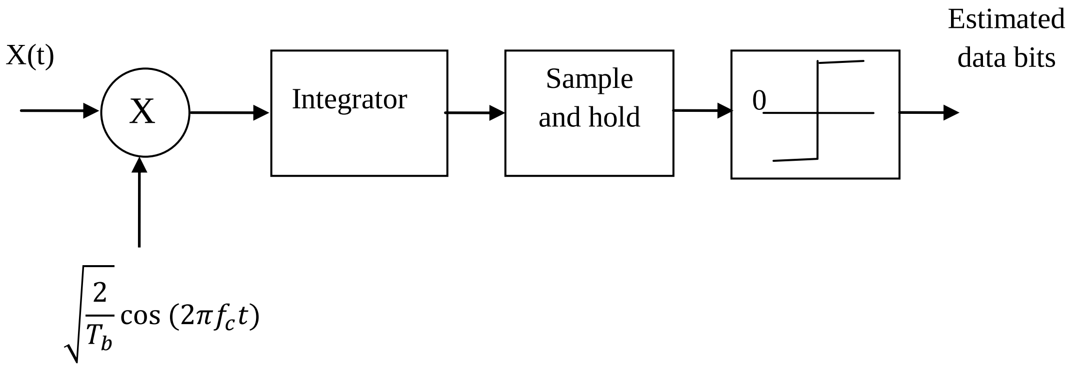

Figure 3.4: BPSK Demodulator using coherent detection

3.2.4 Power Spectral density of BPSK signal

We know that the modulated signal s(t) in terms of its in-phase and quadrature components S_I (t) and S_Q(t), as shown by

$$S(t) = S_I(t)\cos(2\pi f_c t) - S_Q(t)\sin(2\pi f_c t) \qquad (3.5)$$

Comparing (3.2) with (3.5) we conclude that $S_Q(t) = 0$ and that S_I (t) can be seen as a bipolar NRZ random signal composed by equally-likely rectangular pulses p(t) of duration T_b and amplitudes $\pm(\frac{2E_b}{T_b})^{\frac{1}{2}}$. Power spectral density of such waveform can be determined by dividing the squared magnitude of the Fourier transform of p(t) by the pulse duration [7].

$$S_B(f) = \frac{2E_b}{T_b}\frac{T_b^2\mathrm{sinc}^2(f\,T_b)}{T_b} = 2E_b\mathrm{sinc}^2(fT_b) \qquad (3.6)$$

The PSD of the BPSK signal is determined by applying

$$S(f) = \frac{1}{4}[S_B(f - f_c) + S_B(f + f_c)] \tag{3.7}$$

Which results in

$$S(f) = \frac{E_b}{2} \operatorname{sinc}^2[(f - f_c)T_b] + \frac{E_b}{2} \operatorname{sinc}^2[(f + f_c)T_b] \tag{3.8}$$

Figure 3.5 indicates the power spectral density of the BPSK signal.

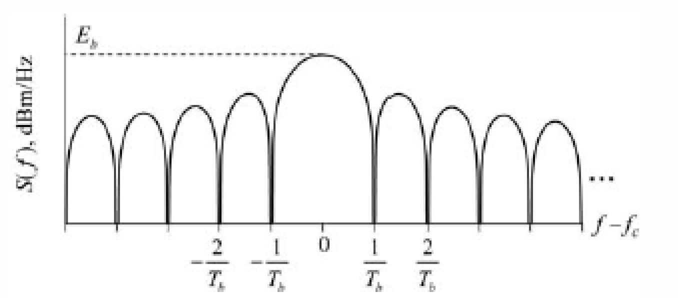

Figure 3.5: Power Spectral density of modulated signal.

We know that minimum theoretical occupied bandwidth B = B_{min}, the one that would be obtained with the use of an ideal brick-wall band-pass filter. For some baseband signals, B_{min} is half of the signalling rate per dimension. For some passband signals B_{min} becomes equal to the signalling rate per dimension, which is the case for the BPSK modulation. Then, its maximum spectral efficiency is given by

$$\rho_{max} = \frac{R_b}{B_{min}} = \frac{R_b}{(\frac{1}{T_b})} = 1 \text{ bit/s/Hz} \tag{3.9}$$

The probability of error for the BPSK signal is given by

$$P_e = Q\left(\sqrt{\frac{2E_b}{N_o}}\right) \tag{3.10}$$

As we increase the transmitted signal energy per bit, E_b , for a specified noise spectral density N_o, the message points corresponding to symbols 1 and 0 move further apart in the constellation shown above, and the corresponding probability of error is reduced according the equation given above.

3.3 Quadrature Phase Shift Keying

The main goal in the design of digital communication system is to achieve low probability of error and effective utilisation of channel bandwidth. Now we will study a bandwidth conserving modulation scheme known as Quadriphase-shift keying(QPSK), which is an example of Quadrature carrier multiplexing [10].

In QPSK, as with binary PSK, information carried by the transmitted signal is contained in phase. In particular, the phase of the carrier takes on one of four equally spaced values, such as $\frac{\pi}{4}$, $\frac{3\pi}{4}$, $\frac{5\pi}{4}$ and $\frac{7\pi}{4}$. For this set of values we may define the transmitted signal as

$$S_i(t) = \begin{cases} \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + (2i-1)\frac{\pi}{2}\right], & 0 \le t \le T_s \\ 0 & \text{elsewhere} \end{cases} \tag{3.11}$$

Where i=1,2,3,4; E_s is the transmitted signal energy per symbol, and T_s is the symbol duration. The carrier frequency f_c equals $\frac{n_c}{T_s}$ for some fixed integer n_c.

Using trigonometric identities, the above equation can be rewritten as

$$S_{QPSK}(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[(i-1)\frac{\pi}{2}\right] \cos(2\pi f_c t) - \sqrt{\frac{2E_s}{T_s}} \sin\left[(i-1)\frac{\pi}{2}\right] \sin(2\pi f_c t) \tag{3.12}$$

If basis functions $\phi_1(t = \sqrt{\frac{2}{T_s}} \cos(2\pi f_c t)$, $\phi_2 = \sqrt{\frac{2}{T_s}} \sin(2\pi f_c t)$ are defined over the interval $0 \le t \le T_s$ for the QPSK signal set, then the four signals in the set can be expressed in terms of the basis signals as

$$S_{QPSK} = \{\sqrt{E_s} \cos\left[(i-1)\frac{\pi}{2}\right] \phi_1(t) - \sqrt{E_s} \sin\left[(i-1)\frac{\pi}{2}\right] \phi_2(t)\} \quad \text{i=1,2,3,4.} \tag{3.13}$$

3.3.1 Constellation plot:

There are two possible constellations for QPSK([7], [9], [10]). They are shown in the Figure 3.6 below.

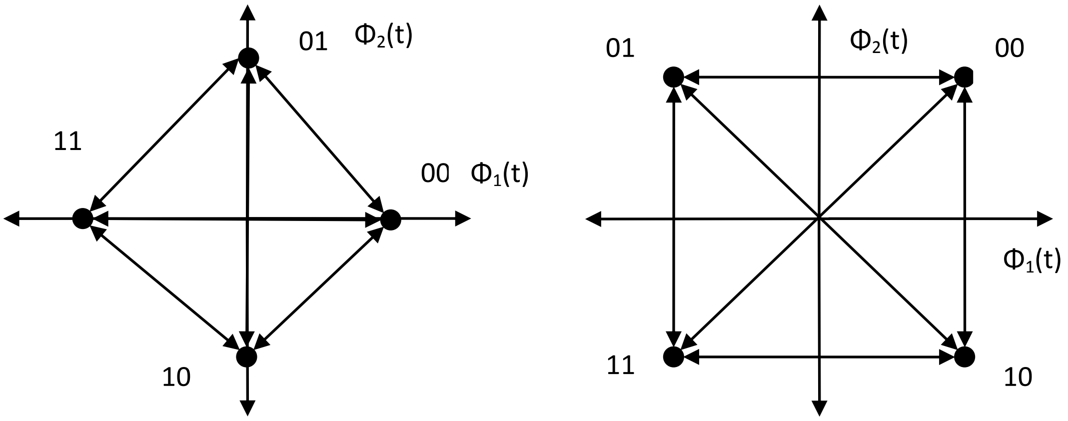

Figure 3.6: QPSK constellation for carrier phases 0, π/2,π,3π/2.

Based on the representation, a QPSK signal can be depicted using a two dimensional constellation diagram with four points as shown in the Figure 3.6. It should be noted that different QPSK signal sets can be derived by simply rotating the constellation.

From the constellation diagram of QPSK signal, it can be seen that the distance between adjacent points in the constellation is $\sqrt{2E_s}$. Since each symbol corresponds to two bits, then $E_s = 2E_b$, thus the distance between two neighbouring points in the constellation is equal to $2\sqrt{E_b}$.

3.3.2 QPSK modulator & demodulator :

The block diagram of QPSK modulator is shown in Figure 3.7 below [7]. The input data bit stream is first serial to parallel converted into two streams, $a_e(t)$ and $a_0(t)$, where $a_e(t)$ carries the even indexed data bits and $a_0(t)$ carries the odd indexed bits, or vice versa. The signal vector coefficients follow the following rule.

$$S_i = \begin{bmatrix} S_{i1} \\ S_{i2} \end{bmatrix} = \begin{bmatrix} \sqrt{E}\ \cos\left[(2i-1)\frac{\pi}{M}\right] \\ \sqrt{E}\ \sin\left[(2i-1)\frac{\pi}{M}\right] \end{bmatrix} = \begin{bmatrix} \pm\sqrt{\frac{E}{2}} \\ \pm\sqrt{\frac{E}{2}} \end{bmatrix} = \begin{bmatrix} \pm\sqrt{E_b} \\ \pm\sqrt{E_b} \end{bmatrix}. \quad (3.14)$$

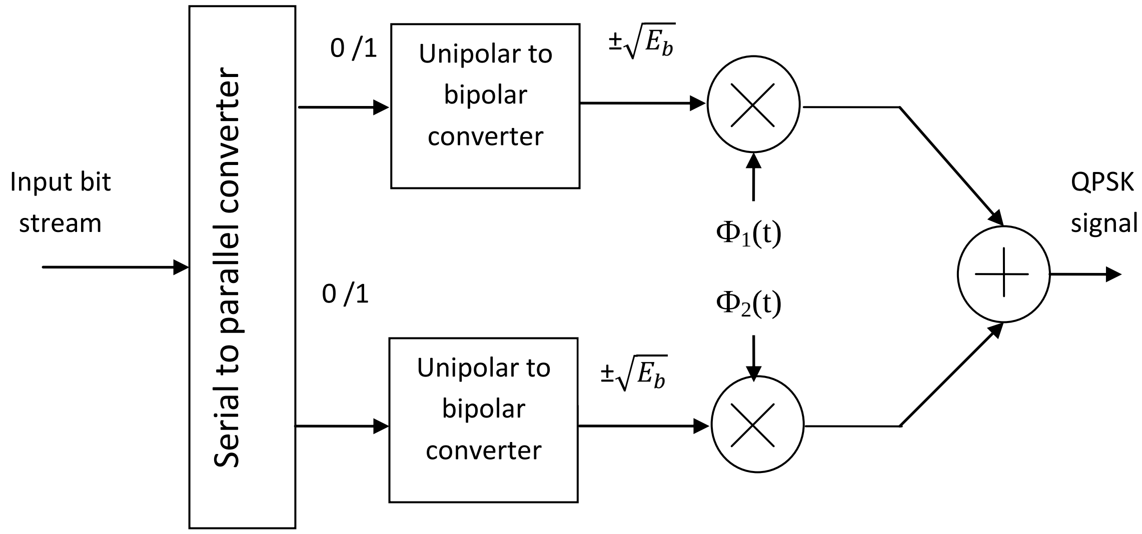

Figure 3.7: QPSK Modulator

If we associate bit "0" with the coefficient $-E_b^{\frac{1}{2}}$ and bit "1" with the coefficient $E_b^{\frac{1}{2}}$. The resultant QPSK constellation becomes the one shown in the Figure 3.6 above, which is a $\frac{\pi}{4}$ counter clock wise rotated version of QPSK constellation. Two of the conveniences mentioned before are related to the easy bit-to-coefficient conversion and to the fact that Gray mapping is being automatically applied by adopting the convention: bit "0" $\rightarrow -E_b^{\frac{1}{2}}$ and bit "1" $\rightarrow E_b^{\frac{1}{2}}$.

Let us take an input bit stream a(t), which is converted into even bit stream $a_e(t)$ and an odd bit stream $a_o(t)$. This bit streams are multiplied with Orthogonal sinusoidal carriers and the resultant outputs are shown in figures below [10]. By adding these two outputs, we will get the required QPSK output as shown Figure 3.8 below, which contains phase transitions.

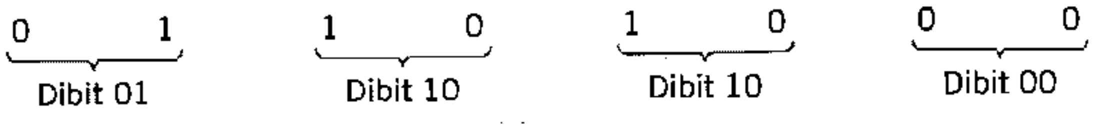

Figure 3.8 a : Input Binary Sequence

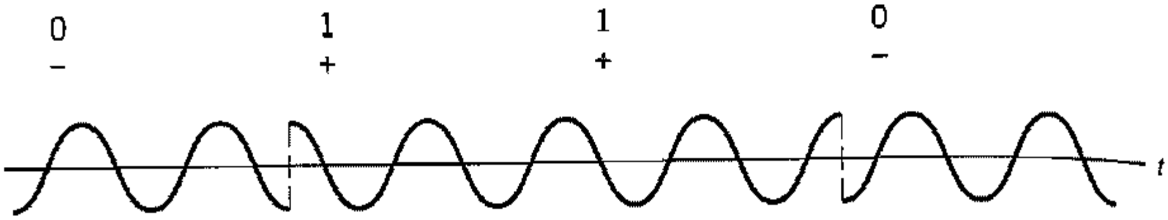

Figure 3.8 b: Binary PSK wave for odd numbered bit sequence

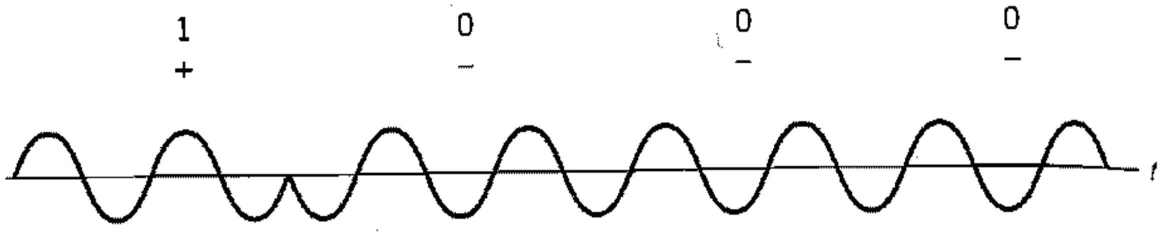

Figure 3.8 c: Binary PSK wave for Odd bit sequence

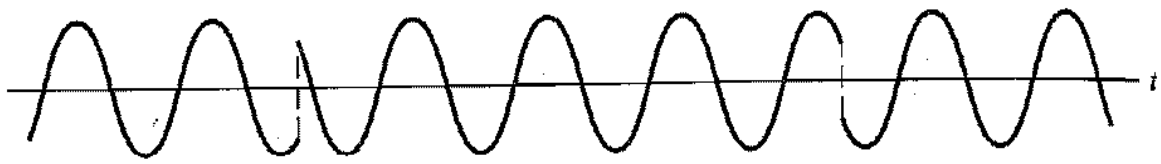

Figure 3.8 d: QPSK waveform

As a consequence, we can detect and estimate independently the streams $a_e(t)$ and $a_o(t)$ by means of two BPSK demodulators in parallel. This is illustrated in the coherent QPSK demodulator shown in Figure 3.9 below. Note that after the estimates of the bit streams $a_e(t)$ and $a_o(t)$ are performed, a parallel-to-serial converter produces the estimate of the transmitted data bit stream. As a consequence, we can detect and estimate independently the streams $a_e(t)$ and $a_o(t)$ by means of two BPSK demodulators in parallel. This is illustrated in the coherent QPSK demodulator shown in Figure 3.9. Note that after the estimates of the bit streams $a_e(t)$ and $a_o(t)$ are performed, a parallel-to-serial converter produces the estimate of the transmitted data bit stream.

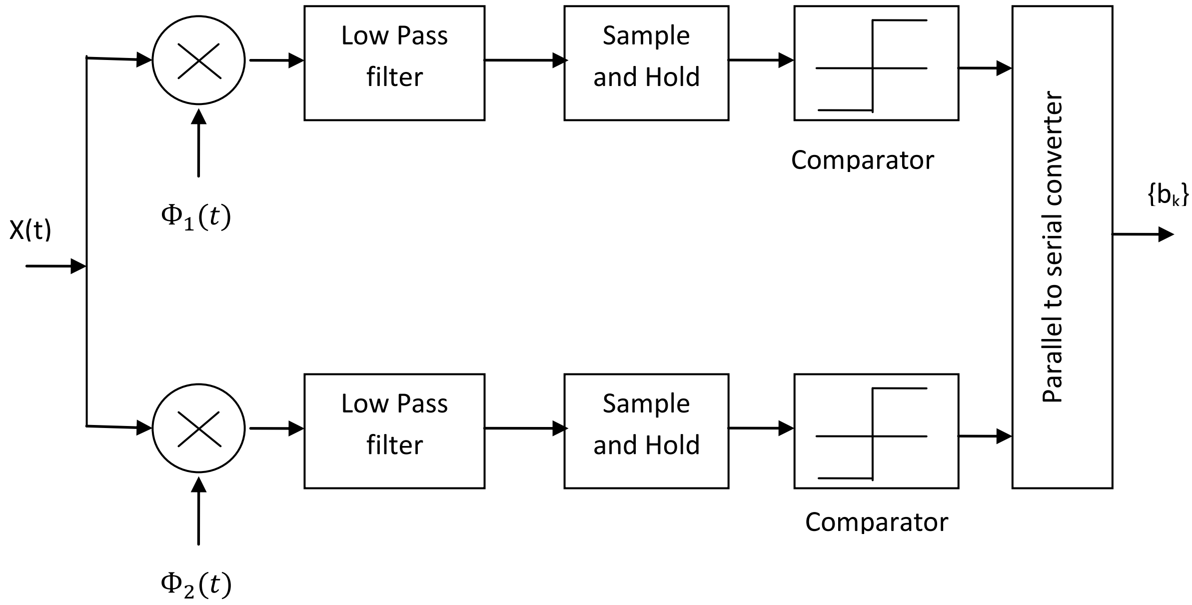

Figure 3.9 :QPSK demodulator by coherent detection

3.3.3 Spectral density of QPSK signal:

The power spectral density of a QPSK signal [9] can be obtained in a manner similar to that used for BPSK, with the bit periods T_b replaced by symbol periods. Hence, the PSD of a QPSK signal using rectangular pulses can be expressed as

$$P_{QPSK} = \frac{E_s}{2}\left[\left(\frac{\sin \pi(f - f_c)T_s}{\pi(f - f_c)T_s}\right)^2 + \left(\frac{\sin \pi(-f - f_c)T_s}{\pi(-f - f_c)T_s}\right)^2\right]$$

$$= E_b\left[\left(\frac{\sin 2\pi(f - f_c)T_b}{2\pi(f - f_c)T_b}\right)^2 + \left(\frac{\sin 2\pi(-f - f_c)T_b}{2\pi(-f - f_c)T_b}\right)^2\right] \tag{3.15}$$

And the Figure 3.10 shows the power spectral density of QPSK signal.

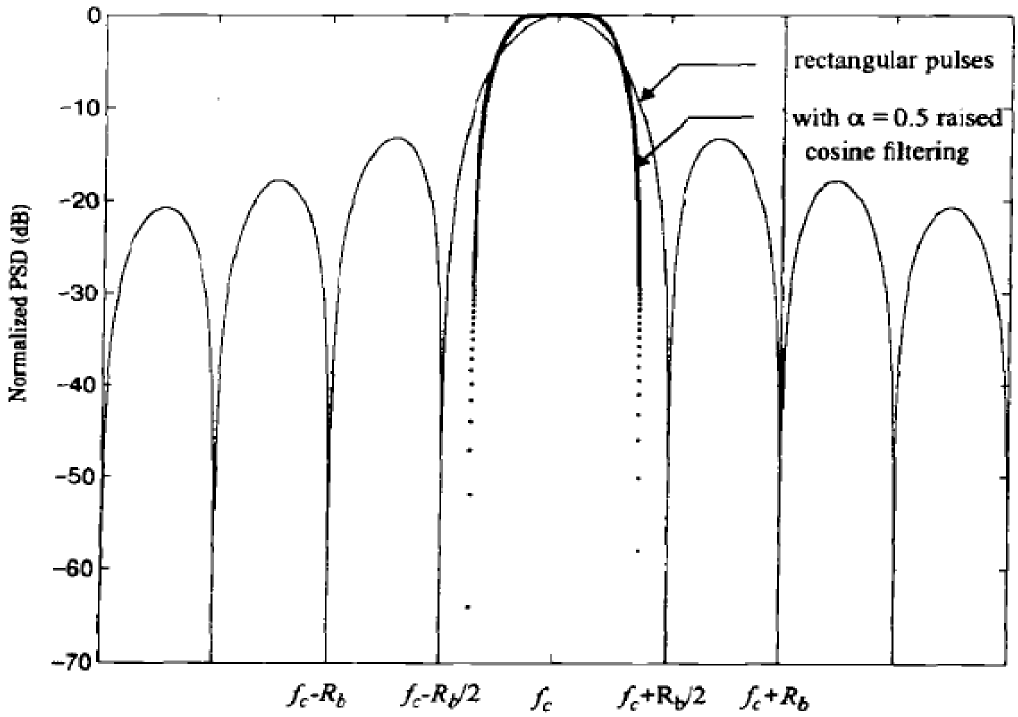

Figure 3.10: PSD of QPSK signal

The average probability of bit error in the additive white guassian noise channel is obtained as

$$P_{e,QPSK} = Q\left(\sqrt{\frac{2E_b}{N_o}}\right) \tag{3.16}$$

A striking result is that the bit error probability of QPSK is identical to BPSK, but twice as much data can be sent in the same bandwidth. Thus when compared to BPSK, QPSK provides twice the spectral efficiency with exactly the same energy efficiency. Similar to BPSK, QPSK can also be differentially encoded to allow noncoherent detection.

3.4 Offset Quadrature Phase Shift Keying(OQPSK)

This is an variant of Quadrature Phase Shift keying method and can be implemented by shifting the Quadrature bit stream by a half bit period. The following constellation plot describes the need for doing so.

3.4.1 constellation

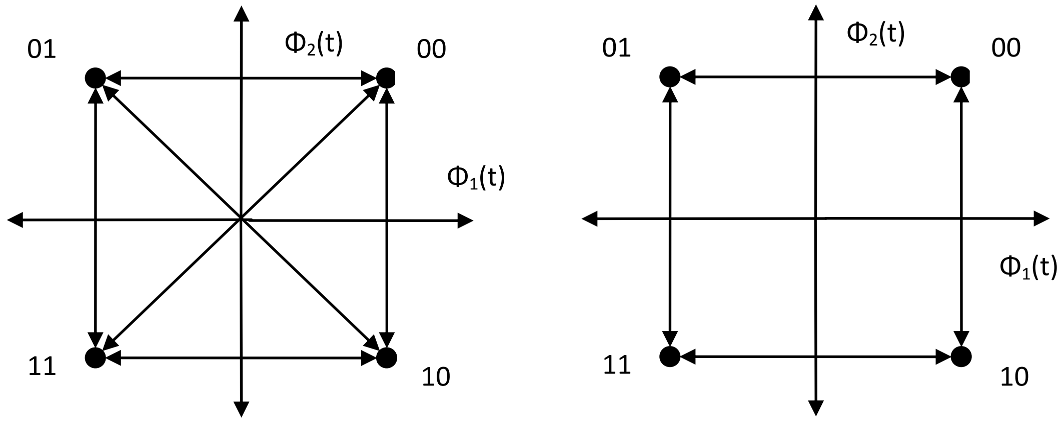

Figure 3.11:Phase transitions between QPSK and OQPSK

The constellation plot shown in Figure 3.11 above includes all possible phase transitions that can arise in the generation of a QPSK signal ([7], [9], [10]). From the QPSK waveform, we observe the following.

1. The carrier phase changes by $\pm$ 180^0 whenever both the inphase and Quadrature components of QPSK signal changes the sign. This situation occured when the input binary sequence switches from dibit 01 to dibit 10.

2. The carrier phase changes by $\pm$ 90^0 whenever the inphase or Quadrature components changes sign. This situation occurs when the input bit sequence switches from dibit 10 to dibit 00, during which the inphase component changes sign, whereas the Quadrature component is unchanged.

3. The carrier phase is unchanged when neither the inphase component nor the Quadrature component changes sign. This situation occurs when dibit 10 is transmitted in two successive symbol intervals.

Situation 1 and, to a much lesser extent, situation 2 can be of a particular concern when the QPSK signal is filtered during the course of transmission, prior to detection. Specifically, the 180^0 and 90^0 phase shifts in the carrier phase can result in changes in the carrier amplitude (i.e., envelope of the QPSK signal), thereby causing additional symbol errors on detection.

The extent of amplitude fluctuations exhibited by QPSK signals may be reduced by using offset QPSK. In this variant of QPSK, the bit stream responsible for generating the

Quadrature component is delayed(i.e., offset) by half a symbol interval with respect to the bit stream responsible for generating inphase component as shown in Figure 3.12. Specifically, the two basis function of offset QPSK are defined by

$$\phi_1(t) = \sqrt{\frac{2}{T}}\cos 2\pi f_c t \quad 0 \leq t \geq T$$
$$\phi_2(t) = \sqrt{\frac{2}{T}}\sin 2\pi f_c t \quad \frac{T}{2} \leq t \leq \frac{3T}{2} \tag{3.17}$$

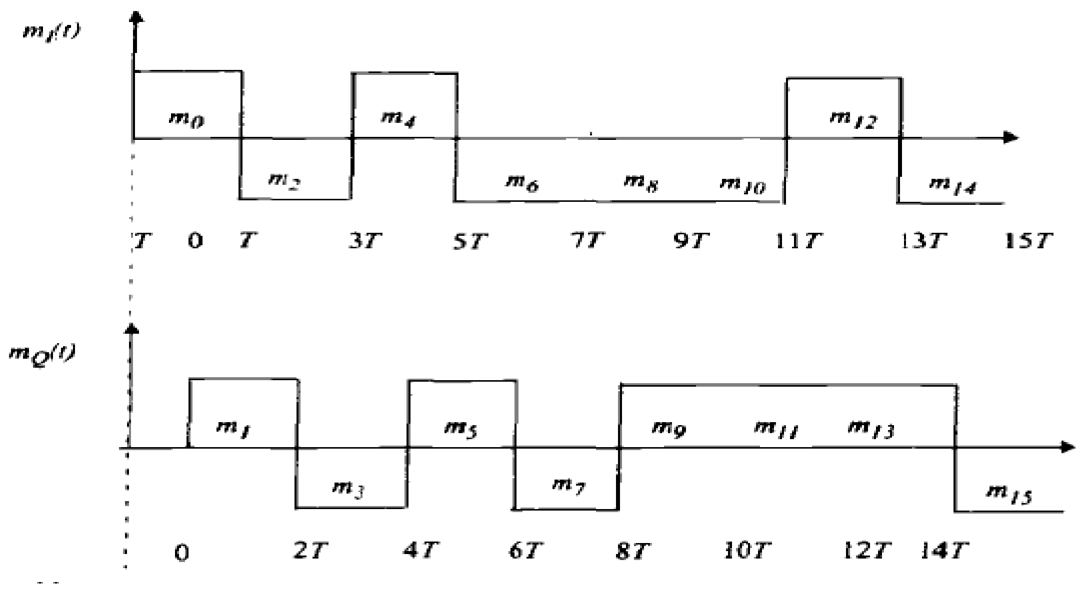

Figure 3.12: Output after adding delay in the quadrature arm of OQPSK modulator.

Unlike QPSK, the phase transitions likely to occur in offset QPSK are confined to $\pm 90^0$, as indicated in the signal space diagram. However, $\pm 90^0$ phase transitions in offset QPSK occur twice as frequently but with the half intensity encountered in QPSK. Since, in addition to $\pm 90^0$ phase transitions, $\pm 180^0$ phase transitions also occur in QPSK, we find that amplitude fluctuations in offset QPSK due to filtering have a smaller amplitude than in the case of QPSK. The equation for OQPSK modulated signal is given by

$$V_{OQPSK}(t) = \sqrt{P_s}\, b_e(t) \cos \omega_0 t + \sqrt{P_s} b_0(t) \sin \omega_0 t \tag{3.18}$$

Despite the delay $\frac{T}{2}$ applied to the basis function $\phi_2(t)$ in equation 3.18 , compared to the equation in QPSK, the offset QPSK has exactly same probability of symbol error in an AWGN channel as QPSK. The equivalence in noise performance between these phase shift keying schemes assumes the use of coherent detection. The reason for the equivalence is that statistical independence of the inphase and Quadrature components applies to both QPSK and OQPSK. We may therefore say that error probability in the inphase and Quadrature components applies to both QPSK and offset QPSK. We may therefore say that the error probability in the inphase or Quadrature channel of a coherent offset QPSK receiver is still equal to $\frac{1}{2}\mathrm{erfc}(\sqrt{\frac{E}{N_0}})$. Hence this formula applies equally well to the offset QPSK.

3.4.2 PSD of OQPSK signal

The spectrum of an OQPSK signal is identical to that of a QPSK signal, hence both signals occupy the same bandwidth [**9**]. The total power spectral density is twice the density generated by either of the Quadrature carriers and is given by

$$G_{OQPSK}(f) = E_b \left\{ \left[\frac{\sin \frac{2\pi(f-f_o)}{f_b}}{\frac{2\pi(f-f_o)}{f_b}} \right]^2 + \left[\frac{\sin \frac{2\pi(f+f_o)}{f_b}}{\frac{2\pi(f+f_o)}{f_b}} \right]^2 \right\} \tag{3.19}$$

The main lobe contains 90 percent of signal energy. Still, the not inconsiderable power outside the mainlobe is a source of trouble QPSK is to be used for multichannel communication on adjacent carriers. If we say, we establish additional channels at carrier frequencies $f_0' = f_0 + f_b$, then the sidelobe associated with first channel, having a peak value at frequency $f_0 + \frac{3f_b}{4}$, will be a source of serious interchannel interference. These sidebands are smaller than the main lobe by 14dB.

The staggered alignment of the even and odd bit streams does not change the nature of the spectrum. OQPSK retains its handlimited nature even after nonlinear amplification, and therefore it is very attractive for mobile communication systems where bandwidth efficiency and efficient nonlinear amplifiers are critical for 1ow power drain.

3.5 Minimum Shift Keying:

Due to the character of baseband signal, we get wide spectrum in QPSK ([7], [9], [10], [11]). This signal consists of abrupt changes, and abrupt changes gives rise to spectral components at high frequencies. In short, baseband spectral range is very large and multiplication by a carrier translates the spectral pattern without changing its form. We must try to alleviate this difficulty by passing the baseband signal through a low pass filter to suppress many sidelobes. Such filtering will cause inter symbol interference. The problem of intersymbol interference in QPSK is so serious that regulatory and standardisation agencies such as the FCC and CCIR will not permit the systems to be used expect with bandpass filtering at the carrier frequency to suppress the side lobes.

Also recall that QPSK and OQPSK are systems in which the signal is of constant amplitude, the information content being borne by phase changes. In both QPSK and OQPSK, there are abrupt phase changes in the signal. In QPSK these changes can occur at the symbol rate $\frac{1}{T_s} = \frac{1}{2T_b}$ and can be as large as 180^0. In OQPSK, phase changes of 90^0 can occur at the bit rate. When such waveforms with abrupt phase changes, are filtered to suppress sidebands, the effect of the filter, at the times of the abrupt phase changes, is to cause substantial changes in the amplitude of the waveform. Such amplitude variations can cause problems in QPSK communication systems which employ repeaters, i.e., stations which receive and rebroadcast signals such as earth satellites. For such stations often employ nonlinear power output stages in their transmitters. These nonlinear stages suppress the amplitude variations. However, precisely, because of their nonlinearity, they generate spectral components outside the range of the main lobe thereby undoing the effect of the bandlimiting filtering and causing inter channel interference. Finally, QPSK with its 180^0 phase changes is a substantially worse than the OQPSK with only 90^0 phase changes.

Now, we will discuss another important modulation technique called Minimum Shift Keying(MSK). There are two important differences between QPSK and MSK.

1. In MSK, the baseband waveform, that multiplies the Quadrature carrier, is much smoother than the abrupt rectangular waveform of QPSK. While the spectrum of MSK has a main centre lobe which is 1.5 times as wide as the main lobe of QPSK, the side lobes in MSK are relatively much smaller in comparison to the main lobe, making filtering much easier.

2. The waveform of MSK exhibits phase continuity, that is, there are not abrupt phase changes as in QPSK. As a result we avoid the intersymbol interference caused by nonlinear amplifiers.

We know that, in the coherent detection of binary FSK signal, the phase information contained in the received signal is not fully exploited, other than to provide for synchronisation of the receiver to transmitter. When performing detection, by proper use of the phase it is possible to improve the noise performance of the receiver significantly. This improvement is achieved at the expense of the increased receiver complexity.

MSK can be viewed as either a special case of continuous phase frequency shift keying (CPFSK), or a special case of offset QPSK with sinusoidal symbol weighting [12] or it can be treated as a special case of Frequency Shift Keying [13]. Consider a continuous phase frequency shift keying (CPFSK) signal, which is defined for the interval $0 \leq t \leq T_b$ as follows:

$$S(t) = \begin{cases} \sqrt{\dfrac{2E_b}{T_b}}\cos[2\pi f_1 t + \theta(0)] & \text{for symbol 1} \\ \sqrt{\dfrac{2E_b}{T_b}}\cos[2\pi f_2 t + \theta(0)] & \text{for symbol 0} \end{cases} \tag{3.20}$$

Where E_b is the transmitted signal energy per bit, and T_b is the bit duration. The phase $\theta(0)$, denoting the value of the phase at time t=0, sum up the past history of the modulation process upto time t=0. The frequencies f_1 and f_2 are sent in response to binary symbols 1 and 0 appearing at the modulator input, respectively.

Another useful way of representing the CPFSK signal s(t) is to express it in the conventional from of an angle modulated signal as follows [10]:

$$s(t) = \sqrt{\frac{2E_b}{T_b}}\cos[2\pi f_c t + \theta(t)] \tag{3.21}$$

Where $\theta(t)$ is the phase of s(t). Where the phase $\theta(t)$ is a continuous function of time, we find that the modulated signal s(t) itself is also continuous at all times, including the interbit switching times. The phase $\theta(t)$ of a CPFSK signal increases or decreases linearly with time during each bit duration of T_b seconds, as shown by

$$\theta(t) = \theta(0) \pm \frac{\pi h}{T_b} t, \quad 0 \le t \le T_b \tag{3.22}$$

Where the plus sign corresponds to sending symbol 1, and the minus sign corresponds to sending symbol 0; the parameter h is to be defined. Substituting θ(t) in s(t) and comparing it with CPFSK signal, we deduce the following pair of relations.

$$\begin{aligned} f_c + \frac{h}{2T_b} &= f_1 \\ f_c - \frac{h}{2T_b} &= f_2 \end{aligned} \tag{3.23}$$

Solving above two equations for f_c and h, we will get

$$\begin{aligned} f_c &= \frac{1}{2}(f_1 + f_2) \\ h &= T_b(f_1 - f_2) \end{aligned} \tag{3.24}$$

the nominal carrier frequency f_c is therefore arithmetic mean of the frequencies f_1 and f_2. The difference between frequencies f_1 and f_2 , normalized with respect to the bit rate $\frac{1}{T_b}$, defines the dimensionality parameter h, which is refered to as deviation ratio. Some researchers given a simplified technique for generating MSK signal [14].

3.5.1 Signal Space diagram of MSK

We may express the CPFSK signal S(t) interms of its inphase and Quadrature components as follows [10]:

$$S(t) = \sqrt{\frac{2E_b}{T_b}} \cos[\theta(t)] \cos(2\pi f_c t) - \sqrt{\frac{2E_b}{T_b}} \sin[\theta(t)] \sin(2\pi f_c t) \tag{3.25}$$

Consider first the inphase component $\sqrt{\frac{2E_b}{T_b}}\cos[\theta(t)]$. With the deviation ratio $h = \frac{1}{2}$, so that

$$\theta(t) = \theta(0) \pm \frac{\pi}{2T_b} t, \quad 0 \le t \le T_b \tag{3.26}$$

Where the plus sign corresponds to symbol 1 and the minus sign corresponds to symbol 0. A similar result holds for θ(t) in the interval $-T_b \le t \le 0$, except that the algebraic sign is not necessarily the same in both intervals. Since the phase θ(0) is 0 or π, depending on the past history of the modulation process, we find that, in the interval $-T_b \le t \le T_b$, the polarity of cos [θ(t)] depends only on θ(0), regardless of the sequence of 1's and 0's transmitted before or after t=0. Thus, for this time interval, the inphase component $S_I(t)$ consists of a half cycle cosine pulse defined as follows.

$$S_1(t) = \sqrt{\frac{2E_b}{T_b}}\cos[\theta(t)]$$

$$= \sqrt{\frac{2E_b}{T_b}}\cos[\theta(0)]\ \cos\left(\frac{\pi}{2T_b}t\right)$$

$$= \pm\sqrt{\frac{2E_b}{T_b}}\cos\left(\frac{\pi}{2T_b}t\right), -T_b \le t \le T_b \qquad (3.27)$$

Where the plus sign corresponds to θ(0) = 0 and the minus sign corresponds to θ(0) = π. In a similar way, we may know that, in the interval $0 \le t \le 2T_b$, the Quadrature component $S_Q(t)$ consists of a half cycle sine pulse, whose polarity depends only on $\theta(T_b)$, as shown by

$$S_Q(t) = \sqrt{\frac{2E_b}{T_b}}\sin[\theta(t)]$$

$$= \sqrt{\frac{2E_b}{T_b}}\sin[\theta(T_b)]\sin\left(\frac{\pi}{2T_b}t\right)$$

$$= \pm\sqrt{\frac{2E_b}{T_b}}\sin\left(\frac{\pi}{2T_b}t\right),\quad 0 \le t \le 2T_b \qquad (3.28)$$

Where the plus sign corresponds to $\theta(T_b) = \frac{\pi}{2}$ and the minus sign corresponds to $\theta(T_b) = -\frac{\pi}{2T_b}$. Here since phase states θ(0) and $\theta(T_b)$ can each assume one of two possible values, any one of the four possibilities can arise, as described here [9], [10].

1. The phase θ(0) = 0 and $\theta(T_b) = \frac{\pi}{2}$, corresponding to the transmission of symbol 1.
2. The phase θ(0) = π and $\theta(T_b) = \frac{\pi}{2}$, corresponding to the transmission of symbol 0.
3. The phase θ(0) = π and $\theta(T_b) = -\frac{\pi}{2}$, corresponding to the transmission of symbol 1.
4. The phase (0) = 0 and $\theta(T_b) = -\frac{\pi}{2}$, corresponding to the transmission of symbol 0

This is in turn, means that the MSK signal itself may assume any one of four possible forms, depending on the values of $\theta(0)$ and $\theta(T_b)$. From the expansion of CPFSK signal, we deduce that the orthonormal basis functions $\phi_1(t)$ and $\phi_2(t)$ for MSK are defined by a pair of sinusoidally modulated Quadrature carriers:

$$\phi_1(t) = \sqrt{\frac{2}{T_b}}\cos\left(\frac{\pi}{2T_b}t\right)\cos(2\pi f_c t), \quad 0 \le t \le T_b$$
$$\phi_2(t) = \sqrt{\frac{2}{T_b}}\sin\left(\frac{\pi}{2T_b}t\right)\sin(2\pi f_c t), \quad 0 \le t \le T_b \tag{3.29}$$

Correspondingly, we may express the MSK signal in the expanded form

$$S(t) = s_1\phi_1(t) + s_2\phi_2(t), \quad 0 \le t \le T_b \tag{3.30}$$

Where the coefficients s_1 and s_2 are related to the phase states $\theta(0)$ and $\theta(T_b)$, respectively. To evaluate s_1, we integrate the product s(t) $\phi_1(t)$ between the limits – T_b and T_b, as shown by

$$S_1 = \int_{-T_b}^{T_b} s(t)\,\phi_1(t)dt$$
$$= \sqrt{E_b}\cos[\theta(0)], \quad -T_b \le t \le T_b \tag{3.31}$$

Similarly to evaluate s_2 we integrate the product s(t) $\phi_2(t)$ between the limits 0 and $2T_b$, as shown by

$$S_2 = \int_{0}^{2T_b} s(t)\phi_2(t)dt$$
$$= -\sqrt{E_b}\sin[\theta(T_b)], \quad 0 \le t \le 2T_b \tag{3.32}$$

Note that from the above equations:

- Both the integrals are evaluated for a time interval equal to twice the bit duration.
- Both the lower and upper limits of the product integration used to evaluate the coefficients s_1 are shifted by the bit duration T_b with respect to those used to evaluate the coefficient s_2.

- The time interval $0 \leq t \leq T_b$, from which phase states $\theta(0)$ and $\theta(T_b)$ are defined in common to both the intervals.

Accordingly, the signal constellation for an MSK signal is two dimensional, with four possible message points, as shown in the Figure 3.13. The coordinates of the message points are as follows in a counter clock direction:

$(+\sqrt{E_b}, +\sqrt{E_b}), (-\sqrt{E_b}, +\sqrt{E_b}), (-\sqrt{E_b}, -\sqrt{E_b}), (+\sqrt{E_b}, -\sqrt{E_b})$.

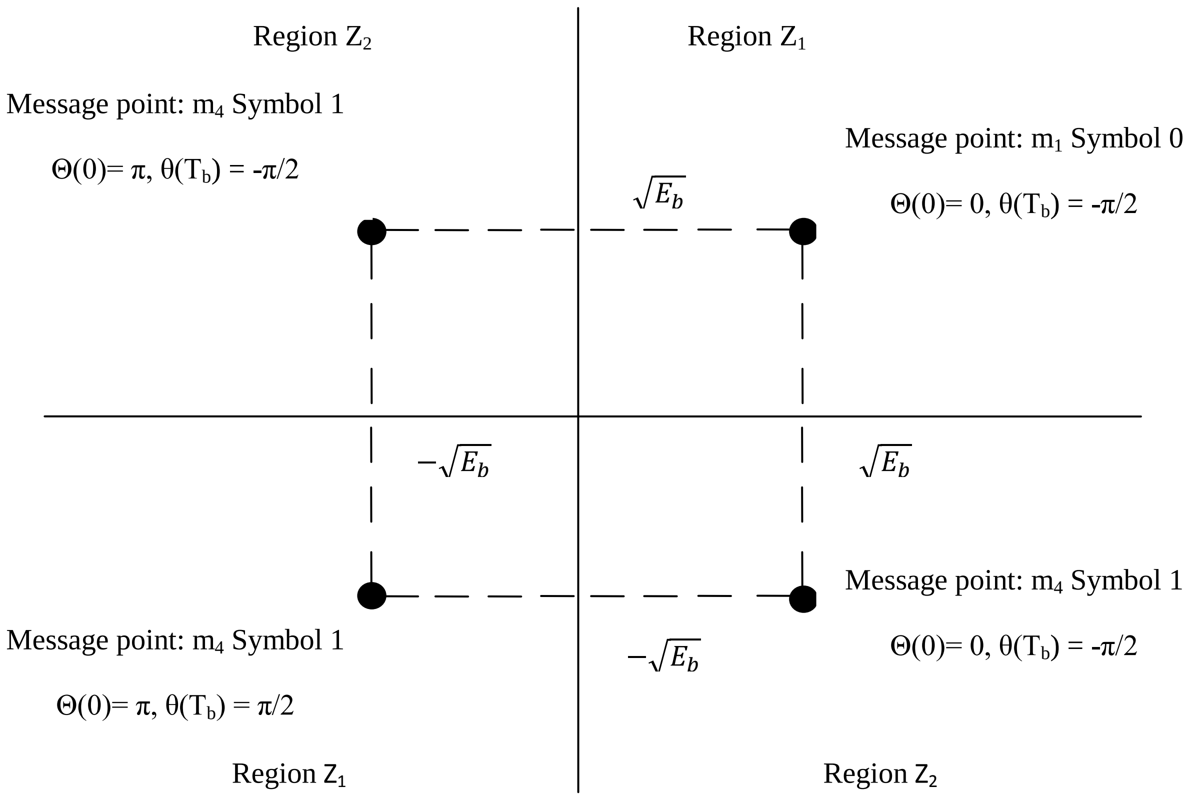

Figure 3.13: Signal space diagram of MSK

The possible values of $\theta(0)$ and $\theta(T_b)$, corresponding to those four message points, as shown in the Figure 3.13. The signal space diagram for the MSK is similar to the QPSK in that both of them have four message points. However, they differ in a subtle way that should be carefully noted. In QPSK the transmitted symbol is represented by any one of the four message points, whereas in MSK one of two message points is used to represent the transmitted symbol at any one time, depending on the value of $\theta(0)$.

Table-4 shows a summary of the values of $\theta(0)$ and $\theta(T_b)$, as well as the corresponding values s_1 and s_2 that are calculated for the time intervals $-T_b \leq t \leq T_b$ and $0 \leq t \leq 2T_b$, respectively. The first column of the table indicates whether symbol 1 or symbol

0 was sent in the interval $0 \le t \le T_b$. Note that the coordinates of the message points, s_1 and s_2, have opposite signs when symbol 1 is sent in this interval, but the same sign when the symbol 0 is sent. Accordingly, for a given input data sequence, we may use the entries in the table to derive , on a bit by bit basis, the two sequences of coefficients required to scale $\phi_1(t)$ and $\phi_2(t)$, and thereby determine the MSK signal S(t).

Table-4: Signal space characterisation of MSK.

Transmitted binary Symbol , $0 \le t \le T_b$	**Phase States in Radians**		**Coordinates of the message points**	
	$\Theta(0)$	$\Theta(T_b)$		
0	0	$-\frac{\pi}{2}$	$+\sqrt{E_b}$	$+\sqrt{E_b}$
1	π	$-\frac{\pi}{2}$	$-\sqrt{E_b}$	$+\sqrt{E_b}$
0	0	$+\frac{\pi}{2}$	$-\sqrt{E_b}$	$-\sqrt{E_b}$
1	π	$+\frac{\pi}{2}$	$+\sqrt{E_b}$	$-\sqrt{E_b}$

3.5.2 Mechanism for implementing MSK

The following example describes the implementation of MSK for the binary sequence 1101000. The input binary sequence is shown in the Figure 3.14 a below. The two modulation frequencies are: $f_1 = \frac{5T_b}{4}$ and $f_2 = \frac{3T_b}{4}$. Assumming that, at time t=0, the phase $\theta(0)$ is zero, the sequence of the phase states are shown in figure, modulo 2π. The polarities of the two sequences of factors used to scale the time functions $\phi_1(t)$ and $\phi_2(t)$ are shown in the top lines of the figure. Note that these two sequences are offset relative to each other by an interval equal to by an interval equal to the bit duration T_b. The waveforms of the resulting two components of S(t), namely $s_1(t)\phi_1(t)$ and $s_2(t)\phi_2(t)$, are also shown in the Figure 3.14 b and Figure 3.14 c. Adding these two modulated waveforms, we get the desired MSK signal S(t) shown in Figure 3.14.

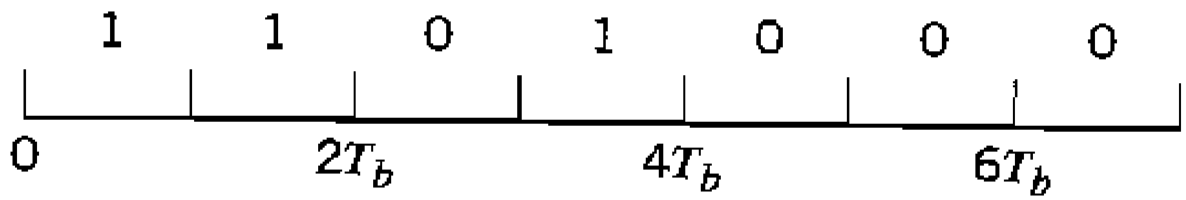

Figure 3.14 a: Input binary sequence

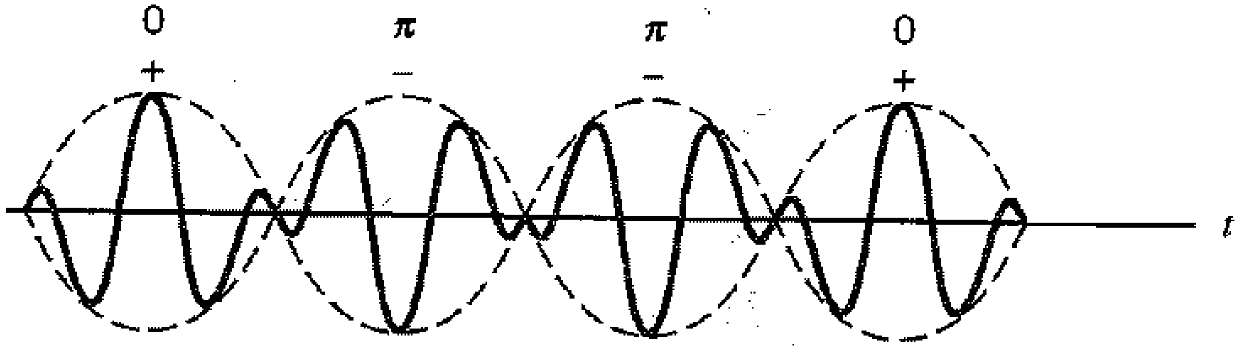

Figure 3.14 b: waveform of a scaled time version $s_1(t)\ \phi_1(t)$.

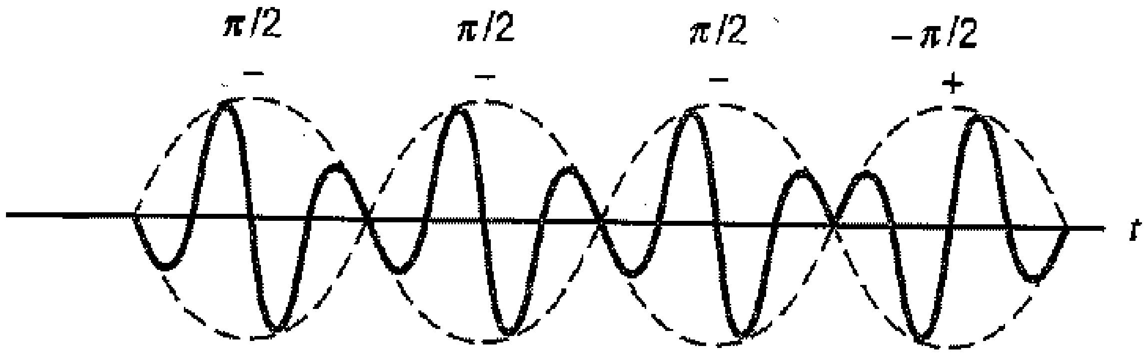

Figure 3.14 c: waveform of scaled time function $s_2(t)\ \phi_2(t)$

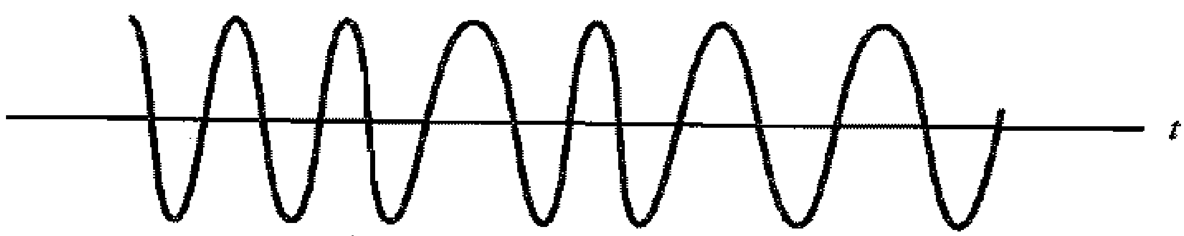

Figure 3.14 d: waveform of the MSK signal s(t) obtained by adding $s_1(t)\ \phi_1(t)$ and $s_2(t)\ \phi_2(t)$ on a bit by bit basis

The bit error rate probability for the coherent MSK is given by

$$P_e = \frac{1}{2}\,\mathrm{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right) \tag{3.33}$$

This is exactly same as that for binary PSK and QPSK. It is important to note, however, that this good performance is the result of the detection of the MSK signal being performed in the receiver on the basis of the observations over $2T_b$ seconds.

The details about the MSK modulator and demodulator are described in the next section.

3.5.3 PSD of MSK signal:

We know that RF power spectrum is obtained by frequency shifting the magnitude squared of the Fourier transform of the baseband pulse shaping function. For MSK, the baseband pulse shaping function is given by [10]

$$p(t) = \begin{cases} \cos\frac{\pi t}{2T} & t \leq |T| \\ 0 & \text{Else where} \end{cases} \tag{3.34}$$

Thus the normalized power spectral density for MSK is given by

$$P_{msk} = \frac{16}{\pi^2}\left(\frac{\cos 2\pi(f+f_c)T}{1.16f^2T^2}\right)^2 + \frac{16}{\pi^2}\left(\frac{\cos 2\pi(f-f_c)T}{1.16f^2T^2}\right)^2 \tag{3.35}$$

Figure 3.15 shows the power spectral density of an MSK signal. The spectral density of QPSK and is also drawn for comparison. From Figure 3.15, it is seen that the MSK spectrum has lower sidelobes than QPSK. Also the comparison between PSD of MSK and OQPSK are given in [15]. Ninety-nine percent of the MSK power is contained within a bandwidth B = $\frac{1.2}{T}$, while for QPSK, the 99 percent bandwidth B is equal to $\frac{8}{T}$. The faster rolloff of the MSK spectrum is due to the fact that smoother pulse functions are used. Also the main lobe of MSK is wider than that of QPSK, and hence when compared in terms of first null bandwidth, MSK is less spectrally efficient than the phase-shift keying techniques.

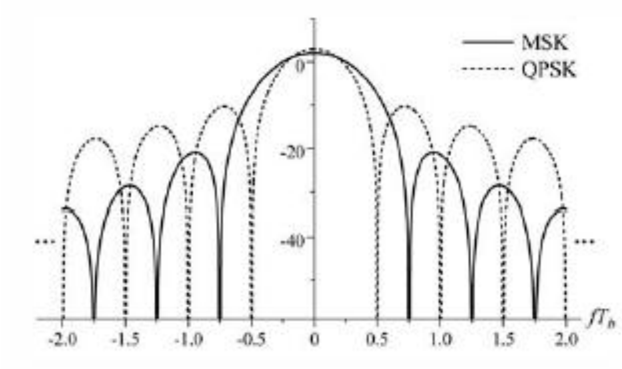

Figure 3.15: PSD of MSK signal as compared to the QPSK and OQPSK signals.

Since there is no abrupt change in phase at bit transition periods, bandlimiting the MSK signal to meet required out-of-band specifications does not cause the envelope to go through zero. The envelope is kept more or less constant even after bandlimiting. Any small variations in the envelope level can be removed by hardlimiting at the receiver without raising the out-of-band radiation levels. Since the amplitude is kept constant, MSK signals can be amplified using efficient nonlinear amplifiers. The continuous phase property makes it highly desirable for highly reactive loads. In addition to these advantages, MSK has simple demodulation and synchronization circuits. It is for these reasons that MSK is a popular modulation scheme for mobile radio communications.

CHAPTER 4

RADIO TRANSCEIVER ARCHITECTURES

CHAPTER 4

RADIO TRANSCEIVER ARCHITECTURES

4.1 INTRODUCTION

The first radio receivers were detectors connected directly to the antenna. Later they replaced with crystal detector, but the first separate component improving the sensitivity of the receiver was a vacuum tube rectifier with the property of amplification. Further improvements are achieved with tuned radio frequency (TRF) receivers using self oscillating detectors, and later with regenerative and super regenerative structures [14].

The use of separate local wave in the receiver producing a beat frequency to an audio frequency with the input signal in a nonlinear element is a principle of heterodyne. The beat frequency is defined as an intermediate frequency in a superhetrodyne receiver. Instead of direct detection, the signal provides higher amplication at IF stage and select a limited band of frequency with fixed filtering. Isolation between the amplifying stages is much easier at IF. Another significant benefit is that the narrower noise bandwidth after the IF filters. Due to these properties, the superhetrodyne receiver is the most sensitive structure for the radio reception.

There are mainly three types of radio Transceivers. These are Direct conversion or zero IF receiver, heterodyne receiver and Low intermediate frequency receiver [16].

4.2 RADIO TRANSCEIVERS

4.2.1 Superhetrodyne Receiver:

Superhetrodyne receiver uses frequency mixing or heterodyning to convert a received signal to a fixed intermediate frequency, which can be more conveniently processed than the original radio carrier frequency.

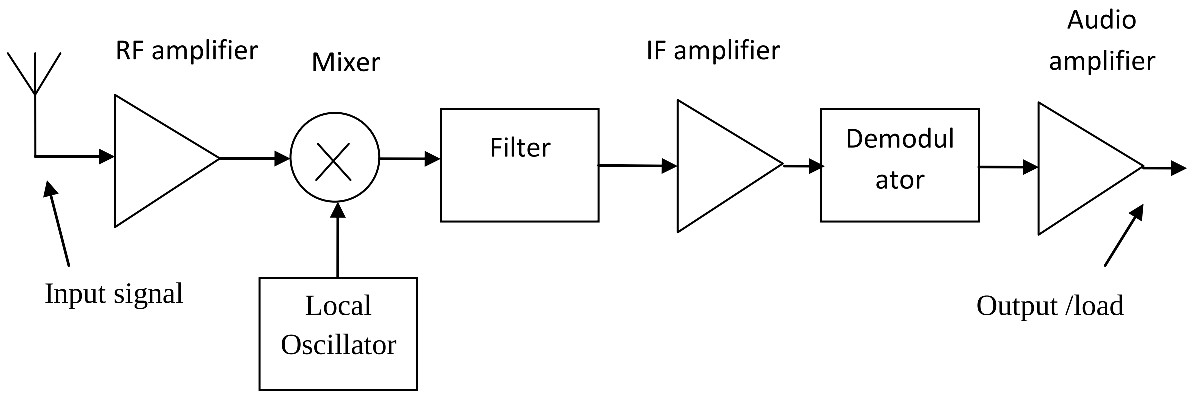

Figure 4.1:Block diagram of superhetrodyne receiver.

The principle of operation of the superhetrodyne receiver depends on the use of frequency mixing or heterodyning. The signal from the antenna is filtered sufficiently at least to reject the image frequency and possibly amplified. A local oscillator in the receiver produces a sine wave which mixes with that signal, shifting it to a specific intermediate frequency (IF), usually a lower frequency. The IF signal is itself filtered and amplified and possibly processed in additional ways. The demodulator uses the IF signal rather than the original radio frequency to recreate a copy of the original modulation (such as audio).

The Figure 4.1 shows the minimum requirements for a single-conversion superhetrodyne receiver design. The following essential elements are common to all superhet circuits: a receiving antenna, a tuned stage which may optionally contain amplification (RF amplifier), a variable frequency local oscillator, a frequency mixer, a band pass filter and intermediate frequency (IF) amplifier, and a demodulator plus additional circuitry to amplify or process the original audio signal.

The advantages of heterodyne scheme are [8]

- Good overall performance.
- Flexibility in frequency planning.
- No DC offset problem.
- Inphase and Quadrature matching is superior.

The disadvantages of superhetrodyne receiver are

- Requirement of external components.

- It suffers from image problems.
- Higher power consumption.
- High implementation results.
- Difficulties in multimode transceivers.

Due to high power consumption and high implementation costs, this superhetrodyne is not suitable for IEEE 802.15.4 standard.

4.2.2 Low IF receiver

In a low-IF receiver, the RF signal is mixed down to a non-zero low or moderate intermediate frequency, typically a few megahertz.

It combines the advantages of both heterodyne and direct conversion scheme.

The advantages of this receiver are [8]

- Avoids DC offset problem.
- Eliminate IF SAW filter, PLL and image filtering.

The disadvantage of this scheme is that it suffers impairments such as even order nonlinearities, local oscillator pulling and local oscillator leakage.

Mainly, the low IF scheme requires stringent image rejection as an adjacent channel becomes its image. Image rejection in low IF can be achieved either in analog or digital domain. In analog domain, we can reduce this image frequency by using complex bandpass filters. By the use of this band pass filters chip size and power consumption will increases. Also it requires a second low frequency digital mixer. In digital domain, the chip size will not increase. The disadvantage in digital domain solution is it requires good inphase and Quadrature phase matching. Also signal bandwidth in low IF is two times that of direct conversion, so ADC sampling rate will be doubled and hence power consumption will be higher. Another disadvantage is design complexity also increases.

4.2.3 Direct conversion scheme:

The direct conversion receiver converts the carrier of the desired channel to the zero frequency immediately in the first mixers. Hence the direct conversion is often called as a zero IF receiver, or a homodyne receiver if the LO is coherently synchronised with the incoming carrier. The synchronization of the LO directly to the RF carrier can be avoided with other techniques in current applications used mostly in optical reception.

Homodyne receivers translates the channel of interest directly from RF to baseband (ω_{IF}=0) in a single stage. Hence these architectures are called Direct IF architectures or Zero-IF architectures. For frequency and phase modulated signals, down conversion must provide Quadrature outputs so as to avoid loss of information. It translates the RF signal directly to the baseband signal.

The block diagram of direct conversion receiver is shown in the Figure 4.2 [14]. Two direct conversion mixers must be used for demodulation already at RF if a signal with Quadrature modulation is received. Otherwise, a single sideband signal with suppressed carrier contains Quadrature information, like QPSK, would alias its own independent single sideband channels in Quadrature over each other. Hence, the RF mixers are already a part of the demodulator although several other processing steps are performed before the detection of bits. This is also a distinct benefit of the direct conversion scheme. Because the information at the both sides of the carrier comes from the same source having an equal power. Hence the image power is always the same with the desired signal and the Quadrature accuracy requirements are only moderate. Thus the required image rejection is realizable with IC technologies even at high frequencies. A low pass filter with a bandwidth of half the symbol rate is suitable for the channel selection. This gives a noise advantages over other architectures and also image noise filtering is needed between the LNA and mixers. The external components in the signal path are now limited to the preselection filter at the input. Hence only the input of LNA must be matched in order to maintain the filter response unchanged. The interfaces between other blocks can be optimized during the design independently to optimize the performance with respect to noise, linearity and power. Of course, flexibility also increases the design complexity.

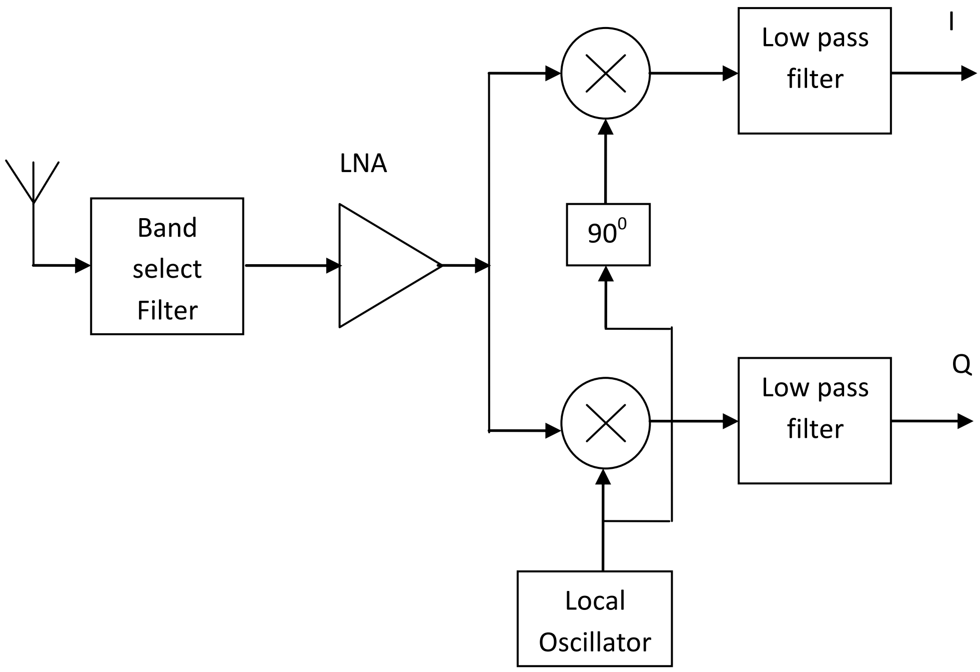

Figure 4.2: Block diagram of direct conversion receiver.

The advantages of this scheme are [8]

- Low cost.
- No image problem.
- No image filters.

The disadvantages of this scheme are

- DC offset.
- I/Q mismatch.
- Even order nonlinearity.
- Flicker noise or $\frac{1}{f}$ noise. This noise is very high in CMOS implementation.
- Local oscillator frequency planning while LO pulling and leakage.

In this type of receiver, DC offset and flicker noise are more dominant. DC offset can be cancelled easily without impairing the signal information using capacitive coupling method or dc offset cancellation technique.

Usually, flicker noise arises from random trapping charge at the oxide silicon interface. For CMOS technology, corner frequency is high (around 1MHz). The flicker noise can be reduced by using passive mixers or by using vertical NPN transistors in CMOS technology for the switching core of active mixers.

On comparing the three transceiver architectures, we were selected the direct conversion scheme because of low power consumption, low cost and high integrable chip. The detailed specifications of IEEE 802.15.4 standard with respect to modulation and demodulation are given in [8].

4.3 MAJOR FUNCTIONS IN TRANSMITTER DESIGN

We used Matlab/Simulink to design ZigBee transmitter design. The functions to be implemented in transmitter are bit to symbol mapping, symbol to chip mapping, serial to parallel conversion, half sine pulse shaping and modulation with high frequency carrier. Initially bit to Chip conversion is achieved by employing direct spread spectrum technique [8]. This spread spectrum technique , serial to parallel conversion and half sine pulse shaping are explained below briefly.

The spread spectrum modulation may be stated as two parts [11].

1. Spread spectrum is a means of transmission in which the data sequence occupies a bandwidth in excess of the minimum bandwidth necessary to send it.
2. The spectrum spreading is accomplished before transmission through the use of a code that is independent of data sequence. The same code is used in the receiver to despread the received signal so that the original data sequence may be recovered.

Spread spectrum modulation was originally developed for military applications, where resistance to jamming is of major concern. However the civilian applications were also benefied from the unique properties of the spread spectrum modulation. For example, in multiple access communications in which a number of independent users are required to share a common communication channel without the external synchronism mechanism, spread spectrum became robust today. There are two types of the spread spectrum techniques.

1. Direct Sequence Spread Spectrum (DSSS)
2. Frequency Hop Spread spectrum (FHSS)

Generally in DSSS, the incoming data sequence is used to modulate a wideband code. This code transforms the narrowband data sequence into a noise like wideband signal. The resulting wideband signal undergoes a second modulation using phase shift keying.

Serial to Parallel conversion and Half sine pulse shaping:

Serial to parallel conversion is implemented by a set of predefined configuration of flip flops, which is explained below([11], [17]). We know that D-flip flop is a one bit storage device. The operation of the D-flip flop is such that at the active edge of the clock waveform, the logic level at D is transferred to the output Q. As a matter of fact, the change in data d(t) will occur slightly after the active edge. Once the flip-flop, in response to an active clock edge, has registered a data bit, it will hold that bit until updated by the occurrence of the next succeeding clock edge.

The mechanism of generating OQPSK waveform is described below [11]. The fundamental circuit for generating the inphase and quadrature data by generating even and odd clock is shown in the Figure 4.3. If we add one more d-flipflop after the D-flipflop at inphase component, we can get QPSK signal. The delay provided by the D-flipflop should be half of bit duration.

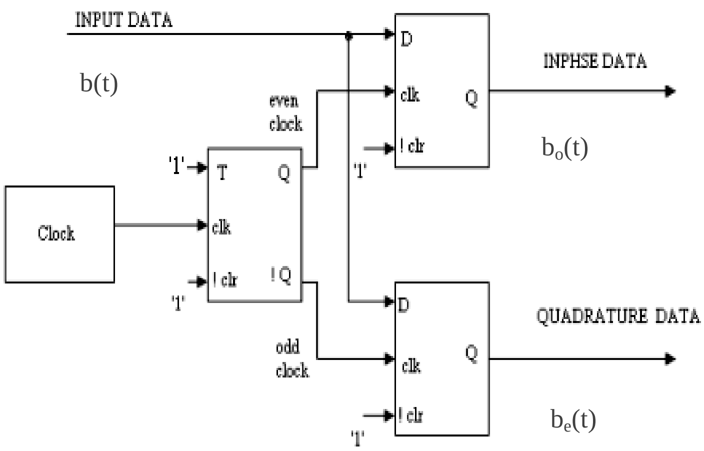

Figure 4.3: Generation of Inphase and Quadrature data.

Here we have arbitrarily assumed that in every case the active edge of the clock waveforms is the downward edge i.e., negative triggering. We know that T-flip flop toggles its output if the input is ‘1’. Here let the toggle flip-flop is driven by a clock waveform whose period is the bit time T_b. The toggle flip-flop generates an even clock waveform and an odd waveform. These clocks have bit period $2T_b$. The active edge of one of the clocks and the active edge of the other are separated by the bit time T_b. Let the bit stream b(t), which is applied as the data input to both type-D flip flops, one driven by the odd and the other is driven by the even clock waveform. The flip-flops register alternate bits in the bit stream b(t) and hold each such registered bit for two bit intervals, that is for a time $2T_b$. If there are input bit sequence numbered 1,2,3,4 etc., then the bit stream $b_o(t)$ registers bit 1 and holds that bit for time $2T_b$ and next it registers bit 3 for time $2T_b$ and next bit 5 for $2T_b$, etc. The even bit stream $b_e(t)$ holds, for times $2T_b$ each , it results the alternate bits numbered 2,4,6 etc.

The bit stream $b_e(t)$ is superimposed on a carrier $\sqrt{2Ps}\cos w_o t$ and the bit stream $b_o(t)$ is superimposed on a carrier $\sqrt{2Ps}\sin w_o t$ by the use of two multipliers, to generate two signals. Let them as $s_0(t)$ and $s_e(t)$ respectively. These signals are then added to generate the transmitted output signal $v_m(t)$ which is given by

$$v_m(t) = \sqrt{2Ps}\ b_o(t) \cos w_0 t + \sqrt{2Ps}\, b_e(t) \sin w_o t \tag{4.1}$$

Where let $S_0(t) = \sqrt{2Ps}\ b_o(t) \cos w_0 t$

$$S_e(t) = \sqrt{2Ps}\, b_e(t) \sin w_0 t$$

Here both $S_e(t)$ and $S_0(t)$ occupy the same spectral range but they are individually identifiable because of the phase Quadrature of the carriers. These four possible output signals have equal amplitude $\sqrt{2Ps}$ and are in phase quadrature; they have been identified by their corresponding values of b_0 and b_e.

In OQPSK, abrupt phase changes occurs due to the multiplication of the abrupt rectangular baseband waveform with the Quadrature carrier. So there is no phase continuity. These abrupt changes give rise to spectral components at high frequencies. In short, the baseband spectral range is very large and multiplication by a carrier translates the spectral pattern without changing its form. It is required to alleviate the difficulty by passing the baseband signal through a low pass filter to suppress the side lobes. Such filtering cause inter symbol interference. In OQPSK, the signal is of constant amplitude, the information content being changed by phase changes in the signal. In OQPSK, the phase changes of 90^0 can occur at the bit rate. When such waveforms with abrupt phase changes, are filtered to suppress the sidebands, the effect of the filter, at the times of abrupt phase changes, is to cause substantial changes in the amplitude of the waveform. Such amplitude variations can cause the problems in QPSK communication systems which employ repeaters, i.e., stations which receive and rebroadcast signals such as earth satellites. For such stations, we employ nonlinear power output stages in their transmitters. These nonlinear stages suppress the amplitude variations. However, precisely because of nonlinearity, they generate spectral components outside the range of the main lobe there by undoing the effect of the bandlimiting filtering and causing interchannel interference. This problem can overcome by, we are using half sine pulse shaping. In this design, half sine was used for pulse shaping to exhibit phase continuity and to smoothen the baseband waveform. This is achieved by multiplying sine wave with a the bit period of $4T_b$ in both I & Q channel.

4.4 Major Functions in Receiver design:

The functions to be implementing in receiver are RF to base band conversion, sampling and thresholding, parallel to serial conversion and despreading. The received signal from the channel is coherently correlated in one arm of the receiver with the result of the multiplication between the in-phase carrier and the shaping function $\cos\left(\frac{\pi t}{2T_b}\right)$. In the other arm, the received signal is correlated with the result of the multiplication between the Quadrature carrier and the shaping function $\sin\left(\frac{\pi t}{2T_b}\right)$. These correlations are made in a $2T_b$ interval, reflecting the duration of the half-cycle cosine and sine shaping pulses, and are time-aligned with these pulses. The estimated sequences at both inphase and Quadrature arms are parallel to serial converted to form serial data of the transmitted sequence. After that signal is despreaded. This signal must be same as the signal that we have sent at the transmitter side [7].

4.5 Comparison between MSK and OQPSK modulators:

In light of the similarities between the MSK and OQPSK modulations, it is possible to design MSK demodulator. It is known from previous discussions that a conventional QPSK modulator can be interpreted as two independent BPSK modulators, each of them making use of one out of two Quadrature carriers. As a consequence, the QPSK demodulator can be implemented as two independent BPSK demodulators. The decisions made by each of them are parallel-to-serial (P/S) converted to form the estimate of the transmitted bit sequence. The OQPSK demodulator follows the same rule, with the difference that one of the estimated parallel streams is offset T_b seconds from the other. Then, before parallel-to-serial conversion these sequences must be aligned in time [7].

The detailed explanation about ZigBee transceiver is given in the next chapter.

CHAPTER 5

IMPLEMENTATION OF ZIGBEE TRANSCIEVER IN MATLAB/SIMULINK

CHAPTER 5

IMPLEMENTATION OF ZIGBEE TRANSCEIVER IN MATLAB/SIMULINK

5.1 Introduction:

In this chapter, the block diagrams of ZigBee transmitter and receiver are to be explained. The major functions like, bit to symbol mapping, symbol to chip mapping, serial to parallel conversion, performing half sine pulse shaping and performing modulation are discussed in the transmitter part. RF to baseband conversion, sampling and thresholding, parallel to serial conversion and despreading were discussed in the receiver part.

5.2 Design of ZigBee Transmitter:

This section describes the implementation of ZIGBEE transmitter system. The implementation was built on Matlab/Simulink using fundamental components in Simulink to demonstrate how reliably complex modulation schemes can be built, cost effectively and efficiently. The design of ZigBee transmitter using OQPSK modulation with half sine pulse shaping is shown in the Figure 5.1 given below ([7], [15]). Here the input bit stream is having a data rate of 250Kbps.

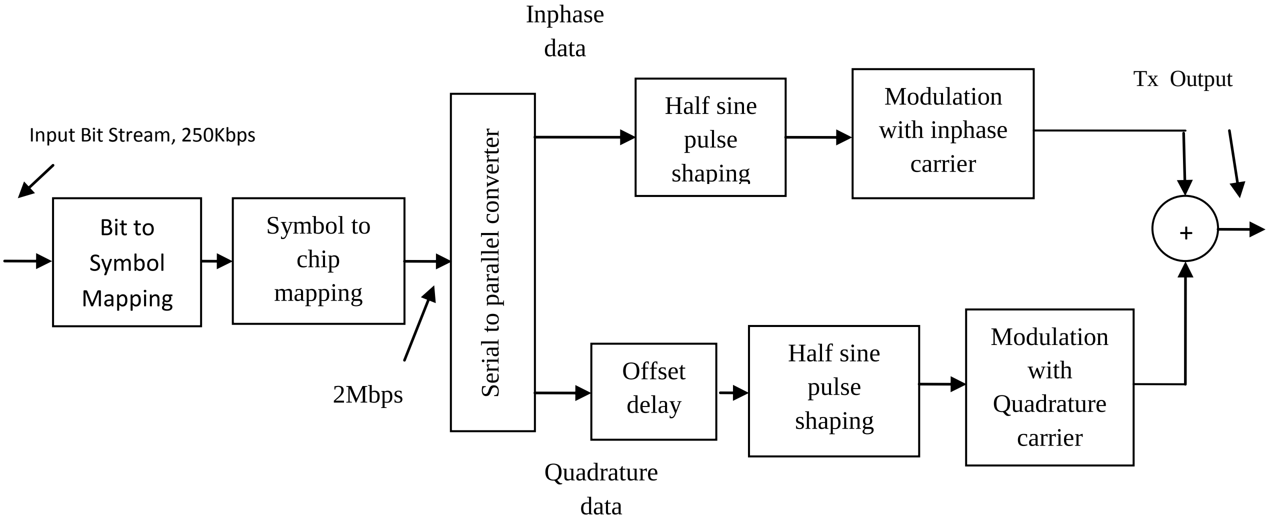

Figure 5.1: Block diagram of ZigBee Transmitter

Now we are mapping 4 input data bits to a symbol having a symbol rate of 62.5Kilo symbols per second. The symbol is then used to select one of 16 nearly orthogonal 32-chip PN sequences to be transmitted and results in a chip rate of two mega chips per second. After that, resultant chip sequence is send to the serial to parallel converter [8]. It is used here to separate the even indexed chips and odd indexed chips. Following this half sine pulse shaping is performed and signal modulated with a 2.4 GHz carrier on the I and Q data stream and add it to get the required transmitter output signal. Step by step procedure to implement ZigBee transmitter using simulink is presented below.

5.2.1 Bit to symbol and Symbol to chip mapping

i. Generating binary data stream: *By using Random integer generator block in the* communication tool box, we can generate binary data stream of 250Kbps. By adjusting the parameters like M- ary number, initial seed, sample time and output data type, we can achieve the fixed binary stream. In a real time scenario, this data stream is supplied by application that will generate information to be transmitted.

ii. *Generating PN sequence*: By using PN sequence generator block in the communication tool box, we can generate 32 bit- PN sequence of 2Mbps data rate. By adjusting the parameters like generator polynomial, initial states, sample time and output data type, we can achieve 32 bit Pseudo Noise code.

iii. *Generating DSSS signal*: After generating the input bit stream and PN sequence, the code generated is converted to NRZ format and multiplied to get a Direct spread spectrum signal.

5.2.2 Serial to parallel converter implementation:

By using flip-flops in Simulink extras tool box, we can get the parallel data from the serial data. The necessary instruments are one clock, one JK flip flop and two D flipflops. The initial conditions of the flip-flops using is zero and the period of the clock was decided by the input data stream. By this way we can easily generate the parallel data technically called as inphase and Quadrature data. Here necessary one bit offset delay is provided by the D flip flop itself.

5.2.3 Performing Half sine Pulse shaping

By the multiplication of NRZ form of inphase data with sine wave generator, we can get a smoothened version of baseband waveform. But we must take care while designing this because edge of inphase data must be in synchronize with phase of the half sine wave .

5.2.4 Performing Modulation :

After obtaining the inphase and Quadrature signals after half sine pulse shaping, we need to do modulation for the transmission of the signal. Generally we do this with the help of high frequency(2.4GHz) sinusoidal carrier. By using sine wave block in Signal Processing Tool Box, sine wave can be generated by adjusting the parameters like amplitude, frequency, sample time, phase and sine type. Now the inphase signal after half sine pulse shaping is multiplied by a sine wave and Quadrature signal after half sine pulse shaping is multiplied by its orthogonal carrier i.e., cosine signal which is nothing but 90^0 phase shift of original sinusoidal carrier.

5.2.5. Output of the ZigBee Transmitter :

Addition of both inphase and Quadrature signals after modulation, generates the required transmitter output. The required output signal is generated by using sum block in commonly used blocks. There will be no phase transitions in the output, which is an advantageous property.

5.3 Design of ZigBee receiver:

There are two type detection schemes available for the detection of original baseband data. They are coherent detection and non-coherent detection. In coherent detection, the phase of carrier that we used in the transmitter and phase of recovered carrier must be same. So proper carrier synchronization is necessary in the coherent demodulation. In case of non-coherent demodulation, there is no need of carrier synchronization. Coherent detection is costlier to implement, that is, the receiver must be equipped with a carrier recovery circuitry, which inturn increases system complexity, and can increase size and power consumption. Additionally, there is no ideal carrier recovery circuit. So, no practical digital communication system works under perfect phase coherence. While Non coherent detection uses previous bit information for extracting the original data and there is no need of using the carrier recovery circuit. Non-coherent detection is simpler, but it suffers from performance degradation as compared to coherent detection, but this difference can be small in practice for some modulation schemes due to the specifics of the modulation and also due to the penalty caused by imperfections in the carrier recovering process [7].

This section describes the implementation of ZigBee receiver system. Here we are concentrating on the MSK coherent detection technique for recovering original data in receiver. The block diagram of the ZigBee Receiver is shown in Figure 5- 2 below.The step by step procedure to implement ZigBee receiver using Simulink is presented below.

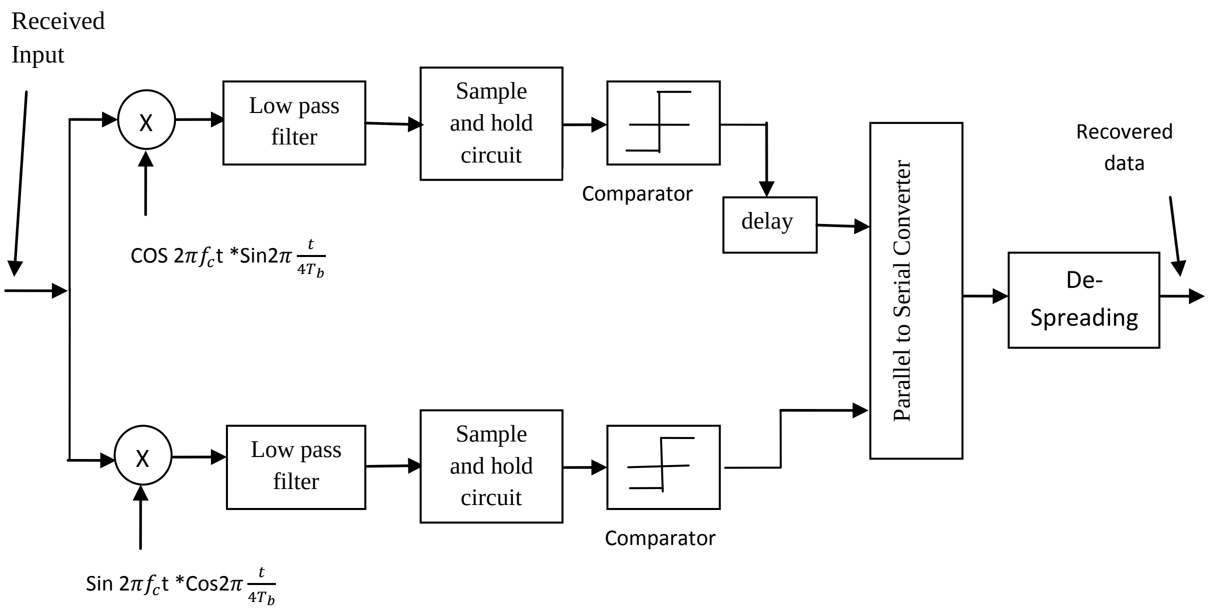

Figure 5- 2: Design of ZigBee Receiver.

In the receiver configuration of ZigBee, we are using a MSK demodulator and a multiplier for despreading. This multiplier is supplied by a PN sequence data that is an exact replica that used in the transmitter. The data coming from the MSK demodulator (i.e. at the parallel to serial converter) is having a data rate of 2Mbps. From this data, the original data is extracted by multiplying with the PN sequence data. But the 2Mbps data obtained at the output of parallel to serial converter contains some offset delay. This offset delay must introduced in the PN sequence data while multiplying with 2Mbps data, So that output contains original bit stream without any errors.

The receiver design is explained below. The incoming received signal is applied to two synchronous demodulators, consisting of a multiplier followed by a lowpass filter. Here one multiplier is supplied with a signal, which is the multiplied signal of carrier cosw$_0$t and cos($2\pi \frac{t}{4T_b}$). Another multiplier is supplied with quadrature signals i.e., with sinw_0t*sin($2\pi \frac{t}{4T_b}$). This multiplier output contains the baseband data and higher frequency harmonics. The multiplied signal is passed through a lowpass filter for avoiding harmonics. Generally, a third order Butterworth filter having a cutoff frequency of $\frac{2}{T_b}$Hz is used for the extraction of baseband data. The resultant data is passed through a sampler. A simple zero order sample and hold circuit is enough for this purpose. This device simply holds the data for one bit period.

This sampled data is passed through a decision device. Decision device is a simple comparator which contains a threshold value for making a decision. If the input to the comparator is greater than the threshold value, it decodes the bit as '1' otherwise it decodes as '0'. Suppose if S_1 and S_2 are two input data levels, then we used to take the threshold as

$$S^*(t) = \frac{S_1+S_2}{2}$$

Therefore, the output of comparator is either '1' or '0'. Remember, at the transmitter, we introduced an offset delay of one bit period at the Quadrature data. For compensating that delay effect, at the receiver, we will introduce same one bit delay at the output of comparator in inphase side. After that, by using a multiplexer or a parallel to serial converter, multiplex the delayed inphase data and Quadrature data coming from comparator. This data is having a data rate of 2Mbps which is nothing but transmitted Direct Spread Spectrum data. Now we have to pass this data through a multiplier. This multiplier is supplied by a PN sequence data that is an exact replica that we used in the transmitter. Actually 2Mbps data coming from parallel to serial converter contains a small amount of offset delay. We must introduce this offset delay in the PN sequence data while multiplying with 2Mbps data, So that we will get original bit stream without any errors.

Step by step procedure to implement ZigBee receiver using Simulink is presented below.

A. RF to Baseband conversion
B. Sampling and thresholding.
C. Parallel to serial conversion.
D. Despreading.

5.3.1 RF to Baseband conversion:

i. *Multiplying with RF carrier*: By using sine wave block in the signal processing tool box, a sine wave of frequency 2.4GHz is generated. By adjusting the parameters like amplitude, frequency, sample time, phase and sine type, required sine wave was generated.

ii. *Multiplying with Half sine wave*: By using sine wave block in the signal processing tool box, a sine wave of frequency 500KHz can be generated. By adjusting the parameters like amplitude, frequency, sample time, phase and sine type, the required sine wave was generated.

iii. Lowpass filtering: By using analog filter design block in the signal processing tool box, a low pass filter was generated to pass the resultant data. Generally, a third order Butterworth low pass filter is sufficient for filtering. By adjusting the parameters like design method, filter type, filter order and pass band edge frequency, the required low pass filter was generated.

5.3.2 Sampling and thresholding :

i. Sampling: By using the zero order hold circuit in the Simulink discrete menu, a sample and hold circuit was generated. It samples the signal for every T time period. By setting the sample time in this block, adjust the time period T in zero order hold circuit.

ii. Thresholding: By using compare to constant block in the Simulink logic and bit operations menu. By setting operator, constant value and output data type parameters, we can get the comparator circuit, which compares the sampled data with the predefined threshold value and detects whether the transmitted data is ‘1’ or ‘0’.

5.3.3 Parallel to serial conversion:

By using switch block in the Simulink signal rooting menu, convert the parallel data into serial data. By setting the threshold value and the criterion for parallel to serial conversion, convert parallel data into serial data.

5.3.4 Despreading:

The resulting data coming after serial to parallel conversion is multiplied with the delayed PN sequence. So that original is recovered data with small amount of delay.

The incoming bit stream and the resultant output both are same but with a small amount of delay.

CHAPTER 6

SIMULATION RESULTS

CHAPTER 6

SIMULATION RESULTS

6.1 SIMULATION RESULTS: STEP BY STEP

From the previous chapters, it is visible that there are two types of modulations used in the ZigBee transmitter. They are direct spread spectrum modulation and modulation with high frequency carrier. The detailed explanation about the simulation results of transmitter part was explained in section 6.1.1. Similarly, at the receiver end, demodulation and despreading operations have to be performed. The explanation about simulation results of receiver was explained in 6.1.2.

6.1.1 **At the transmitter end:**

The simulation results at each step are shown below. The following figures demonstrate simulation results for ZigBee transmission system. The results are displayed in the form of snapshots of scope signals. These figures demonstrate we can know easily what happens exactly inside a ZigBee transmitter.

The input stream is generated from a random integer generator. This is presented in Figure 6.1. It has a data rate of 250Kbps. i.e., it generates a bit for every 4μs as shown in the time axis. Each division for the time axis is taken as 10μs.

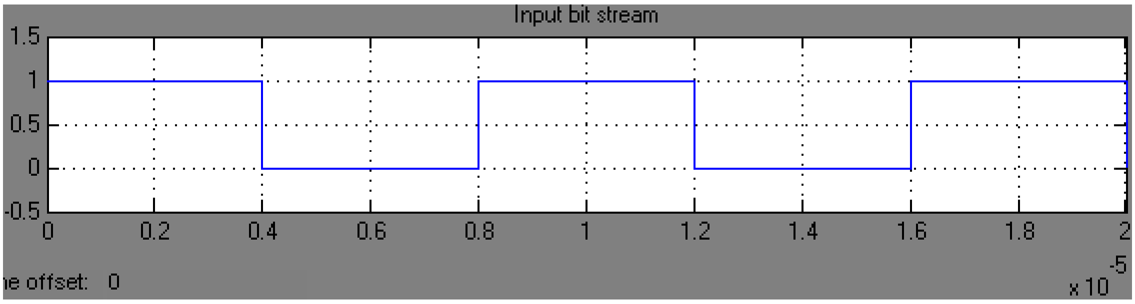

Figure 6.1: Input bit stream.

The 250Kbps input data is mapped into a symbol, making the symbol rate 62.5Kilo symbols per second. The Pseudo Noise code is generated from a PN sequence generator. This is presented in Figure 6.2 and it has a data rate of 2Mbps. That means each bit in the PN sequence is having a time period of 0.5 µs. Clearly the chip rate is equal to eight times the bit rate or it is equal to 32 times of symbol rate.

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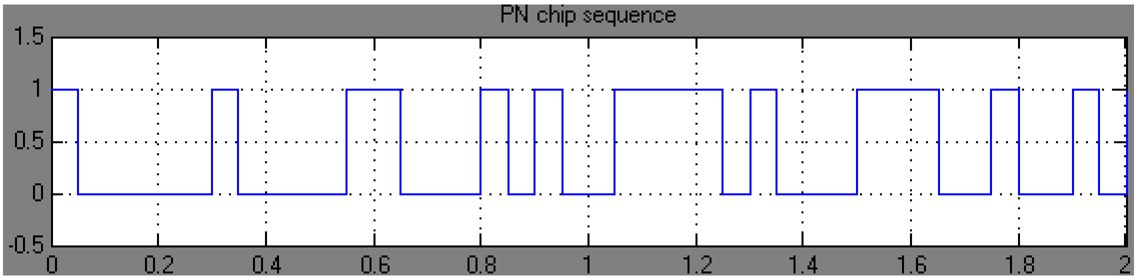

Figure 6.2: Output of the pseudo random generator.

Direct Sequence Spread signal is generated by converting the input bit stream and PN bit sequence into NRZ form and multiply the resultant data. This is shown in Figure 6.3

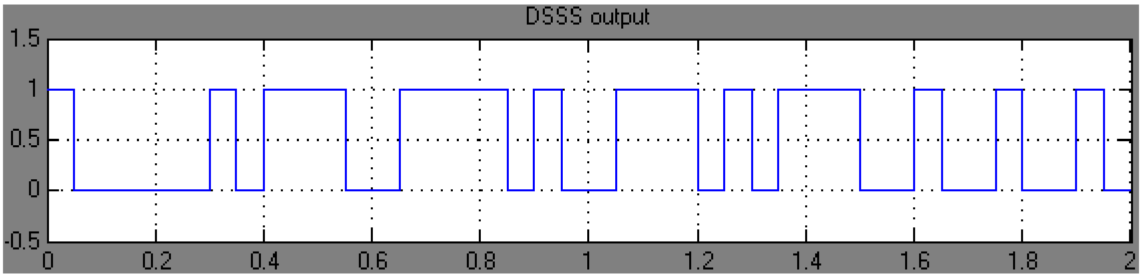

Figure 6.3: Output of the Direct spread spectrum

The clock signal used for the flip-flops is shown in the Figure 6.4. From this clock, two clocks are generated using T flip-flop. These two clocks are complement to each other. The period is double to that of original clock signal. These are called as inphase clock and Quadrature clock. These clock signals are presented in Figure 6.5 and Figure 6.6 respectively.

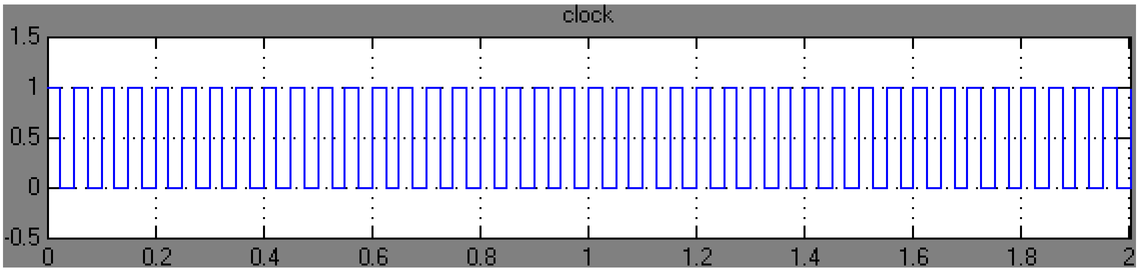

Figure 6.4: Clock

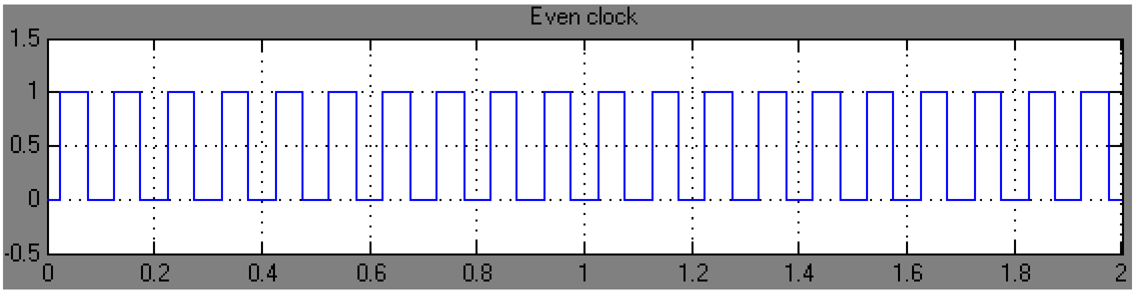

Figure 6.5: Inphase clock

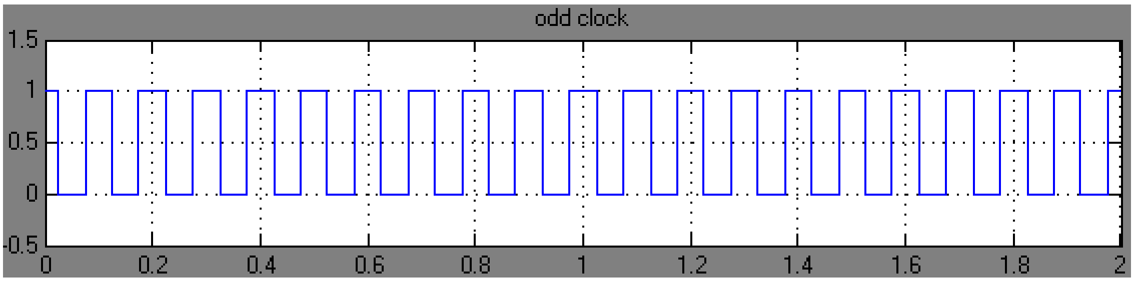

Figure 6.6: Quadrature clock

The DSSS signal has a data rate of 2Mbps. The inphase data and Quadrature data after serial to parallel conversion are shown in Figure 6.7 and Figure 6.8. They are having a data rate of 1Mbps. By doing this, we can transmit two bits at a time and hence bandwidth efficiency improved.

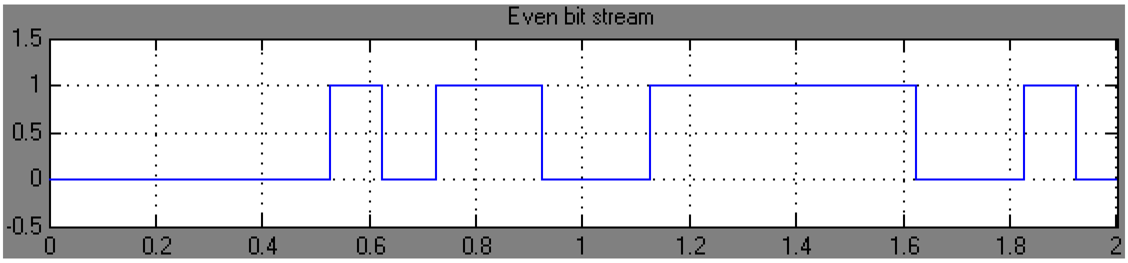

Figure 6.7: Inphase Data

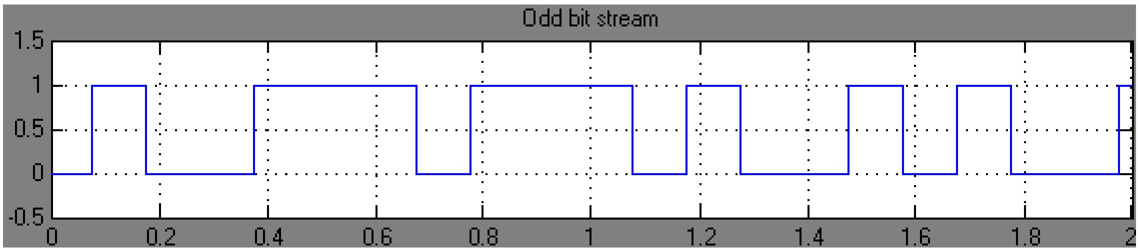

Figure 6.8: Quadrature Data

Following the separation of two bit streams, the two signals are pulse shaped through a half sine wave. Here this sine wave has a time period of 2μs. These Inphase and Quadrature signals after half sine pulse shaping are shown in Figure 6.9 and Figure 6.10 respectively.

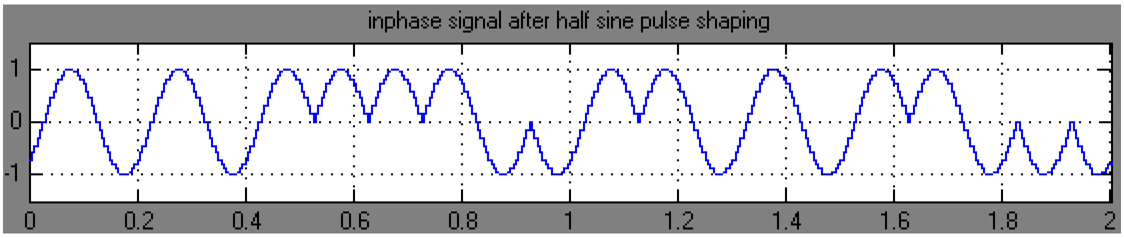

Figure 6.9: Inphase signal after half sine pulse shaping.

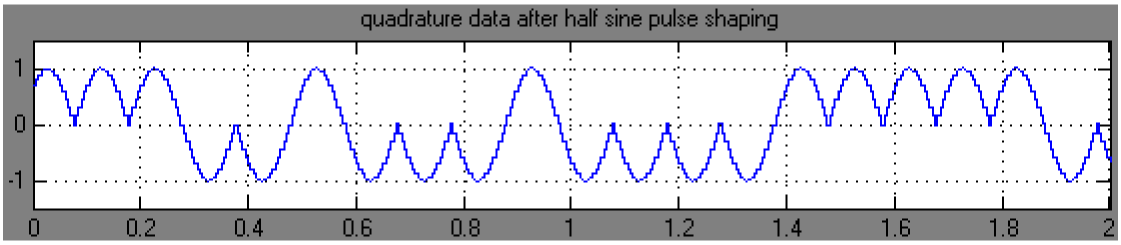

Figure 6.10: Quadrature signal after half sine pulse shaping

Following the pulse shaping, the two signals are multiplied with a 2.4GHz high frequency carrier signal for doing modulation. The inphase signal and Quadrature signal after modulation are shown in Figure 6.11 and Figure 6.12.

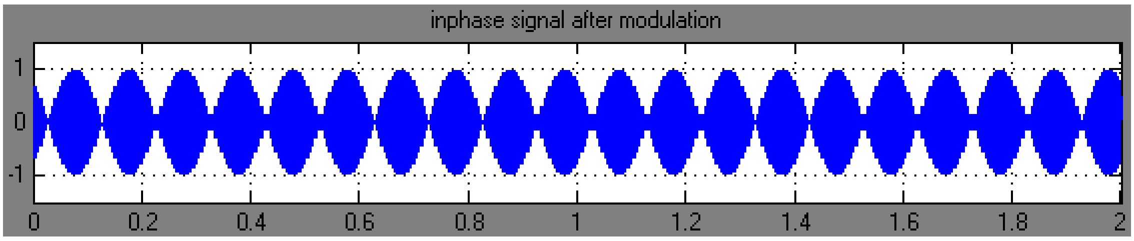

Figure 6.11: Inphase signal after modulation

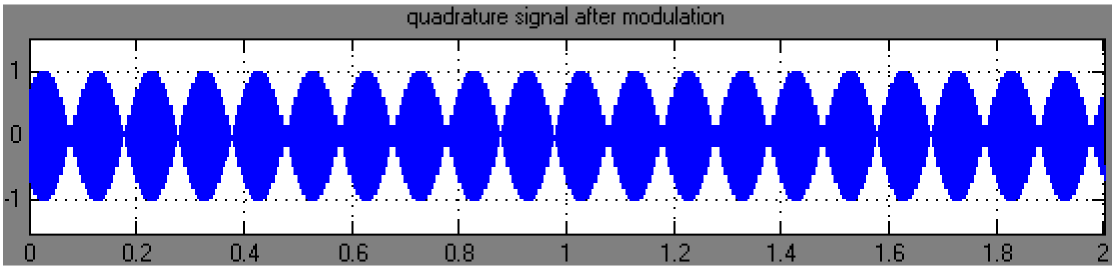

Figure 6.12: Quadrature signal after modulation

Human eye cannot predict the changes in the inphase and quadrature signal. By reducing the sample period, we can observe the changes in the signal and are shown in the Figure 6.13 & Figure 6.14 below and also the final transmitted output of ZigBee transmitter is shown in Figure 6.15.

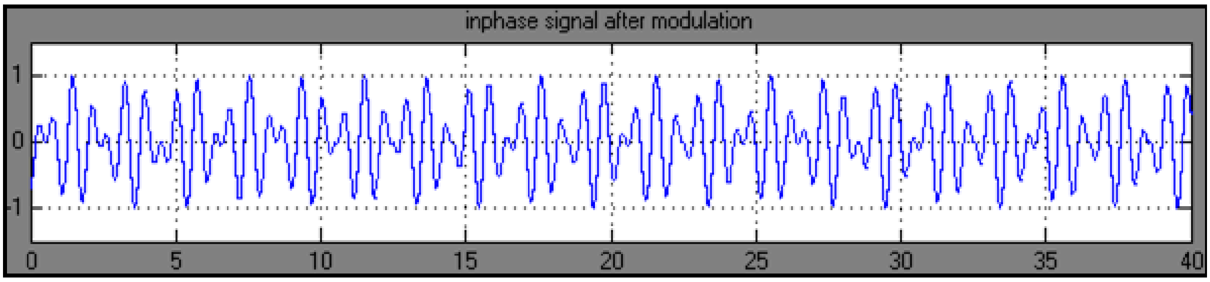

Figure 6.13: Inphase siganl after modulation.

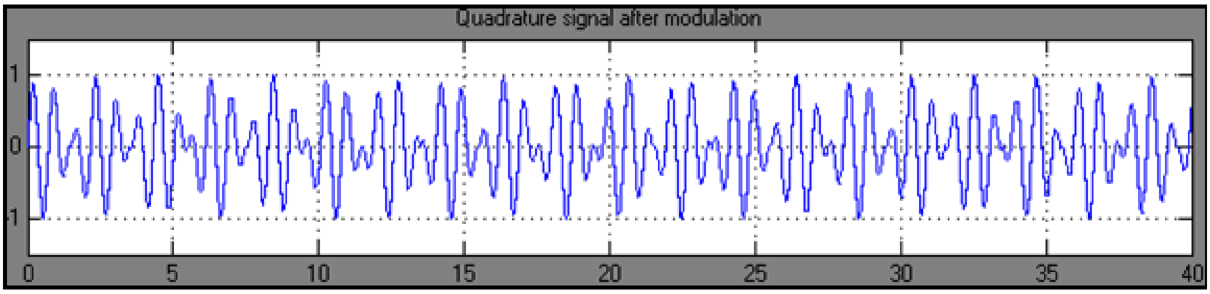

Figure 6.14: Quadrature signal after modulation.

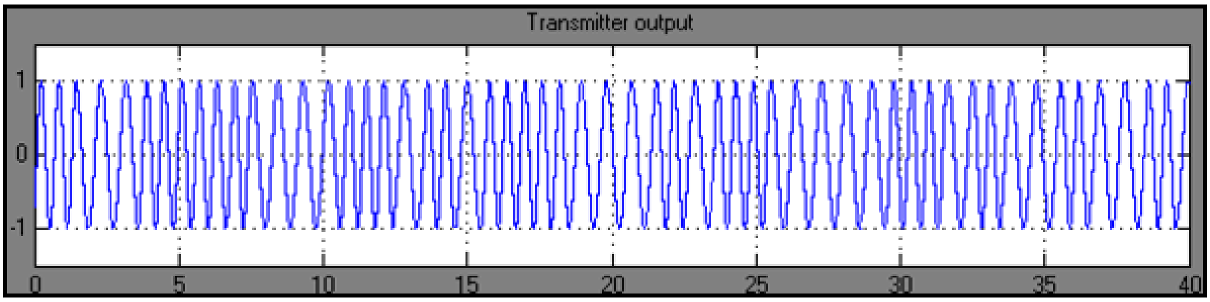

Figure 6.15: Output of ZigBee Transmitter.

Clearly , the results shows that there is no phase transitions in the output by using OQPSK with half sine pulse shaping. By taking this as advantage, an efficient power amplifier design is possible in the realization of hardware.

6.1.2 At the receiver end

By observing these figures, we can know easily what happens exactly inside the ZigBee Receiver.

The transmitted signal is passed through a AWGN channel. The noisy version of the transmitted signal at the input of receiver is shown in Figure 6.16.

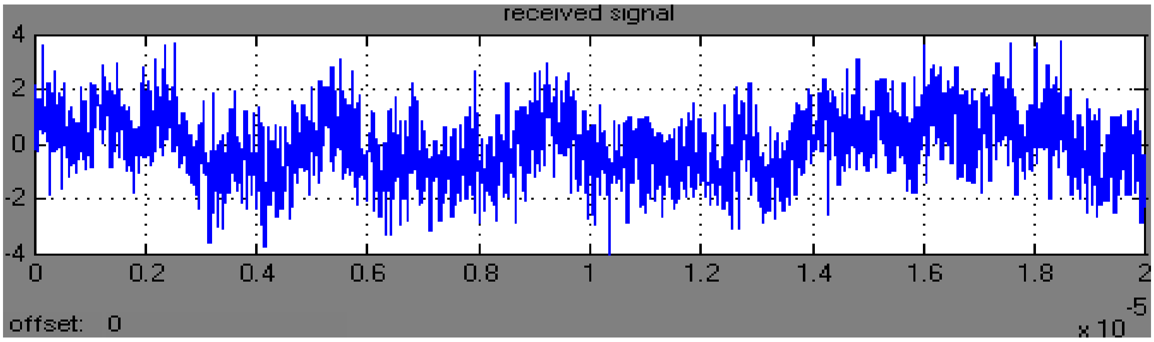

Figure 6.16: Received signal after passing through channel.

The outputs of both inphase and Quadrature signals after multiplying with recovered carrier is shown in Figure 6.17 and Figure 6.18 respectively.

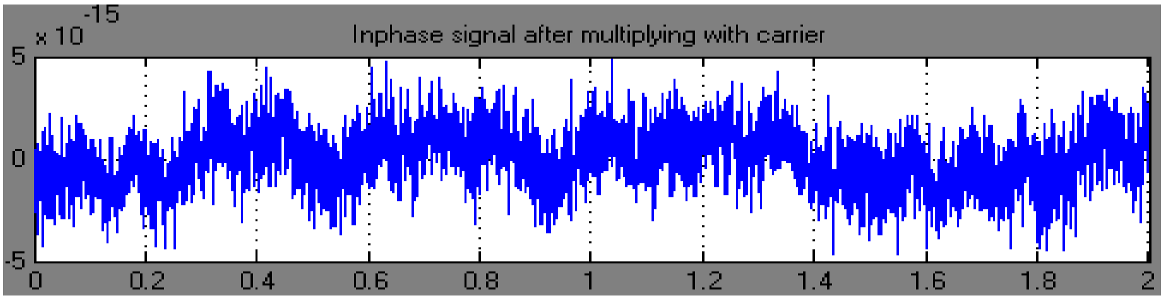

Figure 6.17: Inphase signal after multiplying with carrier.

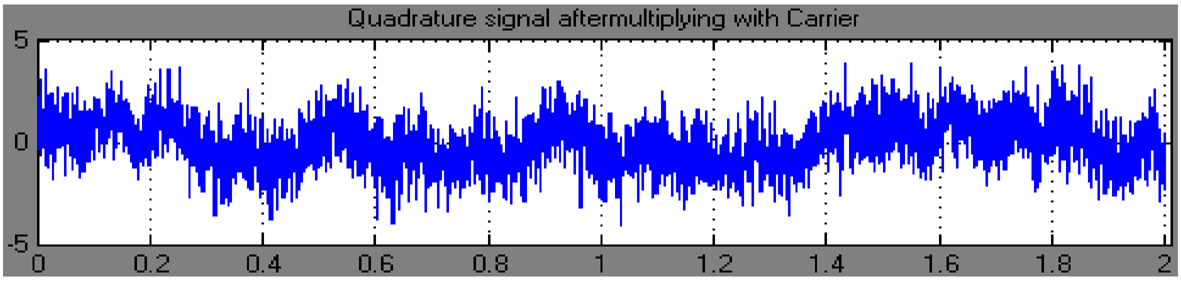

Figure 6.18: Quadrature signal after multiplying with carrier.

The outputs of both inphase and Quadrature signals after multiplying with half sine signal are shown in Figure 6.19 and Figure 6.20 respectively.

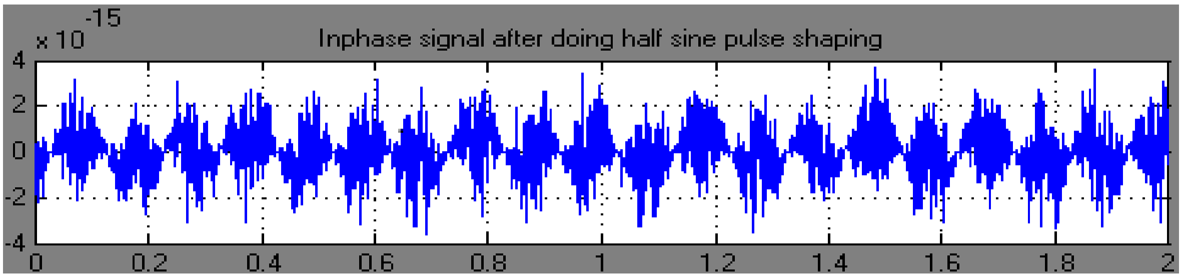

Figure 6.19: Inphase signal after multiplying with half sine pulse shaping.

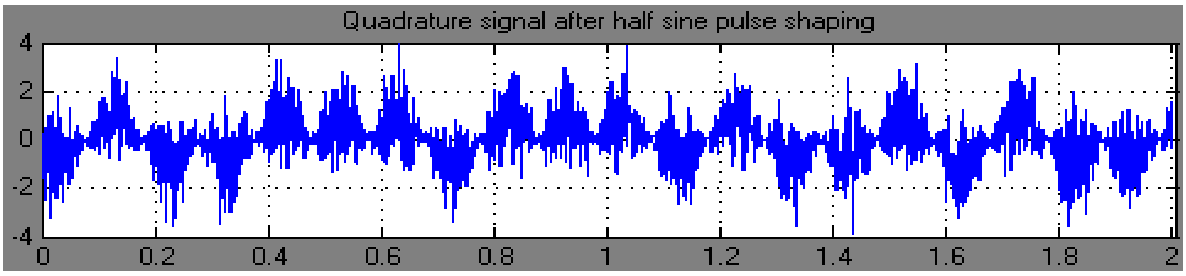

Figure 6.20: Quadrature signal after multiplying with half sine pulse shaping.

The resultant signals containing both the high frequency harmonics and baseband signal components. Separate the baseband signal components form high frequency harmonics by passing through a low pass filter. These signals are passed through a 3rd order Butterworth low pass filter having cutoff frequency of 500 KHz for extracting only baseband data. Outputs after low pass filtering are shown in Figure 6.21 and Figure 6.22.

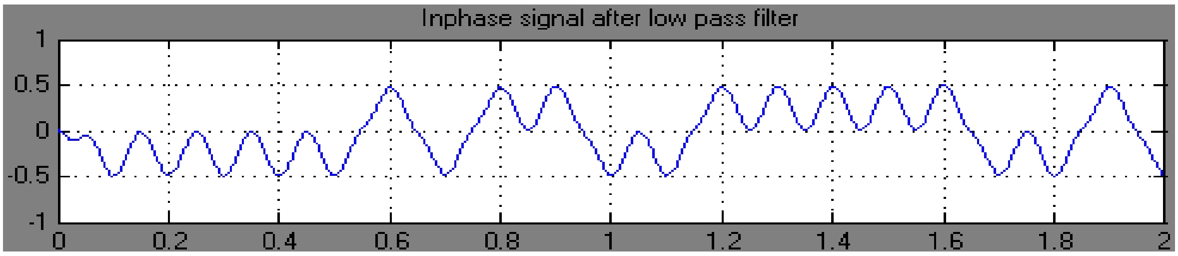

Figure 6.21: Inphase signal after Low pass filtering.

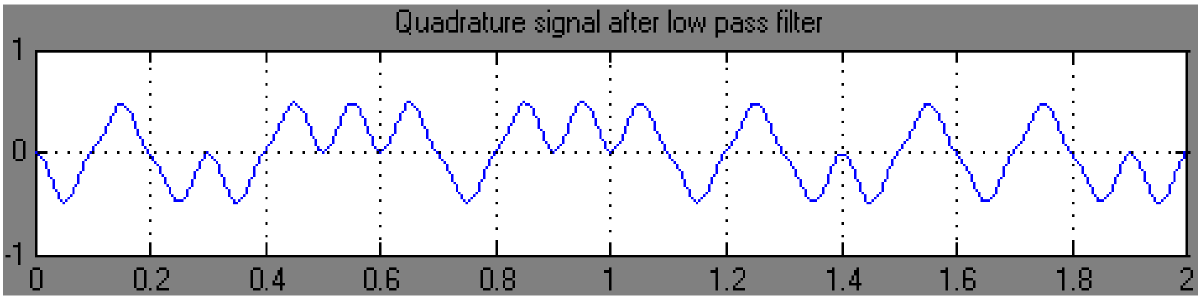

Figure 6.22: quadrature signal after low pass filtering.

Next, this baseband output is passed through a sample and hold circuit for sampling of base band signal. The sample period used for sampling is 2µs. This sampled output at both inphase and Quadrature demodulators are shown in Figure 6.23 and Figure 6.24 respectively.

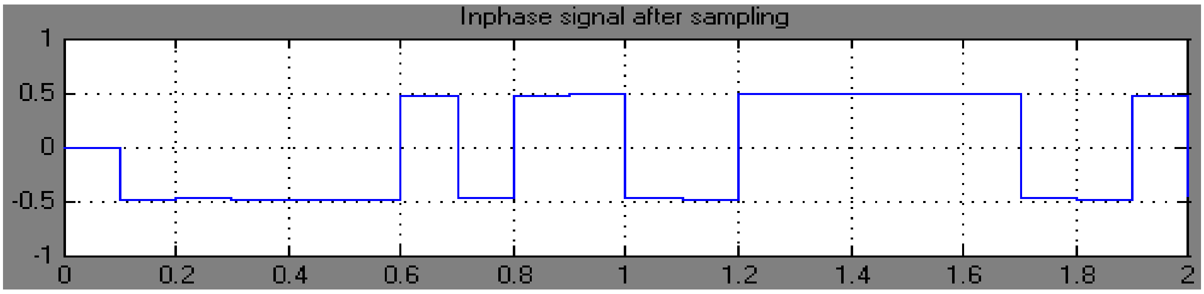

Figure 6.23: Inphase signal after sampling.

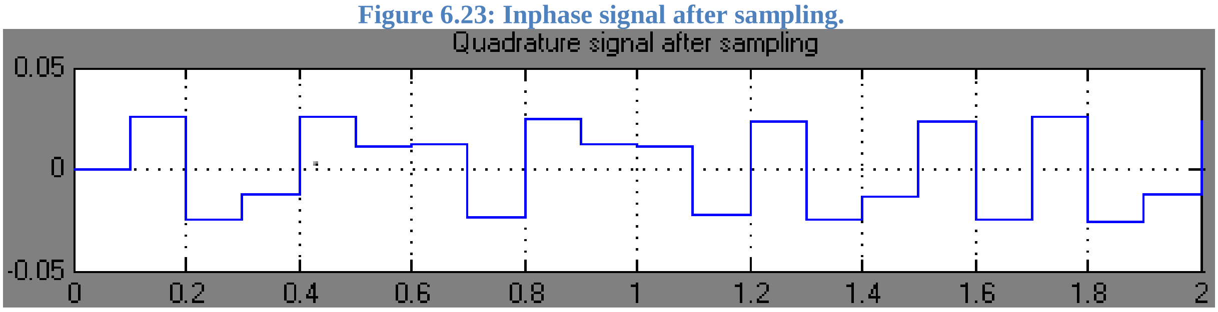

Figure 6.24: Quadrature signal after sampling.

This sampled data is passed through a comparator, for deciding whether the transmitted bit is '1' or '0'. Set the threshold for comparator as '0'. The comparator output at both inphase and Quadrature demodulators are shown in Figure 6.25 and Figure 6.26 respectively.

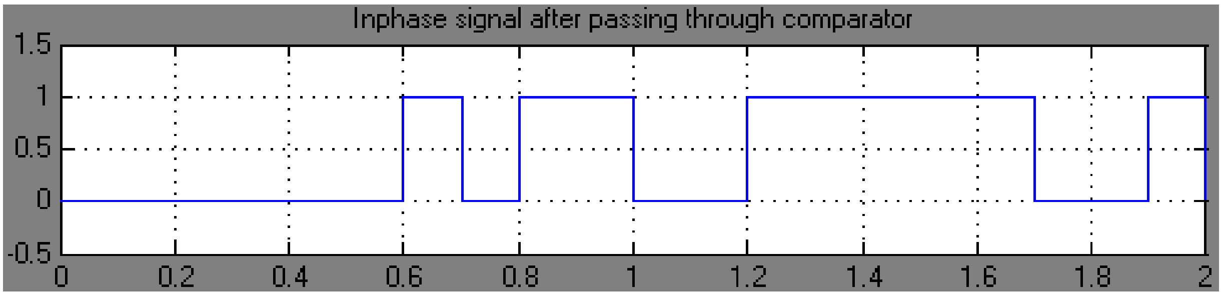

Figure 6.25: Inphase signal after passing through comparator.

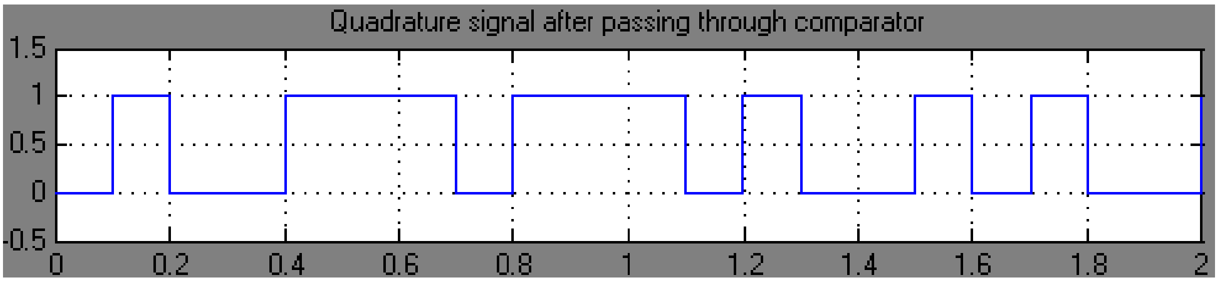

Figure 6.26: Quadrature signal after passing through comparator.

Next, introduce a delay of one bit period at the Quadrature demodulator output. If the delay introduced in the transmitter at Quadrature side, then introduce the same amount of delay in the receiver inphase stage and vice versa. The delayed version Quadrature data is shown in the Figure 6.27.

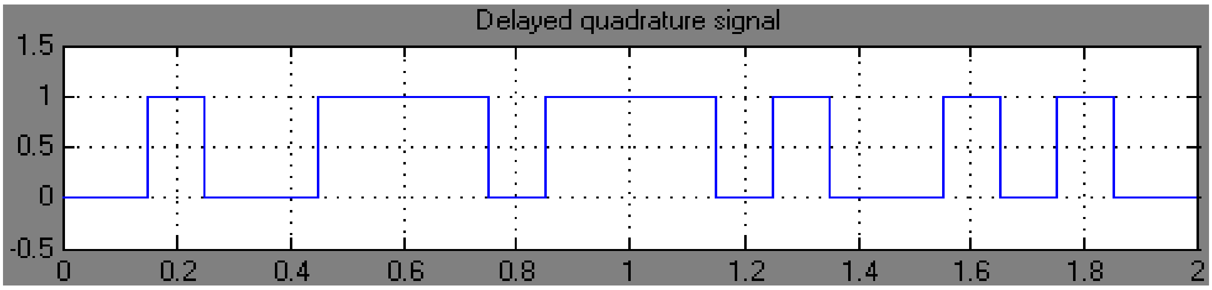

Figure 6.27: Delayed Quadrature signal.

This resulting inphase and Quadrature data is multiplexed using a multiplexer or a parallel to serial to converter and the output after passing through a parallel to serial converter is given in Figure 6.28.

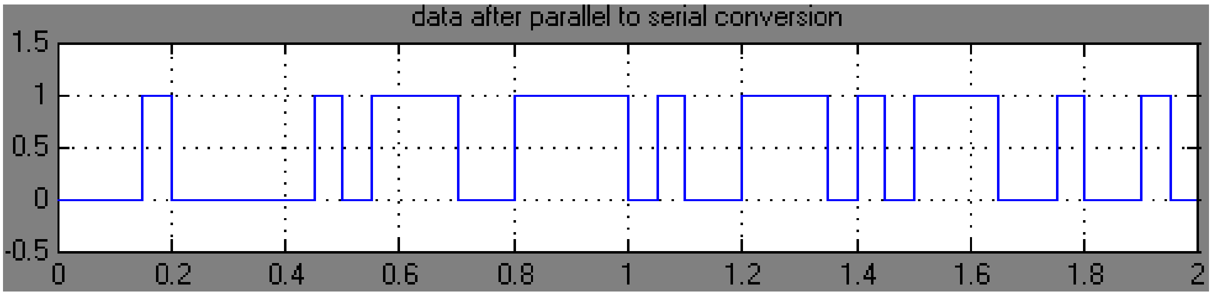

Figure 6.28: Data after parallel to serial conversion.

This output after multiplying with shifted PN sequence data(shown in Figure 6.29) is given in Figure 6.30 below.

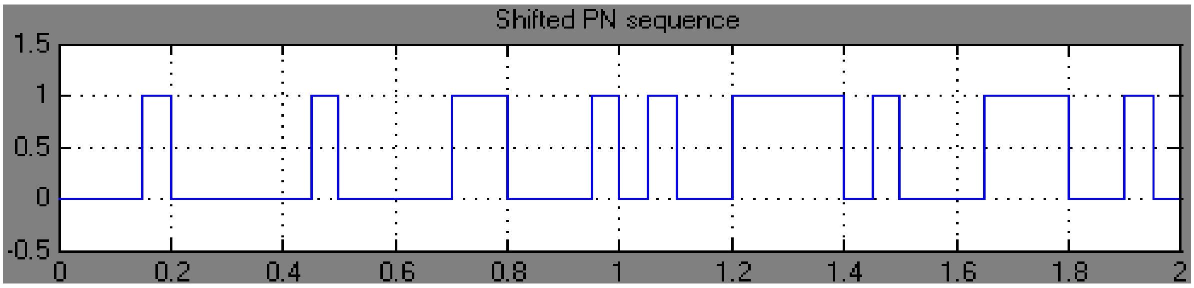

Figure 6.29: Shifted PN sequence.

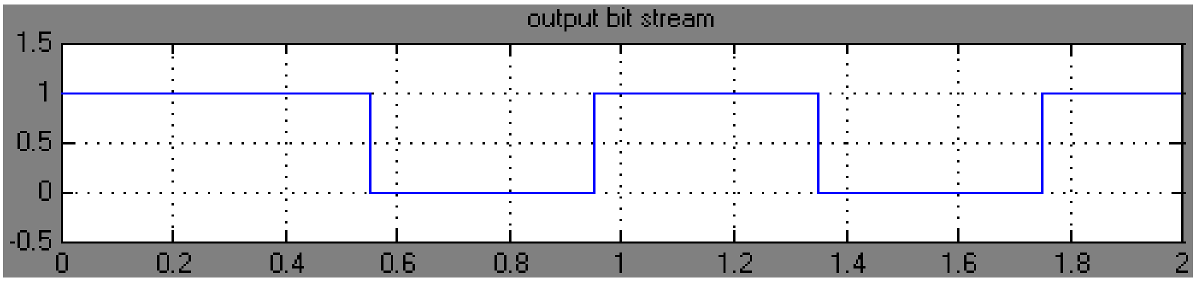

Figure 6.30: Output bit Stream.

The output bit stream shown in Figure 6.30 and the input bit stream shown in Figure 6.1 are same except a small amount of delay (1.5μs).

CHAPTER 7

CONCLUSIONS & SCOPE FOR FUTURE WORK

CHAPTER 7

CONCLUSION & SCOPE FOR FUTURE WORK

7.1 CONCLUSION:

` The work presented here helps to implement a transceiver for ZigBee wireless communication system using Matlab/Simulink. Without using mathematically complex blocks, we designed and tested a ZigBee wireless transceiver in Matlab/Simulink. In this thesis previous work of [8] & [15] are used as reference to implement ZigBee transceiver. A model for MSK has been presented, an analysis of which shows that the theoretical maximum bandwidth efficiency of MSK is 2 bits/s/Hz, the same as for QPSK and Offset QPSK.Here, we are indirectly implementing Minimum Shift Keying modulation and demodulation (OQPSK with half sine pulse shaping). As discussed in chapter 4, Half sine pulse shaping avoids the abrupt phase shifts in the transmitted signal so that it reduced lot of burden and the modulated signal is amplifier friendly in real time scenario. Use of direct spread spectrum technique, reduces the interference effects. As discussed in chapter4, out of the three commonly used radio transceivers (superhetrodyne, Low IF and direct conversion transceiver), the use of direct conversion receiver fulfils the requirement of ZigBee i.e., low cost and low power consumption.

7.2 SCOPE FOR FUTURE WORK

ZigBee transceiver was working successfully only in testing domain in Simulink. By using Verilog HDL, we can establish link from Matlab/Simulink to Cadence and hence we can verify this tested work in real time scenario also. This can be achieved with the help of co simulation blocks in Simulink library.

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