BLOCK TURBO CODE AND ITS APPLICATION TO OFDM FOR WIRELESS LOCAL AREA NETWORK

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

Master of Technology
In
Electronic systems and Communication

By
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Department of Electrical Engineering
National Institute of Technology
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CERTIFICATE

This is to certify that the thesis entitled, " **Block Turbo Code and its application to OFDM for wireless local area network** " submitted by Mr**. Hrudananda Pradhan** in partial fulfillment of the requirements for the award of Master of Technology Degree in Electrical Engineering with specialization in " **Electronic systems and Communiacation** " 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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Last but not least I would like to express my gratitude to my parents, whose love and encouragement have supported me throughout my education.

Hrudananda Pradhan

Abstract

To overcome multipath fading and Inter symbol Interference (ISI), in convolutional single carrier systems equalizers are used. But it increases the system complexity. Another approach is to use a multicarrier modulation technique such as OFDM, where the data stream to be transmitted is divided into several lower rate data streams each being modulated on a subcarrier. To avoid ISI, a small interval, known as the guard time interval, is inserted into OFDM symbols. The length of the guard time interval is chosen to exceed the channel delay spread. Therefore, OFDM can combat the multipath fading and eliminate ISI almost completely.

The another problem is the reduction of the error rate in transmitting digital data. For that we use error correcting Codes in the design of digital transmission systems. Turbo Codes have been widely considered to be the most powerful error control code of practical importance.

Turbo codes can be achieved by serial or parallel concatenation of two (or more) codes called the constituent codes. The constituent codes can be either block codes or convolutional codes. Currently, most of the work on turbo codes have essentially focused on Convolutional Turbo Code (CTC)s and Block Turbo Code (BTC)s have been partially neglected. Yet, the BTC solution is more attractive for a wide range of applications. In this paper, Block Turbo Codes or Turbo Product Codes are used which is similar to the IEEE 802.11a WLAN standard.

In this thesis work simple explanation of BTCOFDM theory is given. The BER performance is evaluated for the Block Turbo coded BPSK and QPSK OFDM system, under both AWGN channel and Rayleigh fading channel. It also compares the BER performance of Block Turbo coded OFDM with the uncoded OFDM. It is verified in the present work that the BTCOFDM system with 4 iterations is sufficient to provide a good BER performance. Additional number of iterations does not show noticeable difference. The simulation results shows that the BTCOFDM system achieves large coding gain with lower BER performance and reduced decoding iterations, therefore offering higher data rate in wireless mobile communications

Contents

	Page
Abstract	i
List of Figures	v
List of Tables	vi
Abbreviations used	vii
Chapter 1: Introduction	1
1.1 Introduction	2
1.2 Motivation	2
1.3 Background literature survey	3
1.4 Thesis outlines	4
Chapter 2: Modulation Techniques in OFDM	5
2.1 Introduction	6
2.2 Digital modulation techniques	6
2.2.1 Three Basic concepts of modulation	6
2.2.2 ASK	7
2.2.3 FSK	7
2.2.4 PSK	8
2.2.5 BPSK	8
2.2.6 QPSK	9
2.2.7 QAM	10
2.3 Bit rate and symbol rate	11
Chapter 3: Block Turbo Code	14
3.1 Introduction	15
3.2 Concatenated codes	16

3.3 Block Turbo Code 17
3.4 Soft decoding of linear block codes 19
3.5 Reliability of decision *D* given by soft-input decoder 20
3.6 Computing the soft decision at the output of the soft-input decoder 23
3.7 Turbo decoding of product codes 24

Chapter 4: Coded orthogonal frequency division multiplexing 26

4.1 Introduction 27
4.2 Frequency division multiplexing modulation 27
4.3 Orthogonal frequency division multiplexing (OFDM) 28
4.4 Coded orthogonal frequency division multiplexing 31
4.5 COFDM transmission & reception 32
4.5.1 Forward error control & Interleaving 33
4.5.2 Symbol Mapping 34
4.5.3 Serial to parallel converter 34
4.5.4 FFT and IFFT 35
4.5.5 Parallel to serial converter 36
4.5.6 Cyclic prefix 36
4.5.7 Receiver 38
4.6 Advantages of COFDM 38
4.6.1 Efficient Modulation and Demodulation 38
4.6.2 High Spectral Efficiency 38
4.6.3 Multipath Delay Spread Tolerance 39
4.6.4 Immunity to Frequency selective fading Channels 39
4.7 Applications of COFDM 39
4.7.1 Digital Audio Broadcasting (DAB) 39
4.7.2 Digital Video Broadcasting (DVB) 40
4.7.3 Wireless LAN Applications 40

Chapter 5: communication channels 41

5.1 Introduction 42

5.2 Additive white gaussian noise (AWGN) 42
5.3 Multipath 43
5.4 Fading 45
5.5 Rayleigh fading channel 46
5.6 Multipath Fading 49
5.6.1Multipath Channel Characteristics 50

Chapter 6: Simulation results and discussion 51

6.1 Introduction 52
6.1.1Matlab simulation model 52
6.1.2Simulation parameters 53
6.2 BER performance of OFDM for different modulations over AWGN channel 53
6.3 BER performance of Block Turbo code under AWGN channel for different iterations 55
6.4 Performance of BTC – OFDM over AWGN and Rayleigh fading channel 57
6.5 Comparison between BTC - OFDM and Convolutional coded OFDM 59
6.6 Performance of BTC – OFDM over multipath fading channel 60
6.7 Effect of Cyclic Prefix length in BTC – OFDM over a multipath fading channel 62

Chapter 7: Conclusion 64

7.1 Conclusion 65
7.2 Scope of future work 65

References 66

List of figures

	Page
Fig 2.1 Baseband information sequence 0010110010	7
Fig 2.2 Binary ASK (OOK) signal	7
Fig 2.3 Binary FSK signal	7
Fig 2.4 Binary PSK Carrier (Note the 180°phase shifts at bit edges)	8
Fig 2.5 (a) Binary modulating signal, and (b) BPSK signal.	9
Fig 2.6 QPSK modulation: (a) binary sequence and (b) QPSK signal.	10
Fig 3.1 Principle of Concatenated codes	16
Fig 3.2 Concatenated encoder and decoder with interleaver	17
Fig 3.3 Construction of product code $\mathrm{P} = c^1 \otimes c^2$	18
Fig 3.4 Block diagram of elementary block turbo decoder	25
Fig 4.1 (a) A Regular-FDM single carrier (b) Orthogonal-FDM	29
Fig 4.2 Frequency Efficiency of OFDM over FDM for Two and Three Subchannels	29
Fig 4.3 freq. spectrum of OFDM	30
Fig 4.4 waveform of carriers in OFDM	30
Fig 4.5 COFDM system diagram	33
Fig 4.6 Block diagram of FFT and IFFT system	35
Fig 4.7 Addition of Guard Period to an OFDM Signal	37
Fig 5.1 Received signal corrupted by AWGN	43
Fig 5.2 Principle of multipath channel	44
Fig 5.3 Delayed wave with incident angle	47
Fig 5.4 Effect of multipath on a mobile station	50
Fig 6.1 Turbo coded OFDM model used for simulation	52
Fig 6.2 BER comparison of uncoded OFDM for different modulations	54
Fig 6.3 Comparison of different iterations for Turbo code under AWGN channel	56
Fig 6.4 Performance of BTC – OFDM over AWGN and Rayleigh fading channel.	58
Fig 6.5 Comparison between Block Turbo coded OFDM and Convolutional coded OFDM	59
Fig 6.6 Performance of BTC – OFDM over multipath fading channel	61
Fig 6.7 Effect of Cyclic Prefix length in BTC – OFDM over a multipath fading channel	62

List of tables

Page

Table 4.1 Parameters of IEEE 802.11 40

Table 6.1 BTC – OFDM Simulation parameters 53

Table 6.2 Simulation parameters of uncoded OFDM
for different modulations over AWGN 54

Table 6.3 SNR comparison of uncoded OFDM under
AWGN channel 55

Table 6.4 simulation parameters of Turbo code under
AWGN channel 56

Table 6.5 Simulation parameters of BTC – OFDM for AWGN
and Raleigh fading channel with Doppler shift f_d = 100 Hz. 57

Table 6.6 Eb/N0 Performance of BTC – OFDM over
AWGN and Rayleigh fading channel. 58

Table 6.7 SNR comparison between BTC – OFDM
and Convolutional coded OFDM 60

Table 6.8 Multipath channel model parameters 60

Table 6.9 Performance of BTC – OFDM over
multipath fading channel 61

Table 6.10 Eb/N0 performance of effect of cyclic prefix
length in BTC – OFDM over a multipath fading channel. 63

Abbreviations used

FDM ---Frequency Division Multiplexing

OFDM ---Orthogonal Frequency Division Multiplexing

RS coding---Reed Solomon coding

WLAN---wireless Local Area Network

DAB ---Digital Audio Broadcasting

DVB ---Digital Video Broadcasting

DFT ---Discrete Fourier Transform

HDSL ---High bit-rate Digital Subscriber Line

ADSL ---Asymmetric Digital Subscriber Line

VDSL ---Very High speed Digital Subscriber Line

HDTV --High Definition Television

SCM --Single Carrier Modulation

FFT ---Fast Fourier Transform

QAM ---Quadrature Division Multiplexing

UHF --Ultra High Frequencies

BTC --Block Turbo Code

BTCOFDM ---Block Turbo Coded OFDM

CDMA ---Code Division Multiple Access

TDMA ---Time Division Multiple Access

ASK --Amplitude Shift Keying

FSK --Frequency Shift Keying

PSK --Phase Sift Keying

BPSK --Binary Phase Shift Keying

QPSK --Qadrature Phase Shift Keying

CP --Cyclic Prefix

FEC --Forward Error Correction

SNR --Signal to Noise Ratio

SER --Symbol Error Rate

BER --Bit Error Rate

Chapter 1

Introduction

1.1 Introduction :

Wireless technologies are the veritable explosions in telecommunication industries. Once exclusively military, satellite and cellular technologies are now commercially driven by ever more demanding consumers, who are ready for seamless communication from their home to their car, to their office, or even for outdoor activities. With this increased demand comes a growing need to transmit information wirelessly, quickly and accurately. To address this need, communications engineer have combined technologies suitable for high rate transmission with forward error correction (FEC) techniques. Orthogonal Frequency Division Multiplexing (OFDM) is the standard being used throughout the world to achieve the high data rates necessary for data intensive applications that must now become routine. A particularly attractive feature of OFDM systems is that they are capable of operating without a classic equalizer, when communicating over depressive transmission media, such as wireless channels, while conveniently accommodation the time- and frequency-domain channel quality fluctuations of the wireless channel.

1.2 Motivation :

Multiple access techniques which are quite developed for the single carrier modulations (e.g. TDMA, FDMA) had made possible of sharing one communication medium by multiple number of users. Multiple techniques schemes are used to allow many mobile users to share simultaneously a finite amount of radio spectrum. The sharing is required to achieve high capacity by simultaneously allocating the available bandwidth (or the available amount of channels) to multiple users.

For the quality communications, this must be done without severe degradation in the performance of the system. FDMA, TDMA and CDMA are the well known multiplexing techniques used in wireless communication systems. While working with the wireless systems using these techniques various problems encountered are (1) multi-path fading (2) time dispersion which lead to inter symbol interference (ISI) (3) lower bit rate capacity (4) requirement of larger transmit power for high bit rate and (5) less spectral efficiency. Disadvantage of FDMA technique is its Bad Spectrum Usage. Disadvantages of TDMA technique is Multipath Delay spread problem. In a typical terrestrial broadcasting, the transmitted signal arrives at the receiver using various paths of different lengths. Since multiple versions of the signal interfere with each other, it becomes difficult to extract the original information.

To overcome multipath fading and Inter symbol Interference (ISI), in convolutional single carrier systems equalizers are used. But it increases the system complexity. Another approach is to use a multicarrier modulation technique such as OFDM, where the data stream to be transmitted is divided into several lower rate data streams each being modulated on a subcarrier. To avoid ISI, a small interval, known as the guard time interval, is inserted into OFDM symbols. The length of the guard time interval is chosen to exceed the channel delay spread. Therefore, OFDM can combat the multipath fading and eliminate ISI almost completely and provides better solution for the above mentioned problems.

1.3 Background literature survey :

The concept of using parallel data transmission by means of frequency division multiplexing (FDM) was published in mid 60's [1]. Some early development with this can be traced back to the 50s [1]. A U.S. patent was filled and issued in January 1970 [2]. The idea was to use parallel data streams and FDM with overlapping sub channels to avoid the use of high-speed equalization and to combat impulsive noise, and multipath distortion as well as to fully use the available bandwidth. The initial applications were in the military communications. Weinstein and Ebert [3] applied the discrete Fourier transform (DFT) to parallel data transmission system as part of the modulation and demodulation process. In the 1980s, OFDM has been studied for highspeed modems [4], digital mobile communications [5] and high-density recording [6].

In 1990s, OFDM has has found its applications in wideband data communications over mobile radio FM channels [9], wireless LAN [8], wireless multimedia communication [10], high-bit-rate digital subscriber lines (HDSL) [11], asymmetric digital subscriber lines (ADSL) [7], digital audio broadcasting (DAB) [12], digital video broadcasting (DVB) [17]. OFDM has been chosen as the modulation technique for the new 5 GHz IEEE802.lla [13] standard as well as High-Performance LAN (HIPERLAN) [14], [15].

For the reduction of the error rate in transmitting digital data we use error correcting Codes in the design of digital transmission systems. Turbo Codes proposed by Berrou in 1993 [16] have been widely considered to be the most powerful error control code of practical importance. Turbo codes have error correcting capability very close to the theoretical performance limits.

1.4 Thesis outlines :

After conveying our motivation and discussing the background literature survey of OFDM and Turbo Coding in this *Introductory* chapter we tried our best to organize this thesis in very systematic way which includes these chapters –

Chapter 2 : This chapter includes the fundamentals behind digital modulation used in OFDM and the also the bit rate and symbol rate calculation.

Chapter 3 : The principle of concatenated codes, Turbo encoders and decoders are included in this section of the thesis.

Chapter 4 : In this chapter of the thesis we more concentrate on the subject matter which is Coded Orthogonal Frequency Division Multiplexing (COFDM). Here we also discuss the applications of COFDM and its advantages.

Chapter 5 : Here we will focus on communication channels that exists in wireless communication and how the communication channels contribute in the BER performance of COFDM.

Chapter 6 : This section is having discussions and analysis on the simulation results.

Chapter 7 : Conclusion symbolizes the whole work & so this section is having some key words, which reflect my labour for this project work under the guidance of my guide.

Last but not the least, *References* give completeness to my thesis and my project work.

Chapter 2

Modulation Techniques in OFDM

2.1 Introduction :

Modulation is the process of facilitating the transfer of information over a medium. Sound transmission in air has limited range for the amount of power our lungs can generate. To extend the range our voice can reach, we need to transmit it through a medium other than air, such as a phone line or radio. The process of converting information (voice in this case) so that it can be successfully sent through a medium (wire or radio waves) is called modulation.

There are two types of modulations: Analog modulation and digital modulation. In analog modulation, an information-bearing analog waveform is impressed on the carrier signal for transmission whereas in digital modulation, an information-bearing discrete-time symbol sequence (digital signal) is converted or impressed onto a continuous-time carrier waveform for transmission. 2G wireless systems are realized using digital modulation schemes.

2.2 Digital modulation techniques :

2.2.1 Three Basic concepts of modulation :

There are three basic types of digital modulation techniques. These are

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

All of these techniques vary a parameter of a sinusoid to represent the information which we wish to send. A general carrier wave may be written:

$$c(t) = A\sin(2\pi f t + \theta) \tag{2.1}$$

A sinusoid has three different parameters than can be varied. These are its amplitude, phase and frequency. Modulation is a process of mapping such that it takes your voice (as an example of a signal) converts it into some aspect of a sine wave and then transmits the sine wave, leaving the actual voice behind. The sine wave on the other side is remapped back to a near copy of your sound.

2.2.2 ASK :

In ASK, the amplitude of the carrier is changed in response to information and all else is kept fixed. In Binary ASK Bit 1 is transmitted by a carrier of one particular amplitude. To transmit 0, we change the amplitude keeping the frequency constant. On-Off Keying (OOK) is a special form of ASK, where one of the amplitudes is zero as shown in fig 2.1 and fig 2.2.

$$\text{Binary } ASK(t) = s(t)\sin(2\pi f t) \tag{2.2}$$

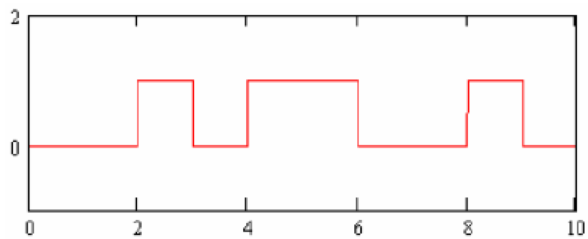


Fig 2.1 Baseband information sequence 0010110010

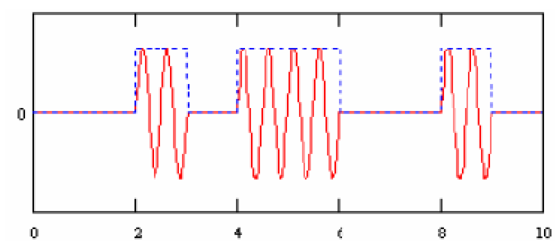


Fig 2.2 Binary ASK (OOK) signal

2.2.3 FSK :

In FSK, we change the frequency in response to information, In Binary FSK one particular frequency for a 1 and another frequency for a 0 is used as shown in fig 2.3 for the same bit sequence as above. In the example below, frequency f1 for bit 1 is higher than f2 used for the 0 bit.

$$\begin{aligned}\text{Binary } FSK(t) &= \sin(2\pi f_1 t) \quad \text{for bit 1}\\ &= \sin(2\pi f_2 t) \quad \text{for bit 0}\end{aligned} \tag{2.3}$$

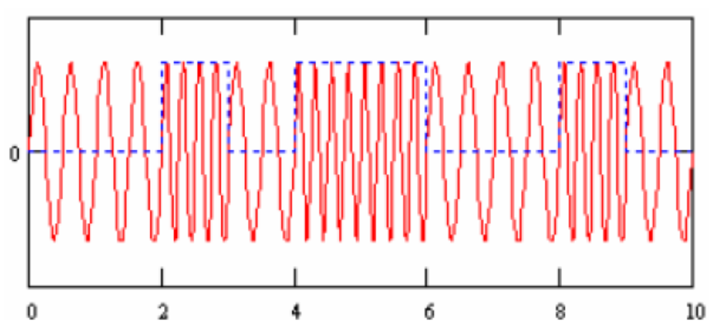


Fig 2.3 Binary FSK signal

2.2.4 PSK :

In PSK, we change the phase of the sinusoidal carrier to indicate information. Phase in this context is the starting angle at which the sinusoid starts. For Binary PSK It has one fixed phase usually 0° when the data is 1. To transmit 0, we shift the phase of the sinusoid by 180°. Phase shift represents the change in the state of the information in this case. ASK techniques are most susceptible to the effects of non-linear devices which compress and distort signal amplitude. The use of phase shift keying produces a constant amplitude signal and was chosen for its simplicity and to reduce problems with amplitude fluctuations due to fading.

Binary $PSK(t) = \sin(2\pi f t)$ for bit 1

$= \sin\,(2\pi f t + \pi)$ for bit 0 (2.4)

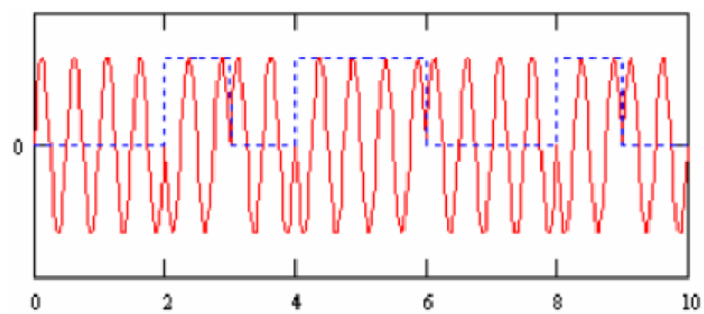


Fig 2.4 Binary PSK Carrier (Note the 180°phase shifts at bit edges)

2.2.5 BPSK :

In binary phase shift keying (BPSK) the transmitted signal is a sinusoid of fixed amplitude. BPSK is the simplest form of PSK. It uses two phases which are separated by 180° and so can also be termed 2-PSK. It has one fixed phase when the data is at one level and when the data is at another level the phase is different by 180°.

In BPSK the data b(t) is a stream of binary digits with voltage levels which, as a matter of convenience, we take to be at +1v and -1v. When b(t) = 1v we say it is at logic level 1 and when b(t) = -1v we say it is at logic level 0. Hence can be written as $V_{BPSK}(t)$ can be written as

$$V_{BPSK}(t) = b(t)\cos\,(\omega_0 t) \qquad (2.5)$$

For binary sequence m(t) = 0 1 0 1 0 0 1, and if the sinusoid s(t) is of amplitude of A, then the resulting BPSK signal will be as shown in the the figure 2.5

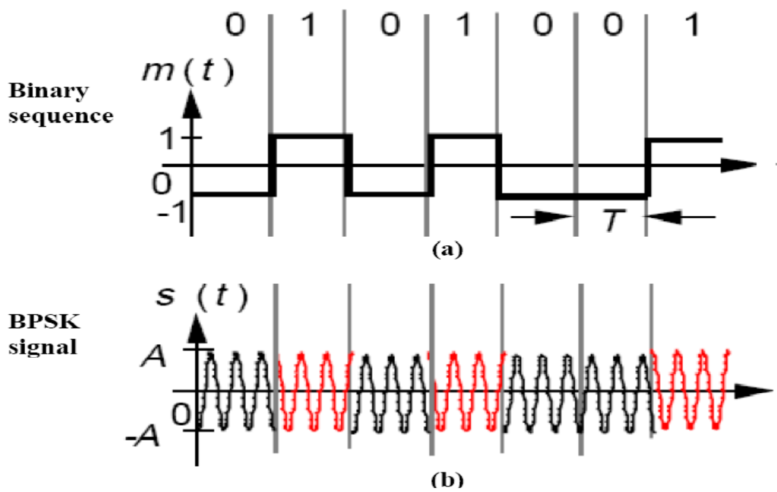
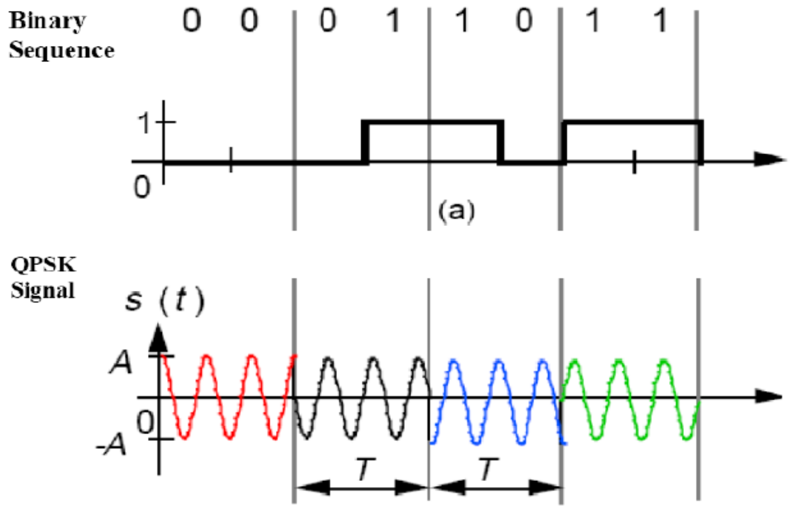


Figure 2.5 (a) Binary modulating signal, and (b) BPSK signal.

2.2.6 QPSK :

QPSK (4-ary PSK) involves changing the phase of the transmitted waveform. Each finite phase change represents unique digital data. A phase-modulated waveform can be generated by using the digital data to change the phase of a signal while its frequency and amplitude stay constant. A QPSK modulated carrier undergoes four distinct changes in phase that are represented as symbols and can take on the values of $\pi/4$, $3\pi/4$, $5\pi/4$, and $7\pi/4$. Each symbol represents two binary bits of data.

For a binary sequence m(t) = 0 0 0 1 1 0 1 1, if the sinusoid s(t) is of amplitude of A, then the resulting QPSK signal will be as shown in the figure 2.6. Phase of the sinusoid is shifted by 90°, 180°, 270°, 360° for data 00, 01, 10, 11 respectively.



(b)

Fig 2.6 QPSK modulation: (a) binary sequence and (b) QPSK signal.

2.2.7 QAM :

ASK is also combined with PSK to create hybrid systems such as *amplitude and phase shift keying* or Quadrature Amplitude Modulation (QAM) where both the amplitude and the phase are changed at the same time. QAM is a modulation scheme which conveys data by changing (*modulating*) the amplitude of two carrier waves. These two waves, usually sinusoids, are out of phase with each other by 90° and are thus called quadrature carriers—hence the name of the scheme.

As for many digital modulation schemes, the constellation diagram is a useful representation. In QAM, the constellation points are usually arranged in a square grid with equal vertical and horizontal spacing, although other configurations are possible. Since in digital telecommunications the data is usually binary, the number of points in the grid is usually a power of 2 (2, 4, 8...). Since QAM is usually square, some of these are rare—the most common forms are 16-QAM, 64-QAM, 128-QAM and 256-QAM. By moving to a higher-order constellation, it is possible to transmit more bits per symbol. However, if the mean energy of the constellation is to remain the same (by way of making a fair comparison), the points must be closer together and are thus more susceptible to noise and other corruption; this results in a higher bit error rate and so higher-order QAM can deliver more data less reliably than lower-order QAM.

Rectangular QAM constellations are, in general, sub-optimal in the sense that they do not maximally space the constellation points for a given energy. However, they have the considerable advantage that they may be easily transmitted as two pulse amplitude modulation (PAM) signals on quadrature carriers, and can be easily demodulated. The non-square constellations achieve marginally better bit-error rate (BER) but are harder to modulate and demodulate. The first rectangular QAM constellation usually encountered is 16-QAM. The reason that 16-QAM is usually the first is that a brief consideration reveals that 2-QAM and 4-QAM are in fact binary phase-shift keying (BPSK) and quadrature phase-shift keying (QPSK), respectively. Also, the error-rate performance of 8-QAM is close to that of 16-QAM (only about 0.5dB better), but its data rate is only three-quarters that of 16-QAM.

2.3 Bit rate and symbol rate :

To understand and compare different PSK modulation format efficiencies, it is important to first understand the difference between bit rate and symbol rate. The signal bandwidth for the communications channel needed depends on the symbol rate, not on the bit rate.

Bit rate is the frequency of a system bit stream. Take, for example, a radio with an 8 bit sampler, sampling at 10 kHz for voice. The bit rate, the basic bit stream rate in the radio, would be eight bits multiplied by 10K samples per second, or 80 Kbits per second. (For the moment we will ignore the extra bits required for synchronization, error correction, etc.).

The symbol rate is the bit rate divided by the number of bits that can be transmitted with each symbol.

$$\text{Symbol Rate} = \frac{\text{Bit rate}}{\text{No:of bits transmitted with each symbol}} \tag{2.6}$$

If one bit is transmitted per symbol, as with BPSK, then the symbol rate would be the same as the bit rate of 80 Kbits per second. If two bits are transmitted per symbol, as in QPSK, then the symbol rate would be half of the bit rate or 40 Kbits per second. Symbol rate is sometimes called baud rate. Note that baud rate is not the same as bit rate. These terms are often confused. If more bits can be sent with each symbol, then the same amount of data can be sent in a narrower spectrum. This is why modulation formats that are more complex and use a higher number of states can send the same information over a narrower piece of the RF spectrum.

The bit error rate (BER) of BPSK in AWGN can be calculated as:

$$P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \tag{2.7}$$

Where $No/2$ = noise power spectral density (W/Hz)

$E_b = P_s T_b$ is the energy contained in a bit duration.

Where P_s = power of sinusoid of amplitude A

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

So that $A = \sqrt{2p_s}$

Thus the transmitted signal is either

$$\begin{aligned} V_{BPSK}(t) &= \sqrt{2p_s}\cos(\omega_0 t) \\ &= \sqrt{2p_s}\cos(\omega_0 t + \pi) \\ &= -\sqrt{2p_s}\cos(\omega_0 t) \end{aligned} \tag{2.8}$$

The probability of bit-error for QPSK is the same as for BPSK:

$$P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \tag{2.9}$$

However, with two bits per symbol, the symbol error rate is increased:

$$\begin{aligned} P_s &= 1 - (1 - P_b)^2 \\ &= 2\,Q\left(\sqrt{\frac{E_s}{N_0}}\right) - Q^2\left(\sqrt{\frac{2E_s}{N_0}}\right) \end{aligned} \tag{2.10}$$

If the signal-to-noise ratio is high (as is necessary for practical QPSK systems) the probability of symbol error may be approximated:

$$P_b \approx 2Q\left(\sqrt{\frac{E_s}{N_0}}\right) \qquad (2.11)$$

Expressions for the symbol error-rate of rectangular QAM are not hard to derive but yield rather unpleasant expressions. They are most easily expressed in a *per carrier* sense:

$$P_{sc} = 2\left(1-\frac{1}{\sqrt{M}}\right)Q\left(\sqrt{\frac{3}{M-1}\frac{E_s}{N_0}}\right) \qquad (2.12)$$

$$P_s = 1-(1-P_{sc})^2 \qquad (2.13)$$

The bit-error rate will depend on the exact assignment of bits to symbols, but for a Gray-coded assignment with equal bits per carrier:

$$P_{bc} = \frac{4}{k}\left(1-\frac{1}{\sqrt{M}}\right)Q\left(\sqrt{\frac{3k}{M-1}\frac{E_b}{N_0}}\right) \qquad (2.14)$$

$$P_b = 1-(1-P_{bc})^2 \qquad (2.15)$$

M = Number of symbols in modulation constellation

E_b = Energy-per-bit

E_s = Energy-per-symbol with *k* bits per symbol $_b kE$

N_0 = Noise power spectral density (W/Hz)

P_b = Probability of bit-error

P_{bc} = Probability of bit-error per carrier

P_s = Probability of symbol-error

P_{sc} = Probability of symbol-error per carrier

Chapter 3

Block Turbo Code

3.1 Introduction :

Turbo codes were first presented at the International Conference on Communications in 1993 by C. Berrou. Until then, it was widely believed that to achieve near Shannon's bound performance, one would need to implement a decoder with infinite complexity or close.

Turbo codes can be achieved by serial or parallel concatenation of two (or more) codes called the constituent codes. The constituent codes can be either block codes or convolutional codes. Currently, most of the work on turbo codes have essentially focused on Convolutional Turbo Code (CTC)s and Block Turbo Code (BTC)s have been partially neglected. Yet, the BTC solution is more attractive for a wide range of applications. In 1994 we proposed a new soft-input/soft-output decoder [26] for all linear block codes and it has been showed that BTC had performances comparable to those of CTC using suboptimal weighting algorithms. BTC offers a good compromise between performance and complexity and is very attractive for implementation.

1. BTC resulted from the combination of three ideas that were known to all in the coding community

2. The utilization of block codes instead of commonly used non-systematic or systematic convolutional codes.

3. The utilization of soft input soft output decoding. Instead of using hard decisions, the decoder uses the probabilities of the received data to generate soft output which also contain information about the degree of certainty of the output bits.

4. Encoders and decoders working on permuted versions of the same information. This is achieved by using an interleaver.

3.2 Concatenated codes :

The power of Forward Error Correction codes can be enhanced by using the concatenated codes, which are shown in Figure 3.1. Concatenated codes were first introduced by Elias in 1954 [28]. The principle is to feed the output of one encoder (called the outer encoder) to the input of another encoder, and so on, as required. The final encoder before the channel is known as the inner encoder. The resulting composite code is clearly much more complex than any of the individual codes. However it can readily be decoded: we simply apply each of the component decoders in turn, from the inner to the outer.

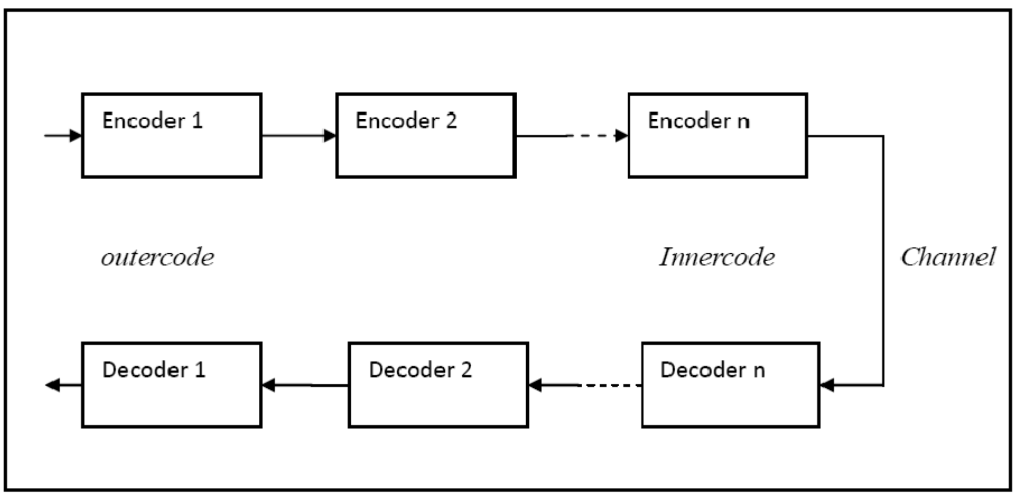


Fig 3.1 Principle of Concatenated codes

This simple scheme suffers a drawbacks which is called error propagation. If a decoding error occurs in a codeword, it usually results in a number of data errors. When these are passed on to the next decoder they may overwhelm the ability of that code to correct the errors. The performance of the outer decoder might be improved if these errors were distributed between a number of separate code-words. This can be achieved using an interleaver/de-interleaver. This interleaver (sometimes known as a rectangular or block interleaver) consists of a two-dimensional array, into which the data is read along its rows. Once the array is full, the data is read out by columns, thus permuting the order of the data.

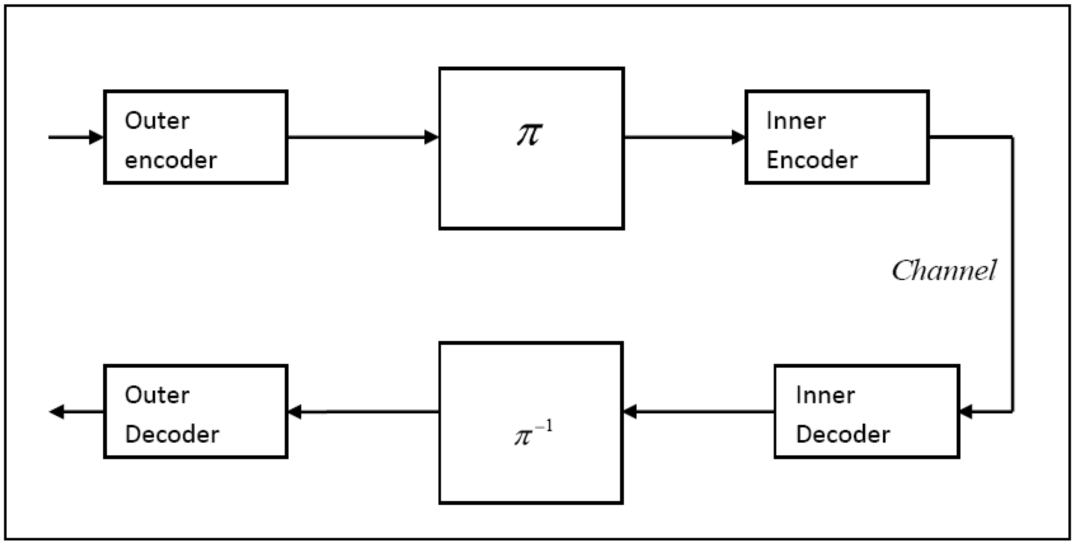


Fig 3.2 Concatenated encoder and decoder with interleaver

Interleaver may be placed between the outer and inner encoders of a concatenated code that uses two component codes, and the de-interleaver between the inner and outer decoders in the receiver, as shown in Figure 3.2 Then, provided the rows of the interleaver are at least as long as the outer codeword"s, and the columns at least as long as the inner data blocks, each data bit of an inner codeword falls into a different outer codeword.Hence, provided the outer code is able to correct at least one error, it can always cope with single decoding errors in the inner code.

3.3 Block Turbo Code :

Block turbo codes (or product codes) are serially concatenated codes [27] which were introduced by Elias in 1954 [28]. The concept of product codes is very simple and relatively efficient for building very long block codes by using two or more short block codes.

Let us consider two systematic linear block codes c^1 with parameters (n_1, k_1, δ_1) and c^2 with parameters (n_2, k_2, δ_2), where n_i, k_i and δ_i stand for codeword length, number of information bits, and minimum Hamming distance, respectively. The product code is obtained (as shown in Figure 3.3) by

1. placing ($k_1 \times k_2$) information bits in an array of k_1 rows and k_2 columns;
2. coding the k_1 rows using code c^2;
3. coding the k_2 columns using code c^1.

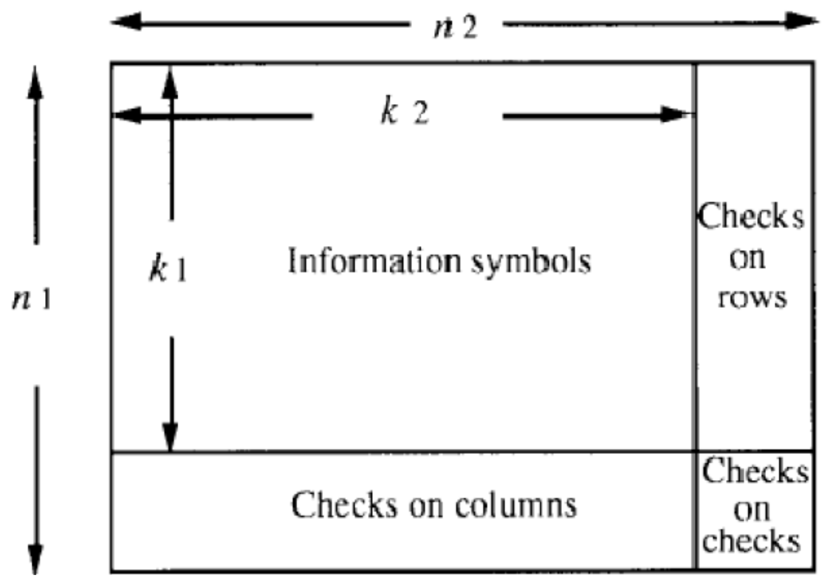


Fig 3.3 Construction of product code P = $c^1 \otimes c^2$

The parameters of the product code P are n = n_1 x n_2, k = $k_1 \times k_2$, $\delta = \delta_1 \times \delta_2$ and the code rate R is given by R = $R_1 \times R_2$, where R_i is the code rate of code c^i. Thus, we can build very long block codes with large minimum Hamming distance by combining short codes with small minimum Hamming distance. Given the procedure used to construct the product code, it is clear that the $(n_2 - k_2)$ last columns of the matrix are codewords of c^1. By using the matrix generator [19], one can show that the $(n_1 - k_1)$ last rows of matrix P are codewords of c^2. Hence, all of the rows of matrix P are codewords of c^1 and all of the columns of matrix P are codewords of c^2.

As indicated by Elias [28], these codes can be decoded by sequentially decoding the rows and columns of P in order to reduce decoding complexity. However, to achieve optimum performance, one must use MLD (soft decoding) of the component codes. Thus, we need soft-input/soft-output decoders to maintain optimum performance when decoding the rows and columns of P. Provided we have a soft-input/softoutput decoder for decoding the rows and columns of P, we can iterate the sequential decoding of P and thus reduce the BER after each iteration as for CTC [16].

3.4 Soft decoding of linear block codes :

Let us consider the transmission of binary elements {0,1} coded by a linear block code c with parameters (n , k , δ) on a Gaussian channel using binary symbols { +1, -1 }. We shall consider the following mapping of the symbols $1 \rightarrow +1$ and $0 \rightarrow -1$. The observation R = ($r_1, r_2, \ldots r_l, \ldots r_n$) at the output of the Gaussian channel for a transmitted codeword E = ($e_1, e_2, \ldots e_l, \ldots\ e_n$) is given by

$$R = E + G \tag{3.1}$$

where components g_l of G are additive white Gaussian noise (AWGN) samples of standard deviation σ.By using MLD, one can show that the optimum decision D = ($d_1, d_2, \ldots d_l, \ldots d_n$) corresponding to the transmitted codeword is given by

$$D = c^i \text{ if } \left|R - c^i\right|^2 \leq \left|R - c^l\right|^2 \qquad \forall \ \ l \ \in \ [1, 2^k] \qquad l \neq i \tag{3.2}$$

where $c^i = ({c_1}^i, \ldots {c_l}^i, \ldots\ {c_n}^i)$ is the ith codeword of c and

$$\left|R - c^i\right|^2 = \sum_{l=1}^{n} \left(r_l - {c_l}^i\right)^2 \tag{3.3}$$

is the squared Euclidean distance between R and c^i.When using an exhaustive search for the optimum codeword D, the computation complexity increases exponentially with k and becomes prohibitive for block codes with k > 6. As suggested by Gallager [29], one needs very long codes in order to approach channel capacity and the exhaustive search is not a realistic solution for those codes considered here with k > 10.

At high SNR, ML codeword D is located in the sphere of radius $(\delta - 1)$ centered on $Y = (\ y_1, \ldots y_l, \ldots y_n\ )$, where $y_l = 0.5\ (\ 1 + sgn(\ r_l))$ and $y_l\ \in \{0{,}1\}$ with a very high probability. Thus, we can limit the reviewed codewords in equation (3.2) to those in the sphere of radius $(\delta - 1)$ centered on Y. To reduce the number of reviewed codewords, only the set of the most probable codewords within the sphere are selected by using channel information R.The procedure used to identify the set of the most probable codewords is the following.

Step 1 : Determine the position of the p = $[\delta/2]$ least reliable binary elements of Y using R. The reliability of the elements of Y is defined later on.

Step 2 : Form test patterns T^q defined as all the n-dimensional binary vectors with a single " 1 " in the least reliable positions and " 0 " in the other positions, two " 1 "s in the least reliable positions and " 0 " in the other positions, and p " 1 "s in the least reliable positions and " 0 " in the other positions.

Step 3 : Form test sequences Z^q where $z_l{}^q = y_l \otimes t_l{}^q$ and decode Z^q using an algebraic (or hard) decoder and add the codeword C^q to subset Ω.

Decision D is then given by applying decision rule (3.2) with the reviewed codewords restricted to the subset of codewords Ω found at step 3 above. Note that the components of the codewords are mapped from{0,1} to {-1,+1}before computing the Euclidean distance. In step 1 the reliability of component y_j is defined using the log-likelihood ratio (LLR) of decision y_j.

$$\Lambda(y_j) = ln\left(\frac{Pr(e_j=+1/r_j)}{Pr(e_j=-1/r_j)}\right) = \left(\frac{2}{\sigma^2}\right) r_j \tag{3.4}$$

If we consider a stationary channel, we can normalize the LLR with respect to constant $2/\sigma^2$, and the relative reliability of y_j is then given by$|r_j|$.

Coming back to the decoding of a product code, the Chase algorithm [30] yields for each row (or column) the decision D of the component block code for a given input data R. To iterate the decoding procedure with maximum efficiency, we must compute the reliability of the decisions given by the Chase algorithm before decoding the columns (or rows).

3.5 Reliability of decision *D* given by soft-input decoder :

Once we have determined decision *D* of a row (or column) of the product code, we have to compute the reliability of each of the components of vector *D* in order to generate soft decisions at the output of the decoder. The reliability of decision d_j is defined using the LLR of transmitted symbol e_j , which is given by

$$\Lambda(d_j) = ln\left(\frac{Pr(e_j=+1/R)}{Pr(e_j=-1/R)}\right) \tag{3.5}$$

Here, the computation of the LLR differs from the previous case [in eqution (3.4)], in that we must take into account the fact that *D* is one of the 2^k codewords of *C*. Thus, by considering the different codewords of *C*, the numerator of (3.5) can be written as

$$Pr(e_j = +1/R) = \sum_{c^i \in s_j^{+1}} Pr(E = c^i/R) \tag{3.6}$$

where s_j^{+1} is the set of codewords $\{c^i\}$ such that $c_j^i = +1$ and the denominator of (3.5) can be written in the form

$$Pr(e_j = -1/R) = \sum_{c^i \in s_j^{-1}} Pr(E = c^i/R) \tag{3.7}$$

where s_j^{-1} is the set of codewords $\{c^i\}$ such that $c_j^i = +1$. By applying Bayes' rule to (3.6) and (3.7) and assuming that the different codewords are uniformly distributed, we obtain for $\Lambda(d_j)$ the following expression:

$$\Lambda(d_j) = ln\left(\frac{\sum_{c^i \in s_j^{-1}} p(R/E=c^i)}{\sum_{c^i \in s_j^{-1}} p(R/c^i)}\right) \tag{3.8}$$

Where $p(R/E = c^i) = \left(\frac{1}{\sqrt{2\pi\sigma}}\right)^n \exp\left(-\frac{|R-c^i|^2}{2\sigma^2}\right)$ (3.9)

is the probability density function of conditioned on This function decreases exponentially with the Euclidean distance between and Let and be the codewords, respectively, in and , at minimum Euclidean distance from By combining (8) and (9), we obtain the following relation:

$$\Lambda(d_j) = \frac{1}{2\sigma^2}\left(\left|R - c^{-1(j)}\right|^2 - \left|R - c^{+1(j)}\right|^2\right) + ln\left(\frac{\sum_i A_i}{\sum_i B_i}\right) \tag{10}$$

Where $A_i = exp\left(\frac{|R-c^{+1(j)}|^2 - |R-c^i|^2}{2\sigma^2}\right) \leq 1$ with $c^i \in s_j^{+1(j)}$

(3.11)

and

$$B_i = exp\left(\frac{\left|R-c^{-1(j)}\right|^2 - \left|R-c^i\right|^2}{2\sigma^2}\right) \leq 1 \quad \text{with} \quad c^i \in s_j^{-1(j)} \tag{3.12}$$

For high SNR, that is, $\sigma \rightarrow 0$, $\sum_i A_i \approx \sum_i B_i \rightarrow 1$ and thus the second term in (3.10) tends to zero. By neglecting the second term in (3.10), we obtain an approximation for the LLR of decision d_j equal to

$$\Lambda'(d_j) = \frac{1}{2\sigma^2}\left(\left|R - c^{-1(j)}\right|^2 - \left|R - c^{+1(j)}\right|^2\right) \tag{3.13}$$

By expanding (3.13) using (3.3), we obtain the following relation:

$$\Lambda'(d_j) = \frac{2}{\sigma^2}\left(r_j + \sum_{l=1,l\neq j}^{n} r_l c_l^{+1(j)} p_l\right) \tag{3.14}$$

$$\text{Where } p_l = 0 \qquad \text{if} \qquad c_l^{+1(j)} = c_l^{-1(j)}$$

$$= 1 \qquad \text{if} \qquad c_l^{+1(j)} \neq c_l^{-1(j)} \tag{3.15}$$

If we suppose that σ is constant, we can normalize $\Lambda'(d_j)$ with respect to the constant $\frac{2}{\sigma^2}$ and we obtain the following equation:

$$r_j' = r_j + w_j \tag{3.16}$$

with

$$w_j = \sum_{l=1,l\neq j}^{n} r_l c_l^{+1(j)} p_l \tag{3.17}$$

The normalized LLR r_j' is taken as the soft output of the decoder. It has the same sign as d_j and its absolute value indicates the reliability of the decision. Equation (3.16) indicates that r_j' is given by the soft-input data r_j plus a term w_j which is a function of the two codewords at minimum Euclidean distance from R and $\{ r_l \}$with $l \neq j$. The term w_j is a correction term applied to the input data and it plays the same role as the extrinsic information in CTC [31]. The extrinsic information is a random variable with a Gaussian distribution since it is a linear combination of identically distributed random variables. Furthermore, it is uncorrelated with the input data r_j'. As for CTC, the extrinsic information plays a very important role in the iterative decoding of product codes.

3.6 Computing the soft decision at the output of the soft-input decoder :

Computing the reliability of decision d_j at the output of the soft-input decoder requires two codewords $C^{+1(j)}$and $C^{-1(j)}$ as shown in (3.13). Obviously, soft decision *D* is one of these two codewords and we must find the second one, which we shall call *C*. *C* can be viewed as a competing codeword of *D* at minimum Euclidean distance from R with $c_j \neq d_j$. Given codeword *C* and *D*, one can show that the soft output is given by the following equation:

$$r_j' = \left(\frac{|R-C|^2 - |R-D|^2}{4} \right) \tag{3.18}$$

To find codeword *C*, one must increase the size of the space scanned by the Chase algorithm (as given in Section 3.4). For that purpose, we increase the number of least reliable bits used in the Chase decoder and also the number of test patterns. It is clear that the probability of finding *C* increases with the value of *p*. On the other hand, the complexity of the decoder increases exponentially with *p* and we must find a tradeoff between complexity and performance. This implies that in some cases we shall not be able to find a competing codeword *C*. In the event where codeword *C* is not found, we must find another method for computing the soft output. The solution we propose is to use the following equation:

$$r_j' = \beta \times d_j \qquad \text{with} \qquad \beta \geq 0 \tag{3.19}$$

This very rough approximation of the soft output is justified by the fact that:

1. The sign of soft output r_j' is equal to d_j [as shown in (3.18)], while only its absolute value or reliability is a function of *C*;

2. If *C* is not found in the space scanned by the Chase algorithm, then *C* is most probably far from R in terms of Euclidean distance;

3. If C is very far from R , then the probability that decision d_j is correct is relatively high and the reliability of d_j is also relatively high.

Thus, we propose to give a predefined value β to the reliability of those components of D for which there is no competing codeword C in subset Ω. The value of β was initially optimized by trial and error [26]. The equation for computing β is given as:

$$\beta \approx \left| ln\left(\frac{Pr(d_j = e_j)}{Pr(d_j \neq e_j)}\right) \right| = ln\left(\frac{Pr(d_j = e_j)}{Pr(d_j \neq e_j)}\right) \tag{3.20}$$

where $Pr(d_j = e_j)$ represents the probability that the decoder takes the correct decision and it takes its values in the interval [0.5, 1]. When $Pr(d_j = e_j) \rightarrow 1$, then $\beta \rightarrow \infty$, and when $Pr(d_j = e_j) \rightarrow 0.5$, then $\beta \rightarrow 0$. Equation (3.20) for computing β is coherent with the notion of reliability of decision d_j. When the probability of taking the correct decision tends to one, the reliability of d_j tends to infinity, and when it tends to 0.5, the reliability tends to zero. In fact, β can be considered as an average value of the reliability of those decisions d_j for which there is no competing codeword C in subset Ω, while (3.18) gives a bit-by-bit estimation of the reliability. It is clear that the soft output given by (3.19) is less accurate than the one using (3.18). However, one can understand that we need a more accurate estimation of the soft output for those decisions where a competing codeword C is at a slightly greater distance from R than D. On the other hand, when C is very far from R, an average value of the reliability can be considered as sufficient.

3.7 Turbo decoding of product codes :

The decoding procedure described below is generalized by cascading elementary decoders illustrated in Fig. 3.4. Let us consider the decoding of the rows and columns of a product code P described in Section 3.3 and transmitted on a Gaussian channel using BPSK signaling. On receiving matrix [R] corresponding to a transmitted codeword [E], the first decoder performs the soft decoding of the rows (or columns) of P using as input matrix [R]. Soft-input decoding is performed using the Chase algorithm (as given in Section 3.4) and the soft output is computed using (3.18) or (3.19).

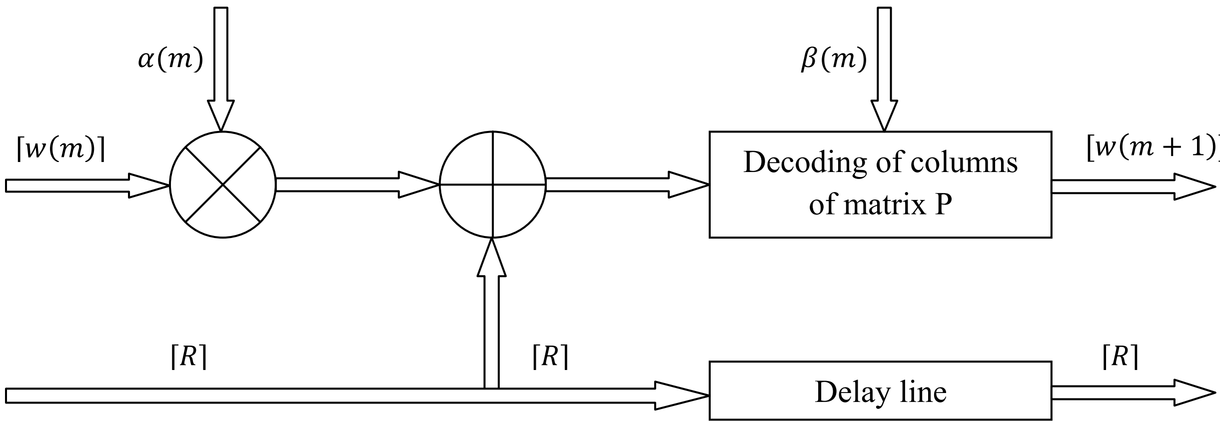


Fig 3.4 Block diagram of elementary block turbo decoder

By subtracting the soft input from the soft output [given in (3.16)] we obtain the extrinsic information[$W(2)$] where index 2 indicates that we are considering the extrinsic information for the second decoding P of which was computed during the first decoding of P. The soft input for the decoding of the columns (or rows) at the second decoding of P is given by

$$[\,R(2)] = [\,R] + \alpha(2)[w(2)] \qquad (3.21)$$

where $\alpha(2)$ is a scaling factor which takes into account the fact that the standard deviation of samples in matrix [R] and in matrix [w] are different (as given in [16]). The standard deviation of the extrinsic information is very high in the first decoding steps and decreases as we iterate the decoding. This scaling factor α is also used to reduce the effect of the extrinsic information in the soft decoder in the first decoding steps when the BER is relatively high. It takes a small value in the first decoding steps and increases as the BER tends to zero.

Chapter 4

Coded Orthogonal Frequency Division Multiplexing system model

4.1 Introduction :

Orthogonal Frequency Division Multiplexing (OFDM) also known as discrete multitone modulation (DMT), is based upon the principle of frequency division multiplexing (FDM), but it utilized as a digital modulation scheme. The bit stream that is to be transmitted is split into several parallel bit streams, typically thousands. The available frequency spectrum is divided into sub-channels and each low rate bit stream is transmitted over one sub channel by modulating subchannel by , modulating a sub-carrier using a standard modulation scheme, for example: PSK, QAM. The sub-carrier frequencies are chosen so that the modulated data streams are orthogonal to each other, meaning that the signals are totally independentand cross talk between the subchannels is eliminated. It is achieved by ensuring that the carriers are placed exactly at the nulls in the modulation spectra of each other.

Orthogonal Division Multiplexing (OFDM) has grown to be the most popular communications systems in high speed communications in the last decade. In fact, it has been said by many industry leaders that OFDM technology is the future of wireless communications.

Coded OFDM (COFDM) is a term used for a system in which the error control coding and OFDM modulation processes work closely together. COFDM , systems are able to achieve excellent performance on frequency selective channels because of the combined benefits of multicarrier modulation and coding.

4.2 Frequency division multiplexing modulation :

Frequency division multiplexing (FDM) extends the concept of single carrier modulation by using multiple subcarriers within the same single channel. The total data rate to be sent in the channel is divided between the various subcarriers. The data do not have to be divided evenly nor do they have to originate from the same information source. Advantages include using separate modulation/ demodulation customized to a particular type of data, or sending out banks of dissimilar data that can be best sent using multiple, and possibly different, modulation schemes.

FDM offers an advantage over single-carrier modulation in terms of narrowband frequency interference since this interference will only affect one of the frequency sub-bands. The other subcarriers will not be affected by the interference. Since each subcarrier has a lower information rate, the data symbol periods in a digital system will be longer, adding some additional immunity to impulse noise and reflections. FDM systems usually require a guard band between modulated subcarriers to prevent the spectrum of one subcarrier from interfering with another. These guard bands lower the system's effective information rate when compared to a single carrier system with similar modulation.

4.3 Orthogonal frequency division multiplexing (OFDM) :

OFDM is a combination of modulation and multiplexing. Multiplexing generally refers to independent signals, those produced by different sources. In OFDM [2], the question of multiplexing is applied to independent signals but these independent signals are a sub-set of the one main signal. In OFDM the signal itself is first split into independent channels, modulated by data and then re-multiplexed to create the OFDM carrier.

If the FDM system above had been able to use a set of subcarriers that were orthogonal to each other, a higher level of spectral efficiency could have been achieved. The guard bands that were necessary to allow individual demodulation of subcarriers in an FDM system would no longer be necessary. The use of orthogonal subcarriers would allow the subcarriers' spectra to overlap, thus increasing the spectral efficiency. As long as orthogonality is maintained, it is still possible to recover the individual subcarrier signals despite their overlapping spectrums.

As an analogy, a FDM channel is like water flow out of a faucet, a whole bunch of water coming all in one stream; In contrast the OFDM signal is like a shower from which same amount of water will come as a lot of small streams. In a faucet all water comes in one big stream and cannot be sub-divided. OFDM shower is made up of a lot of little streams.

Fig 4.1 (a) A Regular-FDM single carrier (b) Orthogonal-FDM

The advantage one over the other is that if I put my thumb over the faucet hole, I can stop the water flow but I cannot do the same for the shower. So although both do the same thing, they respond differently to interference. Both methods carry the exact same amount of data. But in case of any interfere to some of these small streams, only some part of data in the OFDM method will suffer.

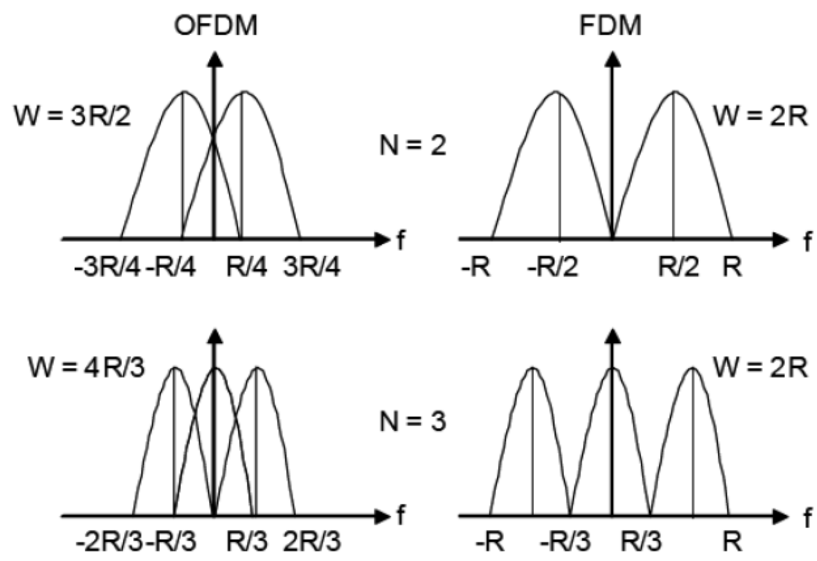


Fig 4.2 Frequency Efficiency of OFDM over FDM for Two and Three Subchannels

Figure 4.2 illustrates the difference between the non-overlapping and overlapping multicarrier modulation [18], techniques. It can be seen that almost half of the bandwidth is saved by overlapping the spectra. As more and more carriers are added, the bandwidth approaches, $\frac{N+1}{N}$ bits per Hz. So larger the number of carriers, better the spectral efficiency.

The main concept in OFDM [19], is orthogonality of the subcarriers. The "orthogonal" part of the OFDM name indicates that there is a precise mathematical relationship between the frequencies of the carriers in the system. It is possible to arrange the carriers in an OFDM Signal so that the sidebands of the individual carriers overlap and the signals can still be received without adjacent carrier's interference. In order to do this the carriers must be mathematically orthogonal.

The Carriers are linearly independent (i.e. orthogonal) if the carrier spacing is a multiple of $1/T_s$. Where, T_s is the symbol duration. The orthogonality among the carriers can be maintained if the OFDM signal is defined by using Fourier transform procedures. The OFDM system transmits a large number of narrowband carriers, which are closely spaced. Note that at the central frequency of the each sub channel there is no crosstalk from other sub channels.

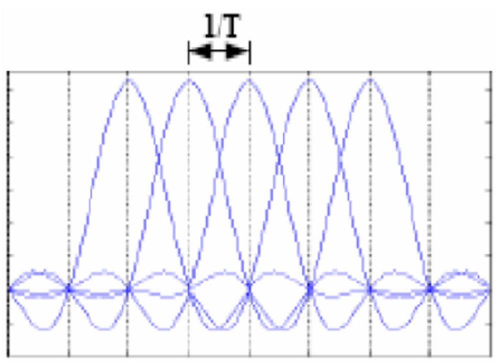


Fig 4.3 freq. spectrum of OFDM

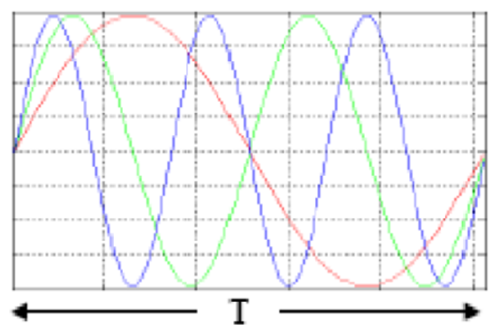


Fig 4.4 waveform of carriers in OFDM

The frequency spectrum of an OFDM transmission is illustrated in figure 4.3. Each sinc of the frequency spectrum in the Fig 4.3 corresponds to a sinusoidal carrier modulated by a rectangular waveform representing the information symbol. One could easily notice that the frequency spectrum of one carrier exhibits zero-crossing at central frequencies corresponding to all other carriers. At these frequencies, the inter carrier interference is eliminated, although the individual spectra of subcarriers overlap. It is well known, orthogonal signals can be separated at the receiver by correlation techniques.

The waveform of some carriers in a OFDM transmission is illustrated in Fig 4.4 The figure indicates the spectrum of carriers significantly over laps over the other carrier. This is contrary to the traditional FDM technique in which a guard band is provided between each carrier. From the Figures illustrated, it is clear that OFDM is a highly efficient system and hence is often regarded as the optimal version of multi-carrier transmission schemes. The number of sub channels transmitted is fairly arbitrary with certain broad constraints, but in practical systems, sub channels tend to be extremely numerous and close to each other. For example the number of carriers in 802.11 wireless LAN is 48 while for Digital Video Broadcast (DVB) it is as high as 6000 sub carriers.

4.4 Coded orthogonal frequency division multiplexing :

Coded OFDM (COFDM) is a term used for a system in which the error control coding and OFDM modulation processes work closely together. COFDM , systems are able to achieve excellent performance on frequency selective channels because of the combined benefits of multicarrier modulation and coding. Due to the effects of noise and multipath fading in the channel, the transmitted signal arrives at the receiver with some errors. The errors in the demodulated data are characterized in terms of a BER, which is directly proportional to the symbol rate and inversely proportional to transmitter power and bit-energy to noise power spectral density ratio . The bit error rate is an important performance parameter of digital communication systems.

In forward error correction coding, a certain number of redundant bits are added to data bits in a particular pattern according to the type of the code. In other words, for every k data bits, n coded bits are transmitted, where $n > k$. In the receiver, the k data bits can be recovered by performing a decoding operation on the n received coded bits.

The transmission conditions in wireless communication channels are severe due to multipath fading and the variation of the signal-to-noise power ratio. Therefore, in order to design a communication system with an acceptable BER, error correction coding must be used to protect the data from transmission errors.

As long as the signal-to-noise ratio (SNR) is high and the channel is relatively flat, error correction coding may be unnecessary in OFDM systems. However, uncoded OFDM systems do not perform well in fading channels.

The code rate r = k/n is a ratio of the number of data bits to the total number of coded bits transmitted per code word.

As a result of redundancy, the number of bit errors can be expected to increase. However, the reliability of demodulated data is increased because redundancy is used to correct some of the errors. A better BER can be achieved at the output of the decoder by using an appropriate coding scheme. The BER improvement provided by the channel coding is generally expressed in terms of the required E_b/N_0 to achieve the same performance without coding. This difference in E_b/N_0 is called coding gain as given by

$$G_{dB} = (E_b/N_0)_{uncoded} - (E_b/N_0)_{coded} \quad (4.1)$$

Using commercially available standard error correction systems, coding gains up to 6 to 9 dB are achievable. Linear block codes, convolutional codes [20], and turbo codes [21], are the most widely used error correction codes. But with the evolution of the Turbo codes (FEC), convolutional and Reed-Solomon codes are replaced by this, in many applications. This is because of Turbo codes come closest to approaching the Shannon limit, [16], the theoretical. This is because of turbo codes uses iterative MAP decoding SISO decoders. This Turbo codes [22], are used in UMTS third-generation cellular standard, as standardized by the Third-Generation Partnership Project (3GPP) because of its better performance compared to all.

Turbo codes are based on parallel convolutional concatenated codes (PCCC) [23]. A concatenated code consists of two separate codes in series in which the first code, called outer code, directly takes the information bits and encodes them. The second code, called inner code, takes the bits coded by the outer code and encodes them. In order to prevent decoding errors from passing from one coder to the other, the errors are distributed by using the interleaving/deinterleaving operation.

4.5 COFDM transmission & reception :

The basic principle of OFDM, as mentioned in the above, is to divide a high-rate data stream into N lower rate streams and to transmit them at the same time over a number of subcarriers. Since the symbol duration is increased, the relative amount of dispersion in time caused by multipath delay spread is decreased. Intersymbol interference (ISI) is another problem, which can almost be eliminated by introducing a guard time in every OFDM symbol. In order to avoid the ICI, an OFDM symbol is cyclically extended by adding a guard time. A general block diagram of the transmitter and the receiver for the COFDM scheme is shown in Figure 4.5. Here is the transmitted sequence and is the estimated sequence of transmitted signal.

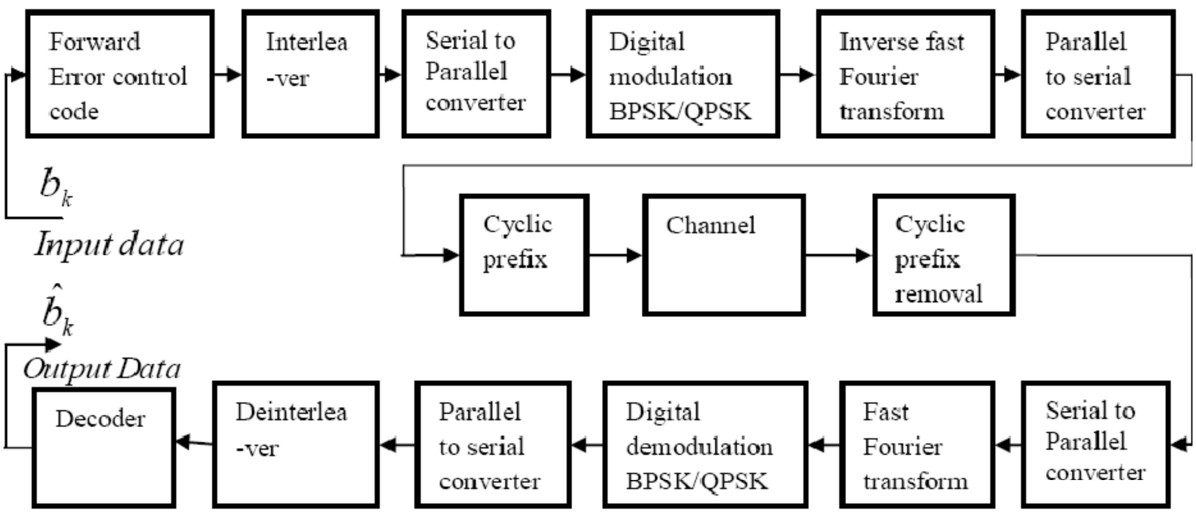


Fig 4.5 COFDM system diagram

4.5.1 Forward error control & Interleaving :

The transmission conditions in wireless communication channels are severe due to multipath fading. Therefore, in order to design a communication system with an acceptable BER for lower level of signal to noise ratio, error correction coding must be used to protect the data from transmission errors. Linear block codes, convolutional codes [20], and turbo codes are the most widely used error correction codes. Turbo codes used in this thesis are based on parallel concatenated convolutional code and have best performance compared to convolutional and Reed Solomon codes due to its soft-in, soft-out (SISO) decoding algorithm. The performance of a decoder is significantly enhanced if, in addition to the " hard decision " made by the demodulator on the current symbol, some additional " soft information " on the reliability of that decision is passed to the decoder. For example, if the received signal is close to a decision threshold (say between 0 and 1) in the demodulator, then that decision has low reliability, and the decoder should be able to change it when searching for the most probable codeword. Making use of this information in a conventional decoder, called soft decision decoding, leads to a performance improvement of around 2dB in most cases.

Soft information usually takes the form of a log-likelihood ratio for each data bit. The likelihood ratio is the ratio of the probability that a given bit is '1' to the probability that it is '0'. If we take the logarithm of this, then its sign corresponds to the most probable hard decision on the bit (if it is positive, '1' is most likely ; if negative, then '0'). The absolute magnitude is a measure of our certainty about this decision.

The performance of the system can be improved if burst errors are distributed over the other code words. This can be achieved by interleaver/de-interleaver. A block interleaver consists of a two-dimensional array, into which the data are read along its rows. When the array is full, the data are read out by the columns, thus the order of the data is permuted. The original order can be received by the corresponding deinterleaver in which the data are read in by columns and read out by rows. But in case of turbo-codes the interleaver is not usually rectangular but, it is pseudorandom, that is the data is read out in a predefined pseudorandom order. The design of interleaver is one of the key features of turbo-codes.

4.5.2 Symbol Mapping :

The interleaved and rearranged data are mapped onto constellation points in accordance with the digital modulation [23] type (BPSK, QPSK, QAM).

4.5.3 Serial to parallel converter :

In OFDM high bit rate data is divided into N low bit rate parallel data streams and then transmitted simultaneously with deferent frequencies.

4.5.4 FFT and IFFT :

OFDM systems are implemented using a combination of fast Fourier Transform (FFT) and inverse fast Fourier Transform (IFFT) blocks that are mathematically equivalent versions of the DFT and IDFT, respectively, but more efficient to implement. An OFDM system treats the source symbols (e.g., the QPSK or QAM symbols that would be present in a single carrier system) at the transmitter as though they are in the frequency-domain. These symbols are used as the inputs to an IFFT block that brings the signal into the time domain. The IFFT takes in N symbols at a time where N is the number of subcarriers in the system. Each of these N input symbols has a symbol period of T seconds. The basis functions for an IFFT are N orthogonal sinusoids. These sinusoids each have a different frequency and the lowest frequency is DC. Each input symbol acts like a complex weight for the corresponding sinusoidal basis function. Since the input symbols are complex, the value of the symbol determines both the amplitude and phase of the sinusoid for that subcarrier. The IFFT output is the summation of all N sinusoids. Thus, the IFFT block provides a simple way to modulate data onto N orthogonal subcarriers. The block of N output samples from the IFFT make up a single OFDM symbol. The length of the OFDM symbol is NT where T is the IFFT input symbol period mentioned above.

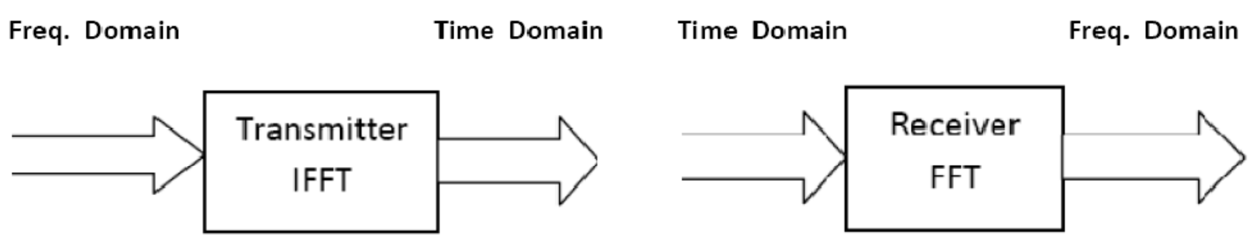


Fig 4.6 Block diagram of FFT and IFFT system

After some additional processing, the time-domain signal that results from the IFFT is transmitted across the channel. At the receiver, an FFT block is used to process the received signal and bring it into the frequency domain. Ideally, the FFT output will be the original symbols that were sent to the IFFT at the transmitter. When plotted in the complex plane, the FFT output samples will form a constellation, such as 16-QAM. However, there is no notion of a constellation for the time-domain signal. When plotted on the complex plane, the time-domain signal forms a scatter plot with no regular shape. Thus, any receiver processing that uses the concept of a constellation (such as symbol slicing) must occur in the frequency- domain. The block diagram in Figure 4.6 illustrates the switch between frequency-domain and time domain in an OFDM system.

4.5.5 Parallel to serial converter :

Parallel data from the IFFT block converted into serial form and then transmitted to guard interval block.

4.5.6 Cyclic prefix :

For a given system bandwidth the symbol rate for an OFDM signal is much lower than a single carrier transmission scheme. For example for a single carrier BPSK modulation, the symbol rate corresponds to the bit rate of the transmission. However for OFDM the system bandwidth is broken up into N subcarriers, resulting in a symbol rate that is N times lower than the single carrier transmission. This low symbol rate makes OFDM naturally resistant to effects of Inter-Symbol Interference (ISI) caused by multipath propagation. Multipath propagation is caused by the radio transmission signal reflecting off objects in the propagation environment, such as walls, buildings, mountains, etc. These multiple signals arrive at the receiver at different times due to the transmission distances being different. This spreads the symbol boundaries causing energy leakage between them. The effect of ISI on an OFDM signal can be further improved by the addition of a guard period to the start of each symbol. This guard period is a cyclic copy that extends the length of the symbol waveform. Each subcarrier, in the data section of the symbol, (i.e. the OFDM symbol with no guard period added, which is equal to the length of the IFFT size used to generate the signal) has an integer number of cycles. Because of this, placing copies of the symbol end-to-end results in a continuous signal, with no discontinuities at the joins [4]. Thus by copying the end of a symbol and appending this to the start results in a longer symbol time. Figure 4.7 shows the insertion of a guard period.

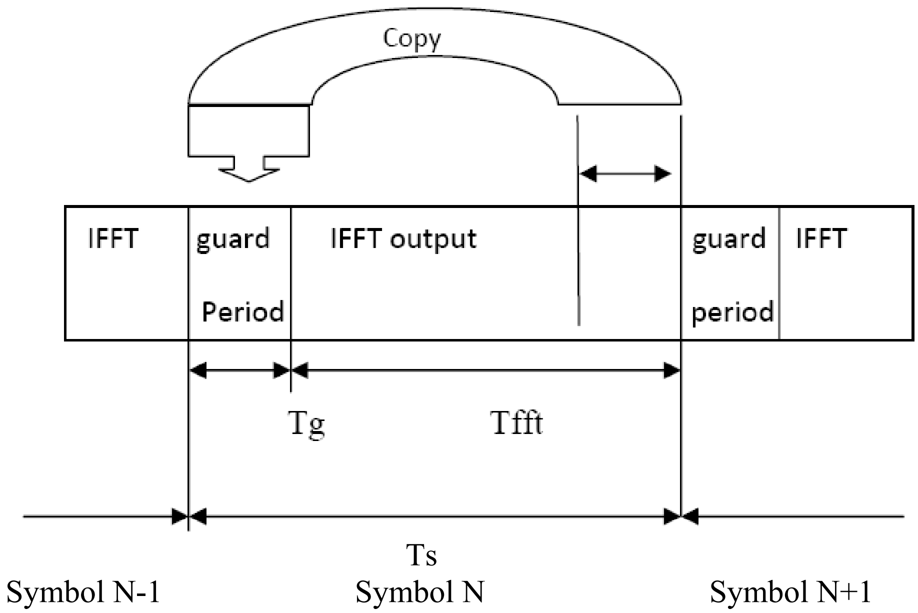


Fig 4.7 Addition of Guard Period to an OFDM Signal

The total length of the symbol is $T_s = T_g + T_{fft}$ where T_s is the total length of the symbol in samples, T_g is the length of the guard period in samples, and T_{fft} is the size of the IFFT used to generate the OFDM signal. In addition to protecting the OFDM from ISI, the guard period also provides protection against time-offset errors in the receiver.

Length Adaptive Cyclic prefix :

Conventional OFDM transmission system uses a fixed-length Cyclic Prefix to counteract Inter-Symbol Interferences (ISI) caused by channel delay spreading under wireless mobile environment. This may cause considerable performance deterioration when the CP length is less than the maximum RMS delay spread of the channel, or may decrease the system power and spectrum efficiency when it is much larger. Although the CPs are crucial to OFDM system, they introduce significant overhead.. So in order to avoid this, a novel Orthogonal Frequency Division Multiplexing (OFDM) transmission with length adaptive cyclic prefix is used. AOFDM-VCPL utilizes the preamble or pilot sub-carriers of each OFDM packet to estimate the channel RMS delay spread and then uses a criterion to calculate the CP length, which finally affects the OFDM transmitter.

4.5.7 Receiver :

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On the receiver side of the COFDM system, corrupted signal by the channel is received and it basically does the reverse operation to the transmitter. The guard period is removed. The FFT of each symbol is then taken to find the original transmitted spectrum. The phase angle of each transmission carrier is then evaluated and converted back to the data word by demodulating the received phase. The data words are then combined back to the same word size as the original data.

4.6 Advantages of COFDM :

OFDM has several advantages over single carrier modulation systems and these make it a viable alternative for CDMA in future wireless networks. In this section, some of these advantages are discussed.

4.6.1 Efficient Modulation and Demodulation :

Modulation and Demodulation of the sub-carriers is done using IFFT and FFT methods respectively, which are computationally efficient. By performing the modulation and demodulation in the digital domain, the need for highly frequency stable oscillators is avoided.

4.6.2 High Spectral Efficiency :

OFDM achieves high spectral efficiency by allowing the sub-carriers to overlap in the frequency domain. At the same time, to facilitate inter-carrier interference free demodulation of the sub-carriers, the sub-carriers are made orthogonal to each other. If the number of subcarriers in 'N', the total bandwidth required is

$$BW_{Total} = \frac{N+1}{T_s} \tag{4.2}$$

For large values of N, the total bandwidth required can be approximated as

$$BW_{Total} = \frac{N}{T_s} \tag{4.3}$$

On the other hand, the bandwidth required for serial transmission of the same data is

$$BW_{Total} = \frac{N}{T_s} \tag{4.4}$$

Thus we achieve very high spectral gain in OFDM compared to the single carrier serial transmission case.

4.6.3 Multipath Delay Spread Tolerance :

OFDM is highly immune to multipath delay spread that causes inter-symbol interference in wireless channels. Since the symbol duration is made larger (by converting a high data rate signal into „N" low rate signals), the effect of delay spread is reduced by the same factor. Also by introducing the concepts of guard time and cyclic extension, the effects of intersymbol interference (ISI) and inter-carrier interference (ICI) is removed completely.

4.6.4 Immunity to Frequency selective fading Channels :

If the channel undergoes frequency selective fading, then complex equalization techniques are required at the receiver for single carrier modulation techniques. But in the case of OFDM the available bandwidth is split among many orthogonal narrowly spaced sub-carriers. Thus the available channel bandwidth is converted into many narrow flat-fading sub-channels. Hence it can be assumed that the sub-carriers experience flat fading only, though the channel gain/phase associated with the sub-carriers may vary. In the receiver, each sub-carrier just needs to be weighted according to the channel gain/phase encountered by it. Even if some sub-carriers are completely lost due to fading, proper coding and interleaving at the transmitter can recover the user data.

4.7 Applications of COFDM :

Orthogonal Frequency Division Multiplexing is a new technology whose applications just being explored. The primary applications are in multimedia push technology and in wireless LAN.

4.7.1 Digital Audio Broadcasting (DAB) :

Digital Audio Broadcasting is a new multimedia push technology, with a good sound quality and better spectrum efficiency. This is achieved by the use of OFDM technology. The

DAB [21], system samples audio at a sample rate of 48 kHz and a resolution of 22bits .Then the data is compressed to between 32 and 384 KBPS. A rate ¼ convolution code is used with constraint length 7. The total data rate is about 2.2Mbps. The frame time is 24ms. QPSK modulation is performed at the transmitter. The advantage of using OFDM for DAB is that the OFDM suffers very little from delay spread and also that the OFDM system has high spectral efficiency.

4.7.2 Digital Video Broadcasting (DVB) :

Digital Video Broadcasting (DVB) is an ETSI standard for broadcasting Digital Television over satellites, cables and thorough terrestrial (wireless) transmission [20]. Terrestrial DVB operates in either of 2 modes called 2k and 8k modes with 1705 carriers and 6817 carriers respectively. It uses QPSK, 16 QAM or 64 QAM subcarrier modulations. It also uses pilot subcarriers for recovering amplitude and phase for coherent demodulation.

4.7.3 Wireless LAN Applications :

The IEEE 802.11 committee has its OFDM parameters are as shown in Table 4.1.

Data rate	6,9,12,18,24,36,48,54 Mbps
Modulation	BPSK, QPSK, 16-QAM, 64-QAM
Coding rate	½, 2/3, 3/4
No of Subcarriers	52
No of pilots	4
OFDM symbol duration	4 μs
Guard interval	800 ns
Sub-Carrier spacing	312.5 kHz
3 dB Bandwidth	16.56 MHz
Channel spacing	20 MHz

Table 4.1 Parameters of IEEE 802.11

Chapter 5

Communication channels

5.1 Introduction :

In wireless communications, a practical communication channel is often modeled by a random attenuation of the transmitted signal, followed by additive noise. The attenuation captures the loss in signal power over the course of the transmission, and the noise in the model captures external interference and/or electronic noise in the receiver. Hence, depending on the application, the mathematical model for the communication system includes a model for the distortion introduced by the transmission medium, and termed the communication channel, or channel for short.

A communications channel refers to the medium through which information is transmitted from a sender (or transmitter) to a receiver. In practice, this can mean many different methods of facilitating communication, including:

1. A connection between initiating and terminating nodes of a circuit.
2. A path for conveying electrical or electromagnetic signals, usually distinguished from other parallel paths.
3. The portion of a storage medium, such as a track or a band, that is accessible to a given reading or writing station or head.
4. In a communications system, the part that connects a data source to a data sink.

Type of communications channels:

1. Simplex communication (1 way)
2. Duplex communication (2 ways)

5.2 Additive white gaussian noise (AWGN) :

In the study of communication systems, the classical (ideal) additive white Gaussian noise (AWGN) channel, with statistically independent Gaussian noise samples corrupting data samples free of intersymbol interference (ISI), is the usual staring point for understanding basic performance relationships. An AWGN channel adds white Gaussian noise ti the signal that passes through it.

In constructing a mathematical model for the signal at the input of the receiver, the channel is assumed to corrupt the signal by the addition of white Gaussian noise as shown in Figure 5.1 below, therefore the transmitted signal, white Gaussian noise and received signal are expressed by the following equation with s (t), n (t) and r (t) representing those signals respectively:

$$r(t) = s(t) + n(t) \tag{5.1}$$

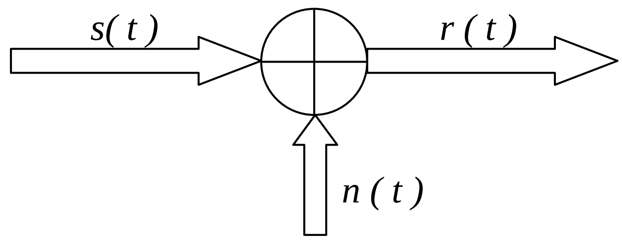

Fig 5.1 Received signal corrupted by AWGN

Where $n(t)$ is a sample function of the AWGN process with probability density function (pdf) and power spectral density as follows

$$\theta_{nm}(f) = \frac{1}{2} N_0 \left[W/_{Hz} \right] \tag{5.2}$$

Where N_0 is a constant and called the noise power density.

5.3 Multipath :

In wireless communications, multipath is the propagation phenomenon that results on radio signals reaching the receiving antenna by two or more paths. Causes of multipath include atmospheric ducting, ionospheric reflection and refraction and reflection from terrestrial object such as mountains and buildings.

The effects of multipath include constructive and destructive interference and phase shifting of the signal. This causes Rayleigh Fading named after Lord Rayleigh. Rayleigh fading with a strong line of sight is said to have a Rician distribution or tobe Rician fading. In digital radio communications such as GSM Multipath can cause errors and affect the quality of communications. The errors are due to Intersymbol Interference (ISI). Equalizers are often used to correct the ISI. Alternatively, techniques such as orthogonal frequency division modulation and Rake receivers may be used.

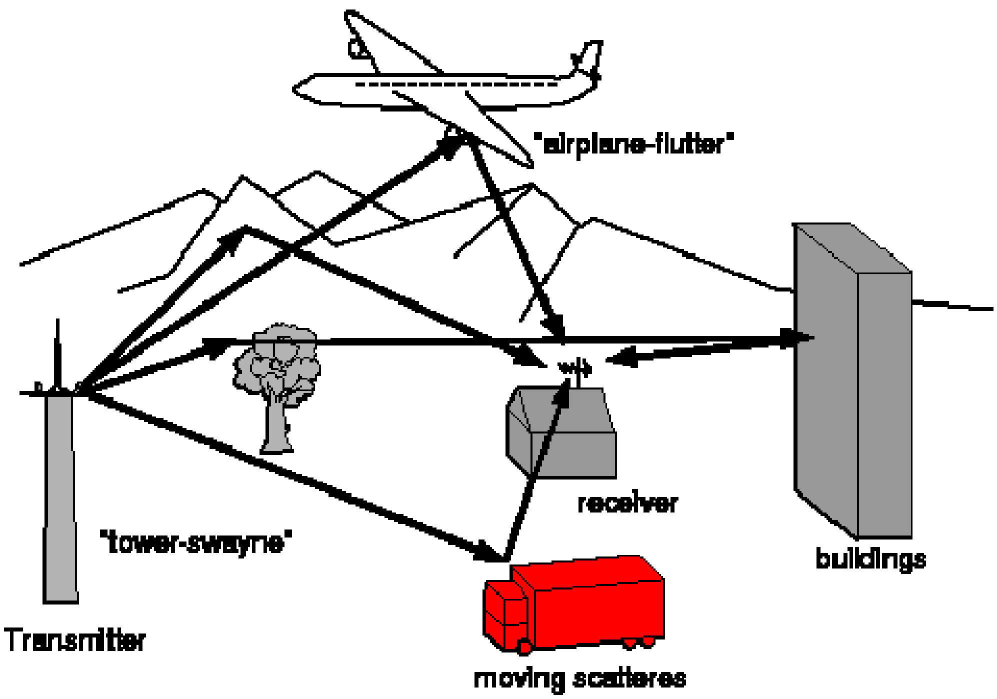

Fig 5.2 Principle of multipath channel

As shown in the Figure 5.2 , the path between base station and mobile stations of terrestrial mobile communications is characterized by various obstacles and reflections. The radio wave transmitted from the base station radiates in all directions. These radio waves, including reflected waves that are reflected off of various objects, diffracted waves, scattering waves, and the direct wave from the base station to the mobile station. Therefore the path lengths of the direct, reflected, diffracted, and scattereing waves are different, the time each takes to reach the mobile station is different. The phase of the incoming wave also varies because of the reflection. As a result, the receiver receives a superposition consisting of several waves having different phase and time of arrival. The generic name of a radio wave in which the time of arrival is retarded in comparison with this direct wave is called a delayed wave. Then, the reception environment characterized by a superposition of delayed waves is called multipath propaation environment.

5.4 Fading :

Fading is about the phenomenon of loss of signal in telecommunications. Fading or fading channels refers to mathematical models for the distortion that a carrier modulated tele-communication signal experiences over certain propagation media. Short team fading also known as multipath induced fading is due to multipath propagation. Fading results from the superposition of transmitted signals that have experienced differences in attenuation, delay and phase shift while travelling from the source to the receiver. It may be caused by attenuation of a single signal.

The most common types of fading are known as “slow fading” and “fast fading” as they apply to a mobile radio environment. Fading refers to the time variation of the received signal power caused by changes in the transmission medium or path.

Slow fading: Shadowing or Large Scale Fading is a kind of fading caused by larger movements of a mobile or obstructions within the propagation environment. Fast fading also known as Multipath fading or small scale fading is a kind of fading occurring with small movements of a mobile. The best way to combat fading is to ensure that multiple versions of the same signal are transmitted, received and coherently combined. This is usually termed diversity and is sometimes acquired through multiple antennas. Mathematically, the simplest model for the fading phenomenon is multiplication of the signal waveform with a time dependent coefficient which is often modeled as a random variable, making the received signal to noise ratio a random quantity.

Fading channel models are often used to model electromagnetic transmission of information over wireless media such as with cellular phones and in broadcast communication. Small scale fading is usually divided into fading based on multipath time delay spread and that based on Doppler spread.

There are two types of fading based on multipath time delay spread:

1. **Flat fading :** The bandwidth of the signal is less than the coherence Bandwidth of the channel or the delay spread is less than the symbol period.

2. **Frequency selective fading :** The bandwidth of the signal is greater than the coherence bandwidth of the channel or the delay spread is greater than the symbol period.

There are two types of fading based on Doppler spread:

1. **Fast fading :** There exist a high Doppler spread and the coherence time is less than the symbol time and the channel variations are faster than baseband signal variation.

2. **Slow fading :** It has a low Doppler spread. The coherence time is greater than the symbol period and the channel variations are slower than the baseband signal variation.

5.5 Rayleigh fading channel :

Raleigh fading [25], is a statistical model for the effect of a propagation environment on a radio signal such as that used by wireless devices. It assumes that the power of a signal that has passed through such a transmission medium (also called a communications channels) will vary randomly or fade according to a Raleigh distribution – the radial component of the sum of two uncorrelated Gaussian random variables. It is reasonable model for tropospheric and ionospheric signal propagation as well as the effect of heavily built up urban environment on radio signals. Raleigh fading is most applicable when there is no line of sight between the transmitter and receiver.

In a multipath propagation environment, the received signal is sometimes weakened or intensified. The signal level of the received wave changes from moment to moment. Multipath fading raises the error rate of the received data.

The delayed wave with incident angle is given by the following equation and corresponding to Figure 5.3, when a continuous wave of single frequency f_c Hz is transmitted from the base station.

$$r_n\,(t) = R_e[e_n\,(t)\,exp\,j\;(2\pi f_c t)] \tag{5.3}$$

Where R_e [ ] indicates the real part of a complex number that gives the complex envelope of the incoming wave from the direction of the number *n*.

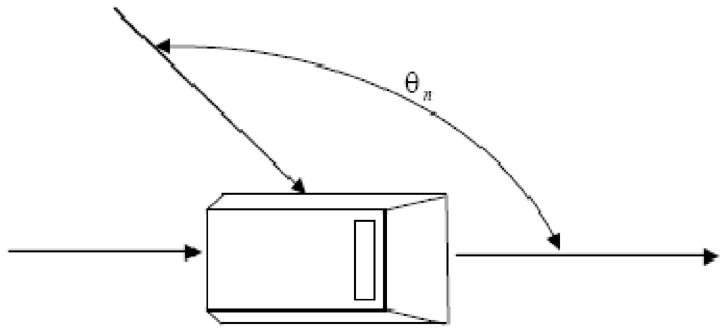

Fig 5.3 Delayed wave with incident angle

Moreover, j is a complex number. e_n is given in equation bellow by using propagation path length from the base station of the incoming waves: *Ln(m)*, the speed of the mobile station, *v* (m/s), and the wavelength, $\lambda(m)$.

$$e_n(t) = R_n(t)\exp j\left(-\frac{2\pi\ (L_n - v_t\cos\theta_n)}{\lambda} + \phi_n\right)$$

$$= x_n(t) + jy_n(t) \qquad (5.4)$$

Where R_n and θ_n are the envelope and phase of the *n*th incoming wave. $x_n(t)$ and $y_n(t)$ are the in-phase and quadrature phase factors of $e_n(t)$, respectively. The incoming *n*th wave shifts the carrier frequency as $v_t\cos\theta_n$ (Hz) by the Doppler shift f_d , has a maximum value of v/λ , when the incoming wave comes from the running direction of the mobile station in $\cos\theta_n = 1$. Then this maximum is the largest Doppler shift. The delayed wave that comes from the rear of the mobile station also has a frequency shift of $-f_d$ (Hz).

It is shown by equation above since received wave *r(t)* received in the mobile station is the synthesis of the above-mentioned incoming waves, when the incoming wave number is made to be *N*.

$$r(t) = \sum_{n=1}^{N} r_n(t)$$

$$= R_e[(\sum_{n=1}^{N} e_n(t))\exp j\,(2\pi f_c t)]$$

$$= R_e[(x(t) + jy(t))(cos2\pi f_c t + jsin2\pi f_c t)]$$

$$= x(t)cos2\pi f_c t - y(t)sin2\pi f_c t \tag{5.5}$$

Where $x\ (t)$ and $y\ (t)$ are

$$\mathrm{x(t)} = \sum_{n=1}^{N} x_n(t)$$

$$\mathrm{y(t)} = \sum_{n=1}^{N} y_n(t) \tag{5.6}$$

and are normalized random processes, having an average value of σ and dispersion of S , when *N* is large enough. The combinational probability density $\rho(x, y)$ is then given by the equation bellow, where *x* = *x(t)* and *y* = *y(t)*.

$$\rho(x, y) = \frac{1}{2\pi\sigma^2} \exp\left(\frac{x^2+y^2}{2\sigma^2}\right) \tag{5.7}$$

In addition, it can be expressed as *r(t)* using the amplitude and phase of the received wave.

$$r(t) = R(t)\cos\left(2\pi f_c t + \theta(t)\right) \tag{5.8}$$

$R(t)$ and $\theta(t)$ are given by

$$R(t) = R = \sqrt{x^2 + y^2} \tag{5.9}$$

$$\theta(t) = \theta = \tan\left(\frac{y}{x}\right) \tag{5.10}$$

By using a transformation of variables $\rho(x, y)$ can be converted into $\rho(R, \theta)$.

$$\rho(R, \theta) = \frac{R}{2\pi\sigma^2} \exp\left(-\frac{R^2}{2\sigma^2}\right) \tag{5.11}$$

integrating $\rho(R, \theta)$ over θ from 0 to 2π, the probability density function $\rho(\mathrm{R})$ is obtained as

$$\rho(R) = \frac{R}{\sigma^2} \exp\left(-\frac{R^2}{2\sigma^2}\right) \tag{5.12}$$

By integrating $\rho(R, \theta)$ over R from 0 to ∞, the probability density function $\rho(\mathrm{R})$ is obtained as

$$\rho(\theta) = \frac{1}{2\pi} \tag{5.13}$$

From these equations, the envelope fluctuation follows a Rayleigh distribution, and the phase fluctuation follows a uniform distribution on the fading in the propagation path.

An expression for simulations of this Raleigh fading is found. Here, the mobile station receives the radio wave as the arrival angle of the receiving incoming wave is uniformly distributed, and the wave number of the incoming waves is *N*.

$$r(t) = x(t) + jy(t)$$

$$= \left[\sqrt{\frac{2}{N_1+1}} \sum_{n=1}^{N_1} \sin\left(\frac{\pi n}{N_1}\right) \cos\left\{2\pi f_d \cos\left(\frac{2\pi n}{N_1}\right) t\right\} + \frac{1}{\sqrt{N_1+1}} cos(2\pi f_d t)\right]$$
$$+ j\sqrt{\frac{2}{N_1}} \sum_{n=1}^{N_1} sin\left(\frac{\pi n}{N_1}\right) cos\left\{2\pi f_d \cos\left(\frac{2\pi n}{N_1}\right) t\right\}$$

(5.14)

Where N_1 is an odd number and is given by

$$N_1 = \frac{1}{2}\left(\frac{N}{2} - 1\right) \quad (5.15)$$

In this case, the following relations are satisfied

$$E[{x_1}^2(t)] = E\left[{y_\varrho}^2(t)\right] = \frac{1}{2}$$

$$E\left[x_1(t)\, y_\varrho(t)\right] = 0 \quad (5.16)$$

5.6 Multipath Fading :

Multipath Fading is simply a term used to describe the multiple paths a radio wave may follow between transmitter and receiver. Such propagation paths include the ground wave, ionospheric refraction, reradiation by the ionospheric layers, reflection from the earth‟s surface or from more than one ionospheric layer, and so on.

Multipath fading occurs when a transmitted signal divides and takes more than one path to a receiver and some of the signals arrive out of phase, resulting in a weak or fading signal. Some transmission losses that effect radio wave propagation are ionospheric absorption, ground reflection and free space losses. Electromagnetic interference (EMI) both natural and man made, interfere with radio communications. The maximum useable frequency (MUF) is the highest frequency that can be used for communications between two locations at a given angle of incidence and time of day. The lowest usable frequency (LUF) is the lowest frequency that can be used for communications between two locations.

5.6.1 Multipath Channel Characteristics :

Because there are obstacles and reflectors in the wireless propagation channel, the transmitted signal arrivals at the receiver from various directions over a multiplicity of paths. Such a phenomenon is called multipath. It is an unpredictable set of reflections and/or direct waves each with its own degree of attenuation and delay.

Multipath is usually described by: Line-of-sight (LOS): the direct connection between the transmitter (TX) and the receiver (RX). Non-line-of-sight (NLOS): the path arriving after reflection from reflectors. The illustration of LOS and NLOS is shown in Figure 5.4

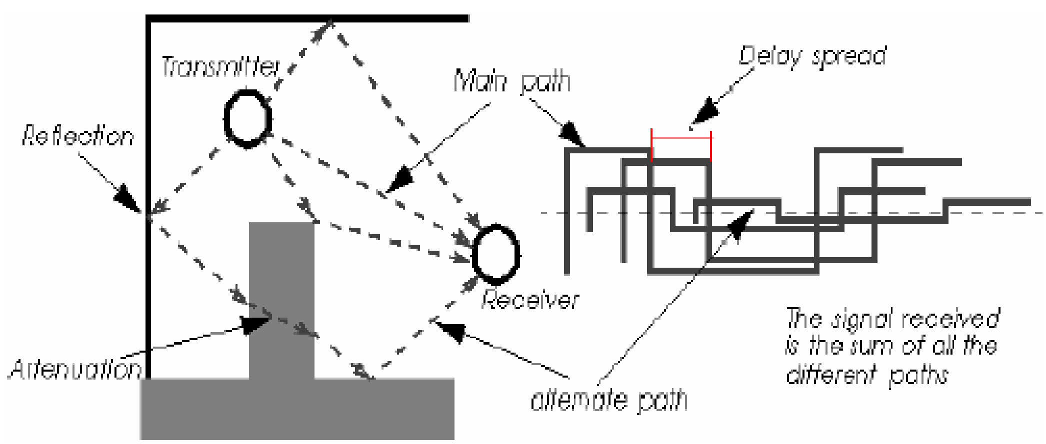

Fig 5.4 Effect of multipath on a mobile station

Characteristics of a Multipath Channel are:

1. Delay spread – this is the interval for which a symbol remains inside a multipath channel
2. Channel can be modeled as a FIR filter with one line of sight (LOS) path & several multipaths, the signals from the multipath being delayed and attenuated version of the signal from the LOS path.

Chapter 6

Simulation results and discussion

6.1 Introduction :

In this chapter, the simulation model used in this thesis is given and the results obtained by using the Block Turbo code as forward error control code for the OFDM system are shown and improvement in performance in terms of bit error rate (BER) are discussed. The Block Turbo coded OFDM (BTC – OFDM) system Performance is evaluated by several numerical simulations in three different channels including AWGN**,** Rayleigh and a multipath channel with three taps. The Doppler shift (f_d)for Rayleigh fading channel here is taken as100 Hz. The effect of the Cyclic prefix length in BTC-OFDM system is also investigated. Both advantages of OFDM and Turbo codes are studied separately to show the advantages of BTC-OFDM system. The specification of the proposed BTC-OFDM scheme is given in Table 6.1.

It is assumed that the channel state information is available at the receiver. It is also assumed that high power amplifier at the transmitter is ideal and the frequency and symbol timing are perfectly synchronized.

6.1.1 Matlab simulation model :

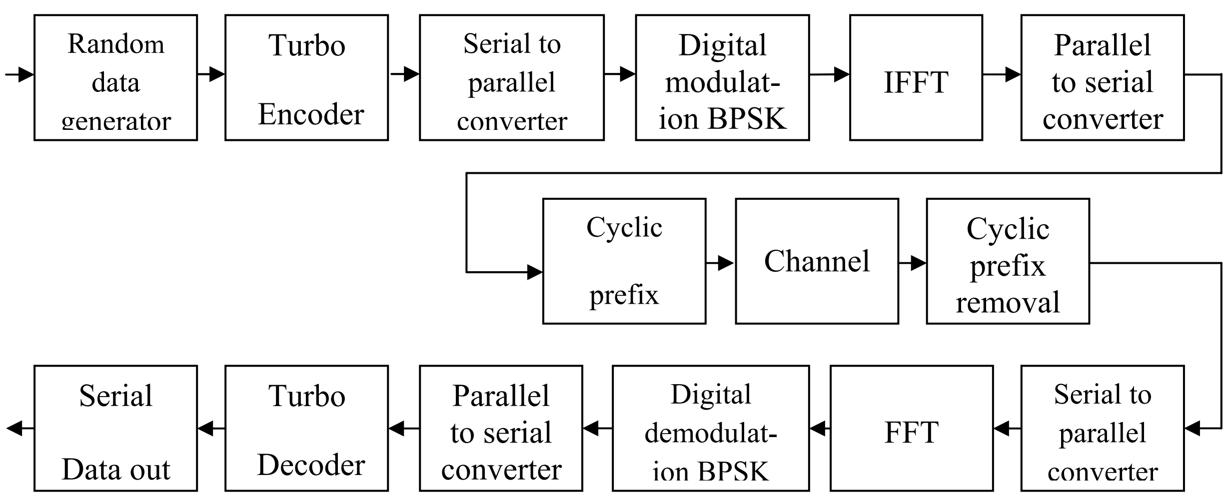

Fig 6.1 Turbo coded OFDM model used for simulation

6.1.2 Simulation parameters :

The simulation parameters used in this thesis are given in the following Table 6.1 and Table 6.2.

Parameters	**Values**
Error correcting Code	Block Turbo Code
Modulation	BPSK
channel	AWGN, Rayliegh fading
Interleaver	Pseudo-random
Number of data sub carriers	48
Number of pilot sub carriers	4
Number of total sub carriers	52
FFT size	48
Coding rate	1/2
Sub carrier frequency spacing	0.3125 MHz(20 MHz / 64)
T_{FFT} (=) : IFFT / FFT period	3.2 μsec
T_{GI} (= T_{FFT}/4): Guard interval time	0.8 μsec
T_{signal} (= T_{FFT} + T_{GI}) : Signal duration	4.0 μsec
T_{symbol} : Symbol interval	4.0 μsec
Iterations	4 , 5

Table 6.1 BTC – OFDM Simulation parameters

6.2 BER performance of OFDM for different modulations over AWGN channel :

Figure 6.2 shows the BER performance of uncoded OFDM for BPSK, QPSK and 16 QAM digital modulations in accordance with the simulation parameter shown in Table 6.2.

parameters	values
Digital modulation	BPSK, QPSK, 16 QAM
FFT size	48
subcarrier	48
Channel	AWGN

Table 6.2 Simulation parameters of uncoded OFDM for differentmodulations over AWGN

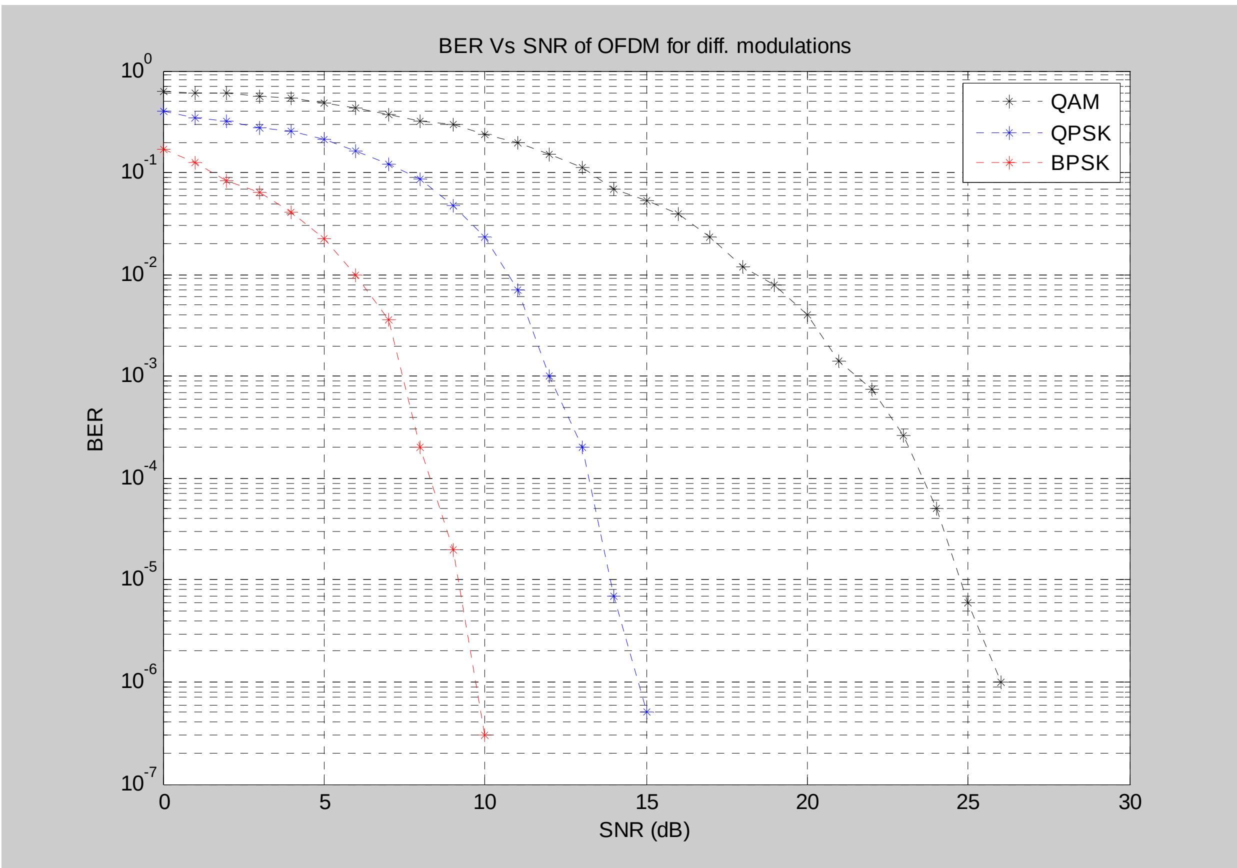

Fig 6.2 BER comparison of uncoded OFDM for different modulations

It was found that the SNR performance of OFDM is similar to a standard single carrier digital transmission. This is to be expected, as the transmitted signal is similar to a standard Frequency Division Multiplexing (FDM) system. Figure 6.2 shows the results from the simulations. The results show that using QPSK the transmission can tolerate a SNR of > 10-12 dB. The bit error rate BER gets rapidly worse as the SNR drops below 6 dB. However, using BPSK allows the BER to be improved in a noisy channel, at the expense of transmission data capacity. Using BPSK the OFDM transmission can tolerate a SNR of > 6-8 dB. In a low noise link, using QAM can increase the capacity. If the SNR is > 25 dB QAM can be used, doubling the data capacity compared with QPSK.

Table 6.3 gives the SNR Comparisons of uncoded OFDM simulation for different digital modulations at BER of 10^{-6} dB and 10^{-5} dB.

Modulation	**SNR for 10^{-6} BER in dB**	**SNR for 10^{-5} BER in dB**
BPSK	~ 9.5	~ 9.0
QPSK	~ 14.5	~ 13.5
QAM	~ 26.5	~ 23.5

Table 6.3 SNR comparison of uncoded OFDM under AWGN channel

6.3 BER performance of Block Turbo code under AWGN channel for different iterations :

Figure 6.3 compares the BER performance of Turbo code under AWGN channel for different iterations. From the result it can be observed that, as the number of iterations increases BER performance increases. Here it is simulated up to 5 iterations and verified that additional number of iterations does not show noticeable difference. The power of the scheme came from the two individual decoders performing the MAP on interleaved versions of the input. Each decoder used information produced by the other as a priori information and outputted a posteriori information. In present wireless scenario, turbo codes are preferred in UMTS third-generation cellular standard, as standardized by the Third-Generation Partnership Project.

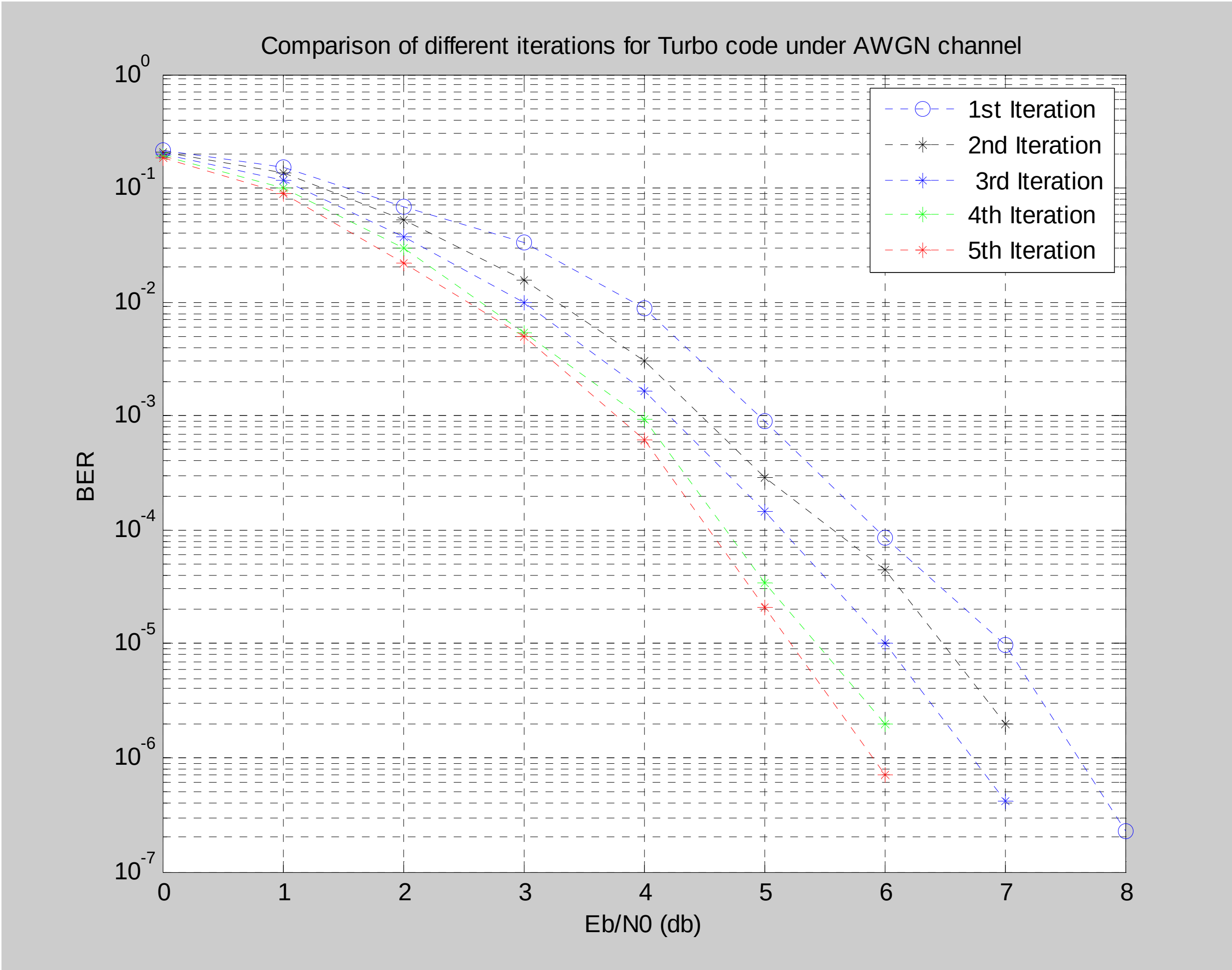

Fig 6.3 Comparison of different iterations for Turbo code under AWGN channel

Table 6.4 shows the simulation parameters of Turbo code simulation under AWGN channel.

Parameters	**values**
Digital modulation	BPSK
Code Rate	1/3
Interleaver	Psuedo-random
Channel	AWGN
Iterations	5

Table 6.4 simulation parameters of Turbo code under AWGN channel

6.4 Performance of BTC – OFDM over AWGN and Rayleigh fading channel :

Figure 6.4 shows the BER performance results of BTC – OFDM over AWGN and Rayleigh fading channels for one iteration and five iterations. These results show that in AWGN channel the BER of the BTC-OFDM system as expected is the same as the BER of single carrier BTC system. BTC – OFDM over Rayleigh fading channel achieves a gain of 7.2 dB at BER of 10^{-5} taking consideration of five iterations.

Table 6.5 shows the simulation parameter of uncoded and BT – COFDM under Rayleigh fading channel with different Doppler shift values for BPSK modulation.

Parameters	**values**
Digital modulation	BPSK
FFT size	48
Turbo code rate	1/3
Channel	AWGN , Rayleigh fading with f_d = 100 Hz

Table 6.5 Simulation parameters of BTC – OFDM for AWGN and Raleigh fading channel with Doppler shift f_d = 100 Hz.

Table 6.6 gives the Eb/N0 Comparisons of Block Turbo coded OFDM simulation under AWGN and Rayleigh fading channels for one iteration and five iterations at BER of 10^{-6} dB and 10^{-5} dB.

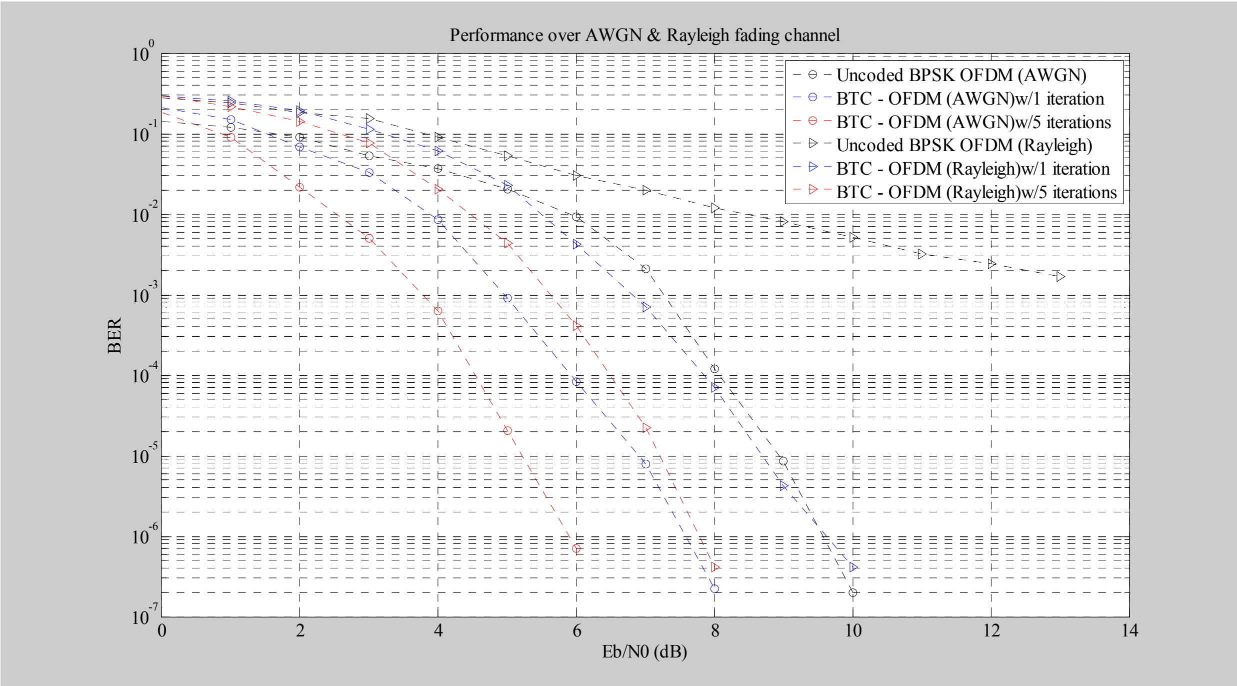

Fig 6.4 Performance of BTC – OFDM over AWGN and Rayleigh fading channel.

Modulation	**Eb/N0 for 10^{-6} BER in dB**	**Eb/N0 for 10^{-5} BER in dB**
Uncoded BPSK OFDM(AWGN)	~ 9.5	~ 9.0
Block Turbo coded OFDM(AWGN) with 1 iteration	~ 7.6	~ 7.0
Block Turbo coded OFDM (AWGN) with 5 iterations	~ 6.0	~ 5.0
Uncoded BPSK OFDM(Rayleigh)	> 14.0	> 14.0
Block Turbo coded OFDM(Rayleigh) with 1 iteration	~ 9.5	~ 8.8
Block Turbo coded OFDM (Rayleigh) with 5 iterations	~ 7.8	~ 7.2

Table 6.6 Eb/N0 Performance of BTC – OFDM over AWGN and Rayleigh fading channel.

6.5 Comparison between BTC - OFDM and Convolutional coded OFDM :

BTC – OFDM offers a good compromise between performance and complexity comparable to those of CTC – OFDM and is very attractive for implementation.

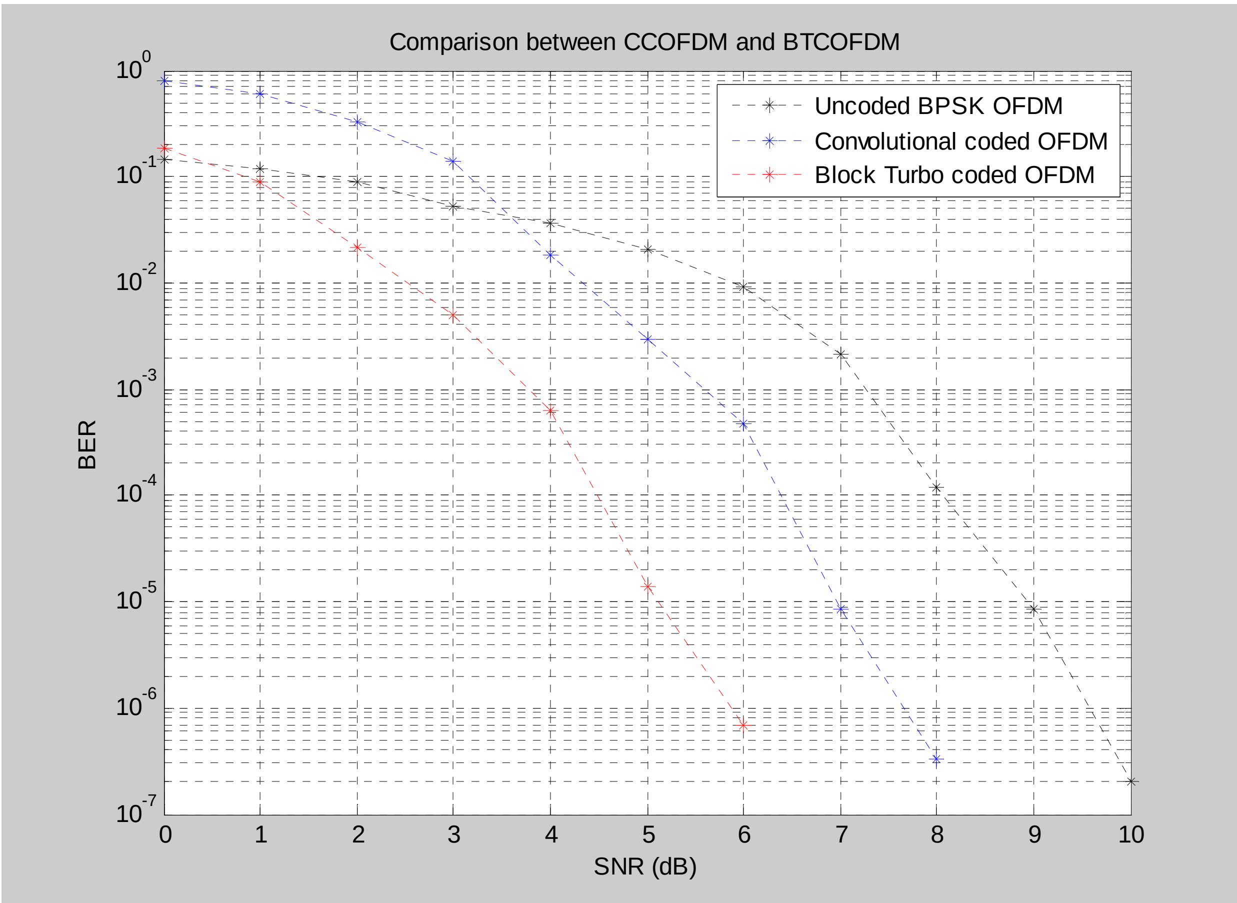

Fig 6.5 Comparison between Block Turbo coded OFDM and Convolutional coded OFDM

Figure 6.5 shows the comparison between Block Turbo coded OFDM and Convolutional coded OFDM and the SNR comparison is summarized in Table 6.7

Modulation	SNR for 10^{-6} BER in dB	SNR for 10^{-5} BER in dB
Uncoded BPSK OFDM	~ 9.5	~ 9.0
Convolutional coded OFDM	~ 7.6	~ 7.0
Block Turbo coded OFDM	~ 6.0	~ 5.0

Table 6.7 SNR comparison between BTC – OFDM and Convolutional coded OFDM

6.6 Performance of BTC – OFDM over multipath fading channel :

Figure 6.6 shows the performance of the BTC-OFDM system over a three tap multipath fading channel. As can be seen, BTC-OFDM provides ~ 8.8 dB gain at BER 10^{-5} over single carrier BTC. Here uncoded OFDM performance for no cyclic prefix (Tg = 0) and a cyclic prefix equal to channel spread (Tg/Td = 1), where Tg is the guard time (cyclic prefix) and Td is the channel delay spread. It can be seen that the difference in performance can be very large, particularly at high Eb/N0, where lack of cyclic prefix results in the loss of performance. Use of Turbo codes can improve the performance and overcome residual Inter symbol Interference (ISI) and Inter carrier Interference (ICI).

Specification of this multipath channel which is a channel model for WLAN, is shown in Table 6.8

Tap	Relative delay (n sec)	Average Power (dB)
0	0	0
1	50	-3.6
2	100	-7.2

Table 6.8 Multipath channel model parameters

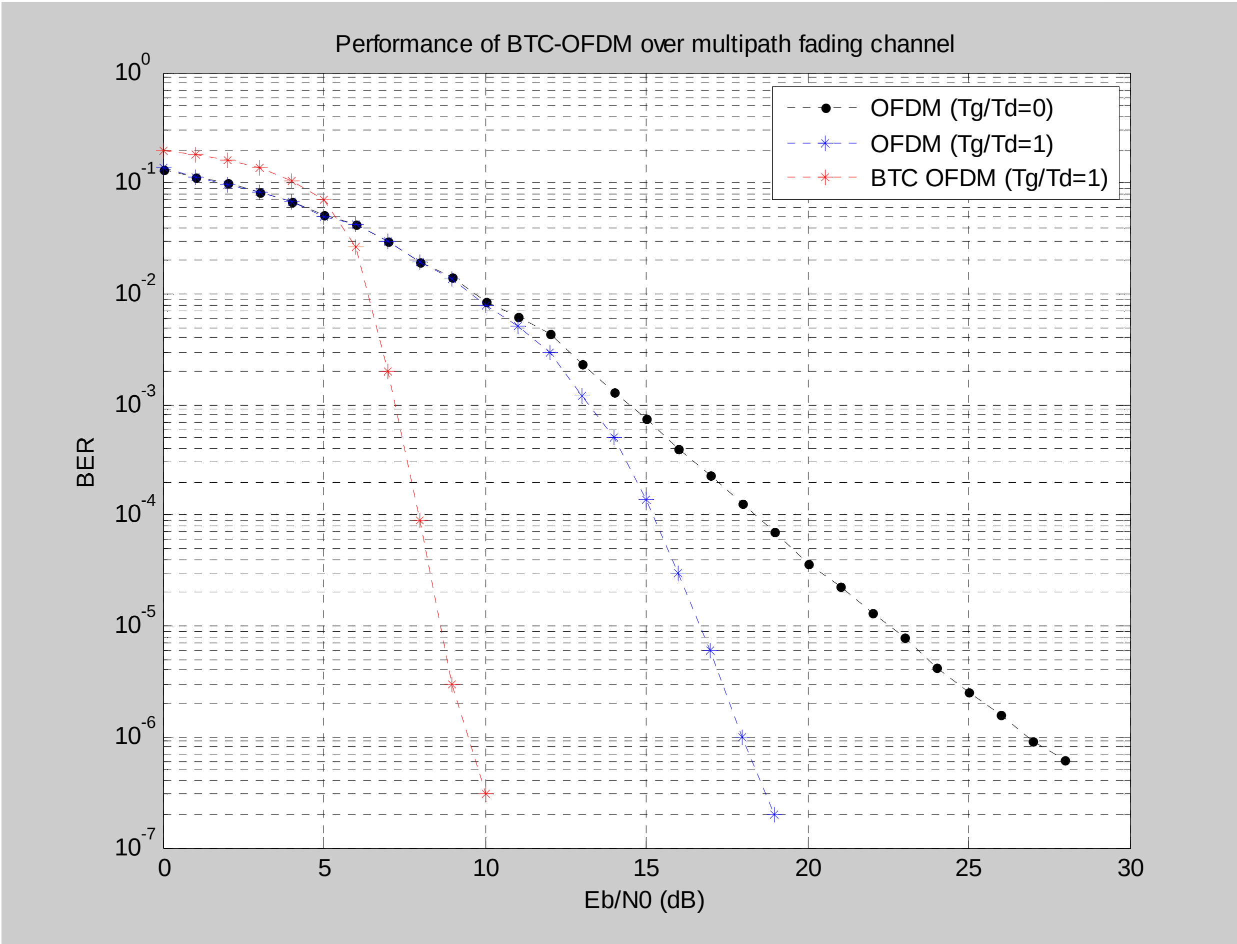

Fig 6.6 Performance of BTC – OFDM over multipath fading channel

The Performance of BTC – OFDM over multipath fading channel is summarized in Table 6.9 .

Modulation	**Eb/N0 for 10^{-6} BER in dB**	**Eb/N0 for 10^{-5} BER in dB**
OFDM (Tg/Td = 0)	~ 27.0	~ 22.5
OFDM (Tg/Td = 1)	~ 18.0	~ 16.8
BTC OFDM (Tg/Td = 1)	~ 9.5	~ 8.8

Table 6.9 Performance of BTC – OFDM over multipath fading channel

6.7 Effect of Cyclic Prefix length in BTC – OFDM over a multipath fading channel :

Figure 6.7 shows the BER performance results of the BTC – OFDM system for different lengths of cyclic prefix. In the first case, no cyclic prefix is used (Tg/Td = 0**)** and in the second case, the length of cyclic prefix is equal to half of the channel delay spread (Tg/Td = 0.5), and in the last case the length of cyclic prefix is equal to the channel delay spread (Tg/Td = 1.0). It can be concluded that Turbo codes allow the use of a shorter Cyclic Prefix and, thereby, a reduction in the overhead associated with the cyclic prefix used in an OFDM system

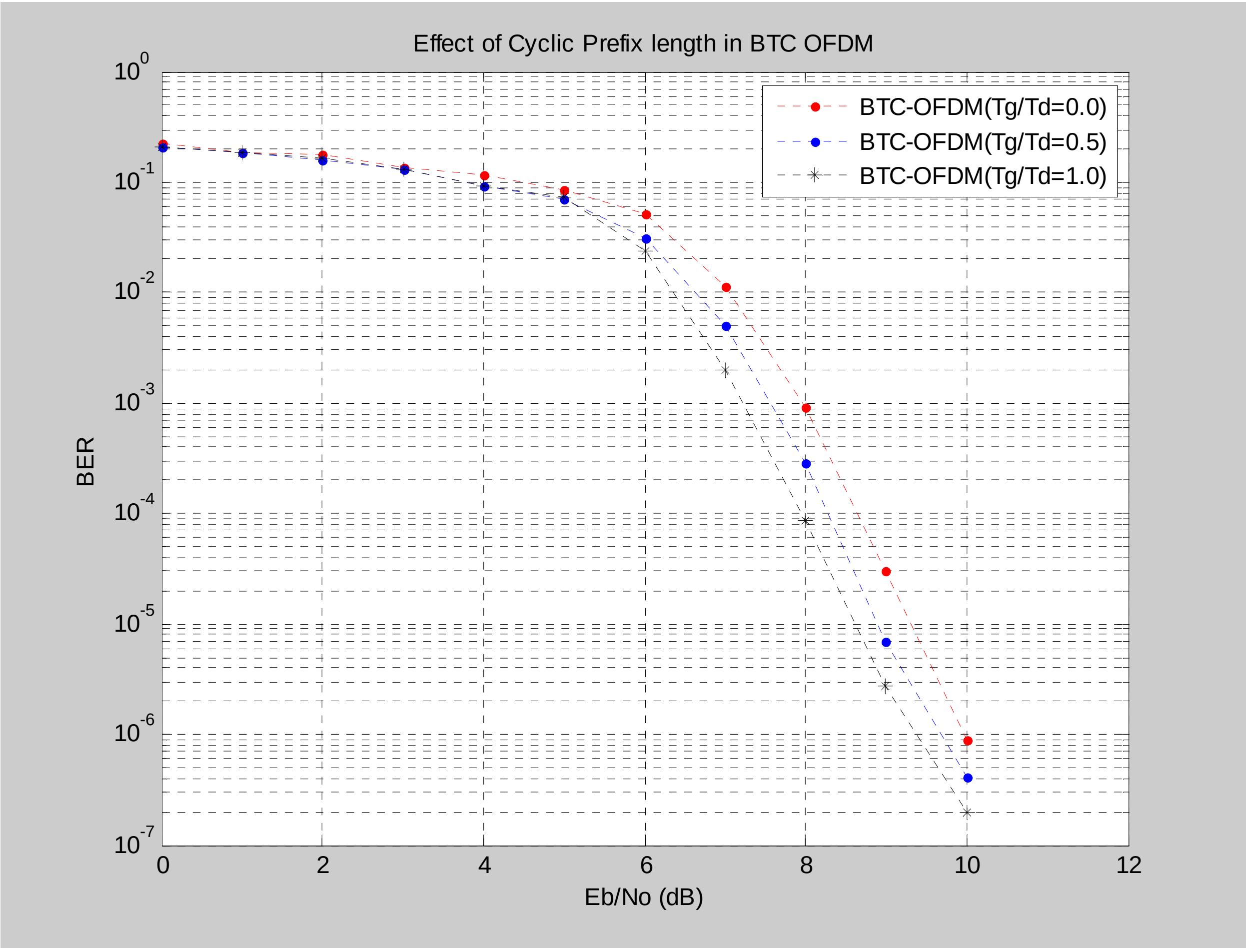

Fig 6.7 Effect of Cyclic Prefix length in BTC – OFDM over a multipath fading channel

The effect of Cyclic Prefix length in terms of Eb/N0 performance in BTC – OFDM over a multipath fading channel is shown in the Table 6.10.

Modulation	**Eb/N0 for 10^{-6} BER in dB**	**Eb/N0 for 10^{-5} BER in dB**
BTC – OFDM (Tg/Td = 0.0)	~ 10.0	~ 9.3
BTC – OFDM (Tg/Td = 0.5)	~ 9.5	~ 9.0
BTC – OFDM (Tg/Td = 1.0)	~ 9.3	~ 8.5

Table 6.10 Eb/N0 performance of effect of cyclic prefix length in BTC – OFDM over a multipath fading channel.

Chapter 7

Conclusion

7.1 Conclusion :

In this thesis BTC-OFDM uses the diversity of both t i e and frequency domains to achieve higher performance and more robustness against frequency and time selective fading. The benefits of coding gain of Turbo codes over Convolutional codes, provide an improvement in performance and efficiency of the proposed scheme over IEEE802.11a. Therefore, BTGOFDM has superior performance and efficiency over EEE802.11a standard.

The performance analysis of the BTC-OFDM system is evaluated by simulations in different channels including AWGN, Rayleigh and a three tap multipath fading channel. The benefits of using OFDM and Turbo codes are studied separately. It is shown that BTC-OFDM can provide a better performance than single carrier BTC system in fading charnels. On the other hand, Turbo codes can eliminate the residual inter symbol interference (ISI) and inter channel interference (ICI) and therefore reduce the length of the required Cyclic prefix in an OFDM system. This decreases the overhead associated with the Cyclic Prefix. The use of Turbo codes in OFDM system for high data rate transmission in wireless LANs, results in a considerable improvement in terms of bit error rate performance and bandwidth efficiency.

7.2 Scope of future work :

The following are the some of the interesting extensions of the present work:

1. An interesting topic for future research is to perform more extensive performance comparisons between FFT based BTC – OFDM and DCT based BTC – OFDM systems under additional real-world channel impairments, such as multipath fading, time dispersion which leads to inter symbol interference (ISI).

2. The main problems with OFDM signal is very sensitive to carrier frequency offset, and its high Peak to Average Power Ratio (PAPR). So, BTC – OFDM systems can be tested for these problems.

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