Spectral Efficiency and BER Analysis In 802.16e Standard

A thesis submitted in partial fulfilment of the requirements for the award of the Degree of

MASTER of ENGINEERING

in

ELECTRONICS AND COMMUNICATION ENGINEERING

Submitted by:

SHYAM PRATAP SINGH
Roll No. 800861013

Under the guidance of:

Dr. KULBIR SINGH
Assistant Professor, ECED

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

THAPAR UNIVERSITY

PATIALA-147004, Punjab, INDIA.

June 2010

CERTIFICATE

I, Shyam Pratap Singh, hereby certified that the work which is being presented in this thesis entitled "*Spectral Efficiency and BER Analysis in 802.16e Standard*" in partial fulfillment of requirements for the award of degree of Master of Engineering in Electronics and Communication from Thapar University, Patiala, is an authentic record of my own work carried under the supervision of Dr. Kulbir Singh.

The matter presented in this thesis has not been submitted in any other University/Institute for the award of Master of Engineering.

(Shyam Pratap Singh)
Signature of the student

This is certified that the above statement made by the candidate is correct to the best of my knowledge

(Dr.Kulbir Singh)
Supervisor

(Dr. A.K. Chaterjee) 9.7.10
Head of Department
ECED, T.U. Patiala

(Dr. R.K. Sharma) 12.7.10
Dean of Academic Affairs
T.U. Patiala

ACKNOWLEDGEMENT

A good job is never the outcome of the efforts of a single person. I feel myself lucky to express my profound sense of gratitude and respect to all those who helped directly or indirectly me throughout this thesis.

First, I would like to thank the **"BABA JI"** the Supreme Power, without his hidden grace, I am nothing.

Again, Supreme Power sends his embodiment in form of "Guide". This work would not have been possible without the continuous supervision and able guidance of my guide **Dr. Kulbir Singh,** his enthusiasm and optimistic approach of guidance will be unforgettable elements of my life. At the same time, I am grateful to Head of the Department **Dr. A. K. Chatterjee** for providing the facilities for the completion of thesis.

At last but not the least I would like to thank my parents, family members and friends whose blessings and wishes are always with me.

Also, I would like to thank to all the members of Electronics and Communication Department whose love and affection made my stay at T.U. campus a memorable.

Place: T.U. Patiala, India (SHYAM PPRATAP SINGH)

ABSTRACT

Spectrum being nature's gift; needs most wise use of it. OFDM system is an excellent way to utilize the spectrum. OFDM is a parallel data transmission system which promises the high data rate with minimum degradation of the quality of service relative to serial communication techniques. In OFDM, a single channel utilizes multiple sub-carriers on adjacent frequencies. In addition the sub-carriers in an OFDM system are overlapping to maximize spectral efficiency. Ordinarily, overlapping adjacent channels can interfere with one another. However, sub-carriers in an OFDM system are precisely orthogonal to one another. Thus, they are able to overlap without interfering. As a result, OFDM systems are able to maximize spectral efficiency without causing adjacent channel interference. In order to obtain the orthogonality the subcarrier frequencies must be spaced by a multiple of the inverse of symbol duration. Other than spectral efficiency OFDM has many advantages, as immunity to impulse interference, resilient to RF interference, robustness to channel fading, resistance to multipath, much lower computational complexity(collectively, they reduces bit error rate).

Also, WiMAX being one of the standards which is using OFDM, has excellent spectral efficiency but there are no limits for researcher's, that is inspiration for "spectral-efficiency and BER analysis in 802.16e". As we know that *with "differential -modulation" over either time samples (Doppler problem) or sub carrier (tone correlation & phase rotation problem) cause a problem. So the suitable modulation techniques are tradeoff between SPECTRAL EFFICIENCY & BER, in simple words as level of modulation is increased BER get degraded but spectral-efficiency get improved. However, an improvement in BER can be achieved using combinations of different FEC coding but at the cost of spectral-efficiency. In present work RS and CC combination has been implemented for BER improvement, without much loss in spectral efficiency.*

In wireless system, neither spectrum nor BER can be compromised at any cost. The tradeoff between modulation techniques and FEC is desirable. *Hence, both the trade-off, the effect of modulation level and the effect of FEC have been analyzed in thesis work. Also, effect of CP addition has been analyzed. For this purpose different channels starting from ideal one i.e AWGN, after that Rayleigh and then finally Nakagami have been taken for analysis.*

TABLE OF CONTENTS

Certificate	i
Acknowledgement	ii
Abstract	iii
Table of Contents	iv
List of Figures	viii
List of Tables	x
List of Abbreviation	xi
CHAPTER 1: INTRODUCTION	**1-10**
1.1 Introduction to OFDM	1
1.1.1 OFDM Basics	1
1.1.2 Maximizing the overall spectral efficiency	*2*
1.1.3 User allocation	*2*
1.2 Mathematical aspect	*3*
1.3 STANDARD 802.16a UPTO 802.16e	*5*
1.3.1 Step by Step Development	5
1.3.2 IEEE 802.16 Protocol Layers	8
1.3.3 Application of IEEE 802.16 based network	8
1.4 Objective of Thesis	9
1.5 Organization of Thesis	9
CHAPTER 2: LITERATURE REVIEW	**11-15**
CHAPTER 3: BER & SPECTRAL EFFICIENCY ANALYSIS	**22-44**
3.1 Physical layer consideration for WiMAX	22
3.2 Overall-all System Model	25
3.2.1 Randomization	25
3.2.2 Forward Error Correction (FEC)	26

3.2.2.1 Convolution Coding 26
3.2.2.2 Connection Representation 27
3.2.2.3 Polynomial Representation 28
3.2.2.4 State Representation and the State Diagram 29
3.2.2.5 The Tree Diagram 30
3.2.2.5 The Trellis Diagram 31
3.2.2.7 The Viterbi Convolutional Decoding Algorithm 33
3.2.2.8 Reed Solomon Codes (RS codes) 33
3.2.2.9 Arithmetic in $GF\ (\mathbf{2}^{r})$ 34
3.2.3 Interleaving 35
3.2.4 Subcarrier modulation and mapping [5-7] 36
3.2.4.1 Binary Phase Shift Keying (BPSK) 37
3.2.4.2 Quadrature Phase Shift Keying (QPSK) 37
3.2.4.3 16-Quadrature Amplitude Modulation 37
3.2.5 Pilot subcarriers 38
3.2.6 IFFT and FFT 38
3.2.7 Cyclic Prefix Addition 38
3.3.1 BER Analysis for different channels and modulation schemes 38
3.3.2. Mathematical Analysis for B.E.R 39
3.3.2.1 BPSK Modulation 39
3.3.2.2 AWGN Channel 40
3.3.2.3 Rayleigh Fading Channel 40
3.3.2.4 QPSK Modulation 41
3.3.2.5 AWGN Channel 41
3.3.2.6 Rayleigh Fading Channel 42
3.3.2.7 QAM Modulation 42
3.3.2.8 Rayleigh Channel 43
3.3.2.9 AWGN Channel 43
3.4 Spectral-efficiency analysis 44

CHAPTER 4: FADING CHANNEL **45-45**

4.1 Fading-definition 45

4.1.1 Multi path fading 45

4.1.2 Types of multi path fading 46

4.1.2.1 Large-Scale Fading 46

4.1.2.2 Small - Scale Fading 47

4.2. Different Channel Models 54

4.3 Simulation Model of Nakagami Channel 56

CHAPTER 5: Results and Discussion **57-63**

CHAPTER 6: CONCLUSION **64**

REFERENCES **65-70**

-

LIST OF FIGURES

NO.	TITLE	PAGE NO.
1.1	Block Diagram of OFDM System	4
2.1	History of OFDM System	11
2.2	Comparison of PAN, WLAN & WiMAX	15
2.3	Comparison of WiMAX & UMTS	16
3.1	Physical layer (802.16)	25
3.2	Convolutional encoder	26
3.3	State diagram for encoder	29
3.4	Tree diagram	30
3.5	Trellis Diagram	32
4.1	Multi Path Fading Channel	45
4.2	Large Scale & Small Scale Fading	47
4.3	Flat Fading Channel Characteristics	50
4.4	Frequency Selective Fading channel Characteristics	51
4.5	Type of fading experienced by a signal as a function T_s and B_s	54
5.1a	BER for AWGN with CC	57
5.1b	BER for AWGN with CC-RS	57
5.2a	Spectral efficiency for AWGN with CC	58
5.2b	Spectral efficiency for AWGN with CC-RS	58
5.3a	BER for Rayleigh with CC-RS	59
5.3b	BER for Rayleigh with CC	59
5.4a	Spectral efficiency for Rayleigh	60
5.4b	Spectral efficiency for Rayleigh CC-RS	60
5.5a	BER for Nakagami with CC	61
5.5b	BER for Nakagami with CC-RS	61
5.6a	Spectral efficiency for Nakagami	62
5.6b	Spectral efficiency for Nakagami CC-RS	62
5.7a	Effect of CP on BASK	63
5.7b	Effect of CP on 64QAM-3/4	63

LIST OF TABLES

TABLE NO.	TITLE	PAGE NO.
2.1	Summery of different standards	7
3.1	Comparison of Both the Standards	24
4.1	Types of Small Scale Fading	48

List of Abbreviations

ADC	Analog to Digital Conversion
ADSL	Asymmetric Digital Subscriber Line
ATM	Asynchronous Transfer Mode
BER	Bit Error Rate
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
QPSK	Quardrature Phase Shift Keying
BS	Base Station
BWA	Broadband Wireless Access
RS	Reed-Solomon
CC	Convolutional Code
CP	Cyclic Prefix
DAC	Digital to Analog Conversion
DFT	Discrete Fourier Transform
DSL	Digital Subscriber Line
ETSI	European Telecommunications Standards Institute
FDD	Frequency Division Duplexing
FDM	Frequency Division Multiplexing
FEC	Forward Error Correction
FFT	Fast Fourier Transform
HIPERMAN	High PERformance Metropolitan Area Network
ICI	InterCarrier Interference
IDFT	Inverse Discrete Fourier Transform
IEEE	Institute of Electrical and Electronics Engineers
ISI	InterSymbol Interference
LAN	Local Area Network
LOS	Line of Sight
MAC	Medium Access Control
MCM	MultiCarrier Modulation
NLOS	Non Line of Sight

OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
QPSK	Quadrature PhaseShift keying
SNR	Signal to Noise Ratio
SS	Subscriber Stations
TDD	Time Division Duplexing
TDM	Time Division Multiplexing
TDMA	Time Division Multiple Access
WAN	Wide Area Network
WiMAX	Worldwide Interoperability for Microwave Access
WirelessMAN	Wireless Metropolitan Network

CHAPTER 1 - INRODUCTION

1.1 Introduction to OFDM system

1.1.1 OFDM Basics

The use of multiple antennas at both ends of w/l links holds the potential to drastically improve the spectral efficiency and link reliability. [1,2] on the other hand OFDM technology can transform frequency selective channel into a set of parallel frequency-flat channel [3]. The combination of both is the best channel utilization through-out all the available w/l system if FRFT is used with proper use of optimal basis function. However in conventional system and in our present discussion also FFT is used.

So in this section we will take the introductory length and width of OFDM in detail. The OFDM is known as ORTHOGONAL FDM, what its mean, in short FDM is technology used for sharing of B.W. by sub-signal of low rate of a given signal. This B.W. sharing may cause interference between different signals. So next step to avoid interference is ORTHOGONALITY. [4] How it does? For non- ideal channel if (1/information rate) > time dispersion of non-ideal channel i.e. duration of impulse response, there will be no interference of symbols, implies that single carrier (channel) of high rate can have more & more interference so get out of this problem we need low information rate i.e. divide information into sub-information then transmit them using sub-carrier .But at the same time being large no. of sub-carrier causes interference between them, called as I.C.I. So what next? Answer is orthogonality between sub-carriers, finally, we get OFDM System [5] being no. of sub-carrier the synchronization is essential parameter for faithful reception. In broad-sense, two way to achieve synchronization. First is to have coherent B.W., which is inversely proportional to Doppler spreading and is statically measure of the range of frequency over which the channel can be considered “flat” (equal gain & linear phase.) Secondly, coherent time, which is again inversely proportional to Doppler spread & is statically measure of the time duration over which channel impulse response is *invariant. On this channel are classical as large scale fading-path loss and small scale fading–path loss. To maintain coherence, we use "coherent modulation" such as X-PSK OR X-QAM etc, as a tail of it requires channel estimation .However with "differential -modulation" over either time samples a (Doppler problem) or sub carrier(tone correlation & phase rotation problem)[3] So the suitable modulation techniques are X-PSK & X-QAM tradeoff between SPECTRAL EFFICIENCY & B.E.R. In a multi- path radio channel frequency selective fading can result in large variation can be as much 30db in received power resulting in similar variation in S.N.R. Again modulation tech. may be fixed modulations, which under worst condition offer an acceptable B.E.R. e.g. BPSK & QPSK with ηspectral = 1-2bps/Hz but provide an excess link margin most of the time, on other hand with adaptive modulation ηspectral increases. E.g., ηspectral =4-6 bps/Hz using 16QAM-64QAM. Again in adaptive modulation we overhead information need to be transmitted as both Tx and Rx must know what modulation currently being used, however using additive modulation the carrier modulation is matched to the S.N.R.*

1.1.2 Maximizing the overall spectral efficiency

Distortion, frequency error and maximum allowable power variation between user limit the maximum modulation technique that can be used the received power for neighboring carrier must have no more than 15-30dB variations at the base station, as Large variations can result in strong signals swamping weaker carriers, similarly frequency errors in the Tx due to synchronization errors and Doppler shift result in a loss of orthogonality between the carriers, a frequency offset of only one percent of the carrier spacing result in the effective S.N.R. being limited to 30dB.

1.1.3 User allocation

There are several methods for allocating carriers to users. The main of them are grouped carriers, spread-out carrier and adaptive carrier allocation. Simplest out them is group carrier, it minimize inter-user interference due to distortion power level variation and frequency errors, however may have fading, which is partly overcome by frequency hopping. In adaptive frequency hopping carrier hopping is based on channel condition. It also virtually eliminates frequency selective fading.

Thirdly comb spread carrier, in it carrier can be allocated in a comb pattern spreading them over the entire system bandwidth, improves diversity and loss of information by a given user. [6]

1.2 Mathematical aspect

As shown in fig. 8.49   k= w/Δf

Where, k: No. of sub-channel.

W: Total B.W.

Δf: sub-channel B.W.

Now, with each sub-channel we associate a carrier known as sub-carrier i.e.

Xk =sin (2.π.fk.t) where k= 0, 1, 2, 3… K-1 and fk mid frequency in k^{th} *sub-channel.*

fk = kf1 or equivalently maintaining 1/T (inverse of symbol rate of each sub-channel)

However conventional orthogonality is defined as; for digital communication

In time domain

$$\int_{T_s} x_p(t)x_q^*(t)dt = \begin{cases} 0 \;\; for & p \neq q \\ G \; for & p = q \end{cases} \qquad (1.1)$$

In frequency domain

$$\int_{T_s} x_p(f)x_q^*(f)df = \begin{cases} 0 \;\; for & p \neq q \\ G \; for & p = q \end{cases} \qquad (1.2)$$

Simplest example of orthogonal system is the integration of product of any combination of sinɵ and cosɵ (excluding sinɵ. sinɵ or cosɵ. cosɵ) as it can be break into sum of two sinusoid and integration over time-period of any sinusoid is always zero. This idea can be generalized as

f2= 2f1, f3 = 3f ………………and so on.

From above discussion we get following which basis of orthogonality condition for OFDM system

1. Each sub carrier has exactly an integer number of cycles in the FFT interval.

2. The number of cycles between adjacent sub carriers differs by exactly one

From above section we conclude that we have to break higher rate data into lower data rate i.e. serial to parallel conversion must be 1st block of OFDM system as shown below,

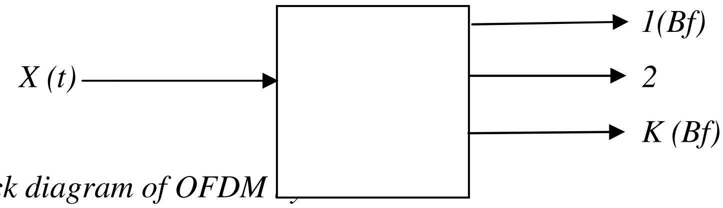
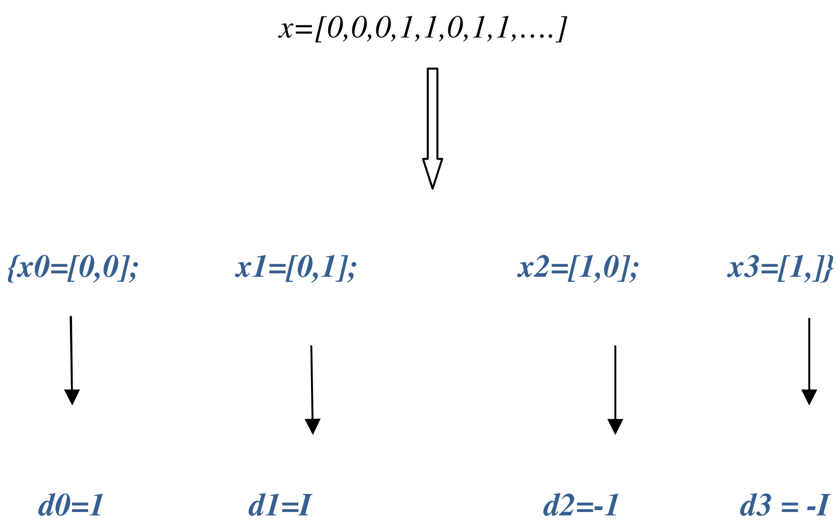


***Figure 1.1**: Block diagram of OFDM system*

Now why, S/P?

Answer is given in previous section which may be state as "each data symbol normally occupies only a small part of available B.W." hence frequency response over each individual channel is relatively flat. As shown in fig Bf bit in each frame parsed into k groups, where i^{th} group is assigned bits $\sum$ i=1k (bi) = Bf and no. of signal points for i^{th} sub-channel is Mi = 2bi

How it can be achieved?



As shown above for s/p conversion is done followed by mapping (last step as X0 is mapped into d0) Now these data can be modulated using separate sub-carrier may be 16-1024 (used presently) and depend on technology[1]. The mathematical tool for this is IFFT [2]. At this point one can easily states that in OFDM system there is SERIES of modulation are there, actually there are three different steps of modulation are

1. *Multiplying by different harmonics of sinusoids(e^{kwt}) corresponding to IFFT*
2. *Modulating the sub-channel (low rate information) corresponding to different MODULATION TECH. (X-QAM. OR X-PSK.*
3. *Modulating the sub-carrier over a single carrier.*

1.3 STANDARD 802.16a UPTO 802.16e

1.3.1. Step by Step Development

802.16e was introduced in 1998[7], by The National Wireless Electronics Systems Test bed (NWEST) of the U.S National Institute of Standards and Technology (NIST) as the result getting "inter-operability" between deferent standards. After-ward there is a lot advancement has been done in this field to develop the standard for fixed and mobile BWA. The IEEE 802.16 standard contains the specification of 1. Physical (PHY) and 2. Medium Access Control (MAC) layers both for BWA. Starting with IEEE802.162001 [8] in 2001 and standard IEEE 802.162004 (2004) in which all advantages of previous standards have been implemented however with the aid of mobility 802.16e was introduced in 2006 named as "WiMAX". This standard specifies the air interface for fixed BWA systems supporting multimedia services in both licensee and un-licensed spectrum [9]. In present section a brief of above development has been discussed.

IEEE 802.162001

It specifies a set of MAC and PHY layer standards used to provide fixed broadband wireless access in a point to point (PTP) or point to multipoint (PMP) topology. The PHY layer uses single carrier modulation in the 10–66 GHz frequency range. Transmission times, durations and modulations are assigned by a Base Station (BS) and shared with all nodes in the network in the form of broadcast. Subscribers need only to hear the base station that they are connected and do not need to listen any other node of the network. Subscriber Stations (SS) has the ability to have variable bandwidth allocation on a burst to burst basis, providing scheduling flexibility. The efficient modulation scheme for this standard is QPSK, 16QAM and 64QAM as. Which can be changed from frame to frame and from SS to SS, depending on the robustness of the connection? Both Time Division Duplexing (TDD) and Frequency Division Duplexing (FDD) as duplexing technique can be used. Differential Quality of Service (QoS) in the MAC Layer is an important feature of 802.162001 [10]

IEEE 802.16c2002

In December 2002, IEEE Standards Board approved amendment IEEE 802.16c [6]. In this amendment detailed system profiles for 10-66 GHz were added and some errors and inconsistencies of the first version of the standard were corrected.

IEEE 8020.16a2003

Multiple physical layer specifications and providing additional physical layer specifications by enhancing the medium access control layer are two additional feature of this version over previous one. This was introduced by IEEE 802.16 working group in January 2003. This amendment added physical layer support for 2-11 GHz. Non Line of Sight (NLOS) operation becomes possible due to inclusion of below 11 GHz range, extending the geographical reach of the network but multipath propagation was the bad consequence of NLOS. To mitigate the consequences, advanced power management technique and adaptive antenna arrays were included in this version. Security was improved in this version, many of privacy layer features were optional which were optional in 802.162001 became mandatory. IEEE 802.16a also mesh topology in addition to PMP has been implemented in it.[11]

IEEE 802.16e2005

This amendment was included in the current applicable version of standard IEEE 802.162004 in December 2005. This includes the PHY and MAC layer enhancement to enable combined fixed and mobile operation in licensed band. The table 2.1 below presents the summery of the step by step advancement of the standard.

	IEEE 802.162001	IEEE 802.16a	IEEE802.16-2004	IEEE 802.16e2005
Completed	December 2001	January 2003	June 2004	December 2005
Spectrum	10-66 GHz	2-11 GHz	2-11 GHz	2-6 GHz
Propagation/channel Conditions	LOS	NLOS	NLOS	NLOS
Bit Rate	Up to 134 Mbps (28 MHz channelization)	Up to 75 Mbps (20 MHz channelization)	Up to 75 Mbps (20 MHz channelization)	Up to 15Mbps (5 MHz channelization)
Modulation	QPSK, 16QAM (optional in UL), 64QAM (optional)	BPSK, QPSK, 16QAM, 64QAM, 256QAM (optional)	256 subcarriers OFDM, BPSK, QPSK, 16QAM, 64QAM, 256QAM	Scalable OFDMA, QPSK, 16QAM, 64QAM, 256QAM (optional)
Mobility	Fixed	Fixed	Fixed/Nomadic	Portable/Mobile

Table 1.1: Summery of different standards

1.3.2 IEEE 802.16 Protocol Layers [9]

The IEEE 802.16 standard is structured in the form of a protocol stack.1 MAC layer 2: Physical Layer. Again the MAC layer is formed with three sub layers: 1. Service Specific Convergence Sub layer (CS) 2. MAC Common Part Sub layer (CPS) and 3. Privacy Sub layer. The MAC CS receives higher level data through CS Service Access Point (SAP) and provides transformation and mapping into MAC Service Data Unit (SDU). MAC SDUs are then received by MAC CPS through MAC SAP. The specification targeted two types of traffic transported through IEEE 802.16 networks: Asynchronous Transfer Mode (ATM) and Packets. Thus, Multiple CS specifications are available for interfacing with various protocols.

The MAC CPS is the core part of the MAC layer, defining medium access method. The CPS provides functions related to duplexing and channelization, channel access, PDU framing, network entry and initialization. This provides the rules and mechanism for system access, bandwidth allocation and connection maintenance. QoS decisions for transmission scheduling are also performed within the MAC CPS. The Privacy layer lies between the MAC CPS and the PHY layer. Security is a major issue for public networks. This sub layer provides the mechanism for encryption and decryption of data transferring to and from PHY layer and is also used for authentication and secure key exchange. However in the present work our main issue is the “physical layer” which is discussed in full detail in section 3.1

1.3.3 Application of IEEE 802.16 based network

IEEE 802.16 supports ATM, IPv4, IPv6, Ethernet and Virtual Local Area Network (VLAN) services SO, it has a wide choice of service possibilities for both voice and data network [12]

Some of them are listed below

1. **Cellular Backhaul**: IEEE 802.16 wireless technology can be an excellent choice for back haul for commercial enterprises such as hotspots as well as point to point back haul applications due to its robust bandwidth and long range.
2. **Residential Broadband**: Practical limitations like long distance and lack off return channel prohibit many potential broadband customers reaching DSL and cable technologies [13]. IEEE 802.16 can fill the gaps in cable and DSL coverage.

3. **Underserved areas:** In many rural areas, especially in developing countries, there is no existence of wired infrastructure. IEEE 802.16 can be a better solution to provide communication services to those areas using fixed CPE and high gained antenna.
4. **Always Best Connected**: As IEEE 802.16e supports mobility [14], so the mobile user in the business areas can access high speed services through their IEEE 802.16/WiMAX enabled handheld devices like PDA, Pocket PC and smart phone.

1.4 Objective of Thesis

The objective of the thesis is to analyze the standard 802.16e, for this two sub-objectives have been taken first is spectral efficiency analysis and second is BER analysis.

First step towards objective is implement to the most ideal channel (AWGN) for our model, for different modulation techniques (*BASK,*BPSK, X-QPSK and X-QAM) using compulsory FEC-CC, analyze its spectral-efficiency and BER and there -after FEC –CC along with RS has been taken and analyzed for the same. Secondly, Rayleigh channel and finally most practical Nakagami channel have been implemented following the same sequence of as in the first step. Whole process has been repeated for a different CP.

1.5 Organization of Thesis

This consisting total six chapters, the chapters are organized as below.

Chapter-1: Introduction, it is consisting of basic of OFDM system as it is the technology for WiMAX standard, and then brief of standard 802.16 has been taken with evolution and application.

Chapter-2: Literature Review, vigorous study of research papers of related field in sequence has been discussed in this section.
Chapter- 3: Theory of BER and Spectral Analysis, in this chapter theory along with mathematical expression for BER for all the modulation techniques under consideration has been discussed in brief .Also concept of spectral efficiency has been taken in this chapter.
Chapter- 4: Different Channel Models, in this chapter first of all a basic of fading has been discussed. Modeling of nakagami channel and finally simulation step has been taken.
Chapter5: Results and Discussion, In this chapter all the results simulated in MATLAB-2009a has been discussed along with comparison tables, starting with most ideal channel (AWGN), Rayleigh and Nakagami for our model, for different modulation techniques (*BASK,* BPSK, X-QPSK and X-QAM) using compulsory FEC-CC, analyze its spectral-efficiency and BER and there -after FEC –CC along with RS has been taken and analyzed for the same, and repeated for different CP.
Chapter-6: Conclusion, in this chapter whole work has been concluded, on the basis of results and also future has been discussed.

CHAPTER 2 - LITERATURE REVIEW

History of OFDM System:

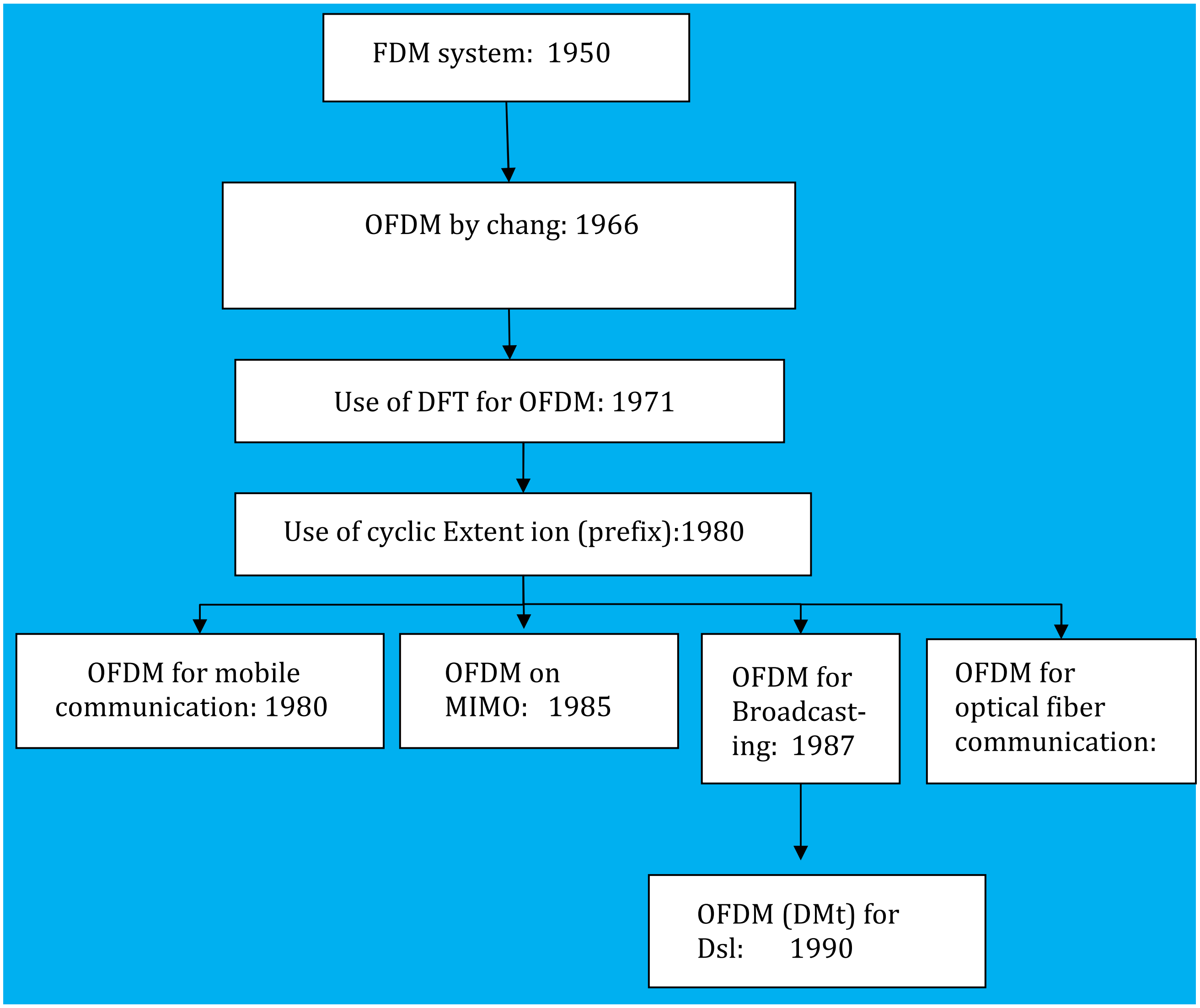


Figure: 2.1 History of OFDM system.

The concept of using parallel-data transmission and frequency-division multiplexing (FDM) was developed in the mid-1960s. Some early development is traced back to the 1950s. A U.S. patent was filed and issued in January 1970. In a classical parallel-data system, the total signal frequency band is divided into *N* non-overlapping frequency sub-channels. Each sub-channel is modulated with a separate symbol, and then the *N* sub-channels are frequency multiplexed. It seems good to avoid spectral overlap of channels to eliminate inter-channel interference. However, this leads to inefficient use of the available spectrum. To cope with the inefficiency, the ideas was proposed in the 1966 by chang [15] of Bell Labs were to use parallel data and FDM with overlapping sub channels, in which each, carrying a signaling rate *b,* is spaced *b* apart in frequency to avoid the use of high-speed equalization and to combat impulsive noise and multipath distortion, as well as to use the available bandwidth fully. In 1971, Weinstein and Ebert applied the discrete Fourier transform (DFT)[16] to parallel data transmission systems as part of the modulation and demodulation processes for OFDM. The cyclic prefix (CP), which is an important aspect of almost all practical OFDM implementations, was proposed in 1980. These are the three key aspects that form the basis of most OFDM systems. The breakthrough papers by Telatar and Foschini on multiple antenna systems fuelled another wave of research in OFDM.[17] OFDM began to be considered for practical wireless applications in the mid–1980s. Cimini of Bell Labs published a paper on OFDM for mobile communications in 1985, while in1987, Lassalle and Alard, based in France considered the use of OFDM for radio broadcasting and noted the importance of combining forward error correction (FEC) with OFDM. Because of this interrelationship, OFDM is often called Coded OFDM (C-OFDM) by broadcast engineers. In the 1990s, OFDM has been exploited for wideband data communications over mobile radio FM channels, high-bit-rate digital subscriber lines (HDSL), asymmetric digital subscriber lines (ADSL), very high-speed digital subscriber lines (VHDSL), digital audio broadcasting (DAB), digital television and HDTV terrestrial broadcasting. The application of OFDM to optical communications has only occurred very recently, but there are an increasing number of papers on the theoretical and practical performance of OFDM in many optical systems including optical wireless single mode optical fiber multimode optical fiber and plastic optical fiber.

The WiMAX Standard:

Broadband Wireless Access (BWA) has emerged as a promising solution for last mile access technology to provide high speed internet access in the residential as well as small and medium sized enterprise sectors. As discussed above section, cable and digital subscriber line (DSL) technologies are providing broadband service. But due to the practical difficulties many urban and suburban locations may not be served by DSL connectivity as it can only reach about three miles from the central office switch [18]. On Broadband wireless Access, because of its wireless nature, it can be faster to deploy, easier to scale and more flexible, thereby giving it the potential to serve customers not served or not satisfied by their wired broadband alternatives. IEEE 802.16 standard for BWA and its associated industry consortium, Worldwide Interoperability for Microwave Access (WiMAX) forum promise to offer high data rate over large areas to a large number of users where broadband is unavailable. This is the first industry wide standard that can be used for fixed wireless access with substantially higher bandwidth than most cellular networks. Wireless broadband systems have been in use for many years, but the development of this standard enables economy of scale that can bring down the cost of equipment, ensure interoperability, and reduce investment risk for operators. The first version of the IEEE 802.16 standard operates in the 10–66GHz frequency band and requires line-of-sight (LOS) towers. Later the standard extended its operation through different PHY specification 2-11 GHz frequency band enabling non line of sight (NLOS) connections, which require techniques that efficiently mitigate the impairment of fading and multipath [19]. Taking the advantage of OFDM technique the PHY is able to provide robust broadband service in hostile wireless channel. The OFD M based physical layer of the IEEE 802.16 standard has been standardized in close cooperation with the European Telecommunications Standards Institute (ETSI) High Performance Metropolitan Area Network (HiperMAN) [18]. Thus, the HiperMAN standard and the OFDM based physical layer of IEEE 802.16 are nearly identical. Both OFDM based physical layers shall comply with each other and a global OFDM system should emerge [19]. The WiMAX forum certified products for BWA comply with the both standards. Communications with direct visibility in the frequency band from 10 to 66 GHz, were dealt by IEEE 802.16 standard, the amendment IEEE 802.16a specifies working in a lower frequency band 2- 11 GHz.

IEEE 802.16d a variation of IEEE 802.16a was basically about optimizing the power consumption of mobile device. IEEE 802.16e is an amendment to IEEE 802.16-2004 and added portability and is oriented to both stationary and mobile deployments. WiMAX standards based projects can work equally well with both the above IEEE standards.

WiMAX vs. Other Wireless Technologies [20-22]

Wireless access to data networks is an area where rapid growth can be expected in coming future and WiMAX can be seen as 4th generation of mobile technology as it has made internet access as well as other multimedia applications possible. In any case, both WLAN and cellular mobile applications are being widely expanded to offer the demanded wireless access. However, they experience several difficulties for reaching a complete mobile broadband access, bounded by factors such as bandwidth, coverage area, and infrastructure costs. On one hand, Wi-Fi provides a high data rate, but only on a short range of distances and with a slow movement of the user. On the other hand, UMTS13 offers larger ranges and vehicular mobility, but instead, it provides lower data rates, and requires high investments for its deployment. WiMAX tries to balance this situation and it fills the gap between Wi-Fi and UMTS, thus providing vehicular mobility (included in IEEE 802.16e), and high service areas and data rates.

Therefore, while WiMAX will complement Wi-Fi and UMTS in some of the possible scenarios where these systems are not sufficiently developed, i.e. they face several problems in the deployment and they do not offer enough capacity to serve all possible users, WiMAX will compete with Wi-Fi and UMTS also in other possible scenarios, where, in general, the costs in the deployment, maintenance, or just the supply of the service would not be profitable

Comparison of WiMAX & Wi-Fi

Wi-Fi also known as WLAN includes 802.11 specifications and is capable to offer data rates upto 54 Mbps in 5.2 GHz and the 80.11b specification, in the 2.4 GHz frequency band, which provides users with data rates of 11 Mbps. This technology has generally a coverage area of 100 meters and fixed channel bandwidth of 20 MHz WiMAX can offer much more than that as already explained and also is doesn't create a conflict with existing Wi-Fi as they are complementary technologies. [3.wxbook.auerbranchmobile] and it flexible BW (1.25MHz to 20MHz.), variable range(2 km to 10km) according to the different development starting from 802.16a to 802.16e. Figure below shows range and data rate variation in PAN, WLAN and WiMAX technology.

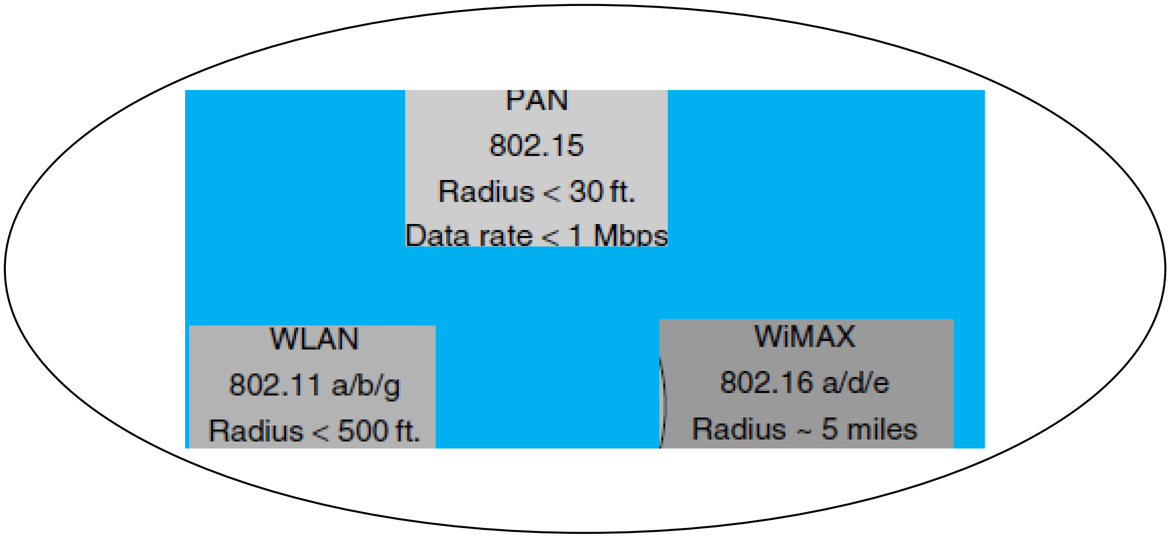


Figure 2.2 Comparison between different technology[23]

Comparison of WiMAX & UMTS

UMTS uses WCDMA as carrier modulation scheme and it provides an integrated solution for mobile voice and data services offering data rates that may decrease with speed of the user This system provides for theoretical bit rates of up to 384 kbps in high mobility situations, which rise as high as 2 Mbps in stationary user environments, employing a 5 MHz channel width WiMAX is becoming a serious threat for 3G cellular networks because of its broadband and distance capabilities, as well as its ability to effectively support voice with full QoS. WiMAX is also able to offer higher data rates than UMTS, but it does not allow the same grade of mobility. However, it is expected to be set up as an alternative to cellular networks, as the investments the operators need to carry out for its deployment are not so high.

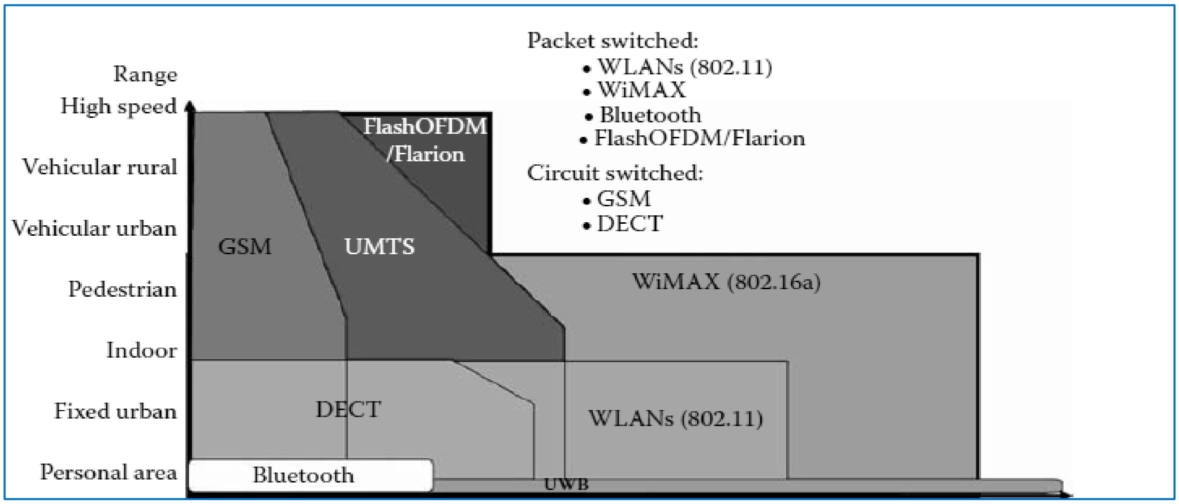


Figure: 2.3 Comparison of WiMAX & UMTS

Spectral-efficiency analysis

Spectral efficiency is defined as the number of bits per second per hertz (b/s/Hz). There are only three techniques are available in literature. On the basis of these techniques there are various models in each. These are modulation techniques, coding techniques and spectrum shaping techniques. Each has been reviewed in brief in this paper. In theory, the number of modulation levels proportional relation with the spectral efficiency. But an increase in modulation result the higher precision is required at the demodulator to detect the phase and frequency. Which require the higher S/N ratio for same BER.

Another way to achieve spectral efficiency is the FEC coding. Different such models are taken and simulated showing significant improvement. Finally, spectrum shaping techniques are included with few different models.

DIGITAL MODULATION TECHNIQUES

There are *basically two type of modulation techniques namely, Linear modulation and continuous modulation [25]. In first type before modulation a high degree of linearity for modulation and RF power amplifier is required. It has more spectral efficiency but a linear power amplifier is essential. The important linear modulations are QPSK, OK-QPSK, π/4 shift QPSK and higher-level PSK. In continuous modulation scheme is quite narrow but at the cost of spectral efficiency. The important* *continuous modulation techniques are MSK GMSK and TFM. From [2],* the average spectral efficiency(S), for perfect CSI, can be given as

$$S = E_{H[n,0],\ldots\ldots,H[n,N-1]}\{\frac{1}{N}\sum_{k=0}^{N-1} a[n,k]\} \quad (2.1)$$

because of the identical statistics of *H*[*n*, *k*] for all *k's,* we can write

$$S = E_{H[n,k]}\{a[n,k] \quad (2.2)$$

However, it is not possible to achieve a perfect CSI, in that case above spectral efficiency give degraded results. The solution for the imperfect CSI problem can be improved using " robust loading algorithm" robust loading algorithm can be brief as suppose the estimated channel gain *H* '[n, k] is the only known information about the current CSI for the *k*th subcarrier of the *n*th

block, and *a*[*n* ,*k*] is computed based on the value of the *n*th block, and *a*[*n, k*] is computed based on the value of H[n, k].Since *Pe*[, *k*] depends on the value of the true channel estimation, *H*[*n, k*], which is assumed unknown in this case, it is not possible to set *Pe*[*n, k*] equal to the target BER. However, we can obtain the average BER with only *H* '[n, k] to be known using CSI. The BER for H' [n, k] is given as

$$\bar{P_e}[n,k] = E_{H[n,k]H'[n,k]}\{P_e[n,k]\} \quad (2.3)$$

Assuming that the conditional probability density function H [n, k] for given H[n, k] is known then$\bar{P_e}[n,k]$ Can be calculated. In "ROBUST LOADING ALGORITHM" $\bar{P_e}[n,k]$ is set equal to P_{target}. In [26] this algorithm has been implemented for both noise channel and delayed channel. For noisy channel assuming MSE of 15dB (achievable) the effect of adaptive channel is negligible. In case of delayed channel the factor $f_d t_d$ is deciding factor for spectral efficiency.

SOURCE CODING TECHNIQUES

Channel coding is techniques of transforming of signals to improve communications performance increasing the robustness against channel impairments noise, interference and fading. Altering the characteristics of the sequences is achieved by coding of the data sequence. The converted sequences which enables decision process, by a transmitter or a receiver, less subject to errors. Channel Coding can be described as a three phase process including Randomization, Forward Error Correction and Interleaving. In [27] an efficient technique for spectrum efficiency has been proposed and simulated, in this paper turbo trellis- coded modulation(TTCM) has been taken as channel coding techniques. The iterative decoding used with TTCM can provide good degree of spectral efficiency and also give protection against partial-band interference. In this model there are two encoders. The systematic outputs of first encoder are ($\log_2$m)/2 MSB of each input symbol, while outputs of the second encoder are the least significant ($\log_2$m)/2 of information bit. The first step in the iterative demodulation and decoding processing is to generate a posteriori probability (APP). For APP first the conditional probability of received signal is derived from CSI. Then received signal is correlated with the signal correspond to every sequence. Averaging of probability is the next step of APP. Now for the first iteration, probability of each sequence is known in advance (prior probability) based on prior knowledge of each symbol. For all future iterations, probability based on information which is derived from the trellis decoders. Next, the turbo decoder implements the iterative MAP decoding. MAP algorithm is based on soft-input ,soft-output algorithm at every decoding stage. For every new APPs previous APPs is used and iteration is process is repeate until a stopping condition is achieved. Then decoder generates hard decisions based on the max APPs. In [3] it is simulated that above model can improve spectral efficiency equal to 1.3bits/sec/Hz, which is 3.2 times greater than conventional system. In [28], a different approach FEC with MMSE has been proposed, the simulation has been done for rate **1/2** convolution- code, with 4 and 16 states using soft output Viterbi algorithm (SOVA). Channel spacing of.752Hz. Observed spectral efficiency is found to be 2.6671bits/sec/Hz.,which is approximately 6.5 times than conventional system.

SPECTRUM SHAPING TECHNIQUES

In [29] spectrum efficiency of OFDM system using root-roll-off filter has been analyzed, it has implemented the above filter at the both end, transmitter and receiver. The benefit of doing so is that compare to raised cosine window, much lower out of band emission level can be achieved, also as impulse response satisfied Nyquist theorem ,the demodulator can regenerate the transmitted symbol stream provided there is no thermal noise .channel distortions and modem impairments. Basically, in the present approach conventional modem has been implemented along with root-roll-off filter. As it well known in the modem regeneration will be perfect provided received signal is sampled at an accurate timing point as inverse Fourier transform converts a transmitted complex symbol into M complex data ,hence BW of filter will be M/Ts. now M times BW gives M times number of sampling points so original data can be regenerated perfectly. In proposed method roll-off factor of filter is 0.2 number of filter taps is 201 at an over-sampling rate of 4 ,number of sub carriers are 64. For above set of data proposed method improves about 60% of the conventional method bandwidth at emission level of -40dB. However in [30] DFE is proposed, applying dual equalization in frequency and time domain, suggested that for given BER it is possible to reduce “guard interval” to get better spectral efficiency. In [31] HC-MCM has been proposed as follows. Let T’ and B’ be the symbol time of a single symbol and average number of message bits in a single sub-carrier in a symbol duration. Then transmission rate is given by R=B’K/T where K is the total number of carriers .On the other hand, when the frequency separation of the subcarriers is assumed to be Δf ,the HC-MCM occupies a bandwidth BW= 2/T+(K-I).Δf [Hz] in total, providing that the main hole bandwidth is taken into account. Therefore, the spectral efficiency for the HC-MCM is given by

$n = \frac{R}{W}$ = (1/TK.B')/[2/T+(k-1). ΔfTs]

= K.B'/[2+(K-1) ΔfTs] (2.4)

The mutual interference due to DFT aliasing can be small. The transmission rate for the HC-MCM will be inversely proportional to the modulation index ΔfTs when ΔfTs is dealt with a normalized symbol duration. **As** simulated in this [31], almost nodegradation appears at

ΔfTs = 0.75 in spite of a 33% increase in BW for 4-QAM conventional OFDM system. Hence a great spectral efficiency improvement. In [32]MCS with CC and SW has been proposed using MCS out of many sequences produced set of sequences having maximum side lobe suppression is selected. There are two steps in it after generation of sequences first selection of sequence with maximum suppression and then selection of sequences with minimum side lobe suppression .Now in SW along with MCS ,the output of MCS side lobe suppression unit is fed Into the SW side lobe suppression unit. SW side lobe Suppression unit performs the multiplication of real Valued weighting factors with the result of MCS side lobe suppression unit. The vector q is fed into the SW side lobe suppression unit

Which outputs

$r = (r_1\ r_2 \dots \dots \dots . r_N)^T$ (2.5)

Hence the vector r is given by,

$r_n = g_n q_n \quad , n = 1,2,3, \dots \dots \dots \dots \dots . N$ (2.6)

n=1,2,...,N, are chosen such that the side lobes of the transmission signal are suppressed weighted vector **r** is modulated on N subcarriers using the I DFT and used for final transmission. However in CC with MCS the output of MCS side lobe suppression unit is fed into the CCs side lobe suppression unit. CCs side lobe suppression unit inserts a few so-called CCs on the left and right hand side of the used OFDM spectrum. CCs don't carry any data but complex weighting factors are associated with them .which are selected in such a way that side lobes of CC can cancel side lobes of original signal .Then normalization is done in order to make the power of the combined signal equal to the power of the original signal .After insertion of CC we need (N+W) sub-carriers for modulation. The side lobes can be sup- pressed by solving the following optimization problem, $min\|B + C_m\|^2$ with condition $\|m\|^2 \leq$ □□□. The combined effect of CC and SW can be summarized as below Numerical results illustrates the effectiveness of the combination of MCS with SW and MCS with CCs for side lobe suppression. BPSK modulation s applied and no channel coding is considered. The number of used subcarriers is set to N =32. The spectra of the OFDM signals with combination of MCS with SW and combination of MCS with CC are illustrated [32]

Here the ratio P for SW is set 0.2. The MCS with SW reduces OFDM side lobes by more than 16 dB. If ratio ρ for SW are increased then even higher side lobe suppression results can be achieved but it degrades the system performance. Next model used included in this paper is[33]partial transmit. In this model very simple principle has been used. By not transmitting the low- energy samples we can achieve significant time (bandwidth) gain. Let a polynomial with degree M is used for cyclic convolution frequency domain signal Xn then in PR-OFDM model it has been shown that as value of M increase a significant narrowing of time domain is observed. In [33] usable bandwidth gain efficient model has been simulated taking following parameters. In this study, we restrict ourselves to an OFDM system using 64 subcarriers and consider PR polynomial orders from M=1, 2, 3 and 4, and d=12, 16, 24 and 32 samples are dropped,(including left and right extremities) respectively. Then assuming no effect on BER or SNR time fraction of time saved d/N is 18.75%, 25, 37.5% and 50% respectively. These saved slots can used to transmit additional OFDM symbols. Corresponding to above saving usable bandwidth gain will be 23%,33%,60%,100% respectively for M=1,2,3,4 and thus partial response model proposed can give to significant bandwidth gains. However loss of SNR and BER has an adverse effect on it. A straightforward calculation shows that PR signaling with M=1, 2, and 3 yields 2.46, 2.66, 3.2 bits/subcarrier, respectively at some SNRs and BER of 5×10^{-3}. For the very same SNRs and BER, in frequency selective channels, using M-ary PSK and plain OFDM yields [34]2.21, 2.25 and 2.3 bits/subcarrier respectively. , we find that PR signaling can yield 11%, 18% and 39% bandwidth (or spectral efficiency) gains respectively. In spectral efficiency for WiMAX has been analyzed using STBC for M-PSK. In the standard specifies multiple techniques to map the data and pilot subcarriers into sub channels are proposed. There are two modes according to the manner that they build sub channel from the subcarriers. The first one is referred to as Distributed Sub-Channelization Mode (DSCM) whereas the second one is referred to as Adjacent Sub-Channelization Mode (ASCM).For a total available bandwidth equal to 20 Mbps; this gain increases the system capacity by 4 Mbps.[35]

CHAPTER 3 - BER & SPECTRAL EFFICIENCY ANALYSIS

3.1 Physical Layer Consideration For WiMAX [11-12]

Three physical layers are used in conjunction with MAC layer to provide reliable end to end links. These three PHY specifications are as follows:-

- Single carrier modulated air interface
- A 256 point FFT OFDM multiplexing scheme
- A 2048 point FFT OFDMA scheme

Here the first, single carrier air interface is basically used with line of sight (LoS) communications where as last two are more suitable for non-LoS applications as equalization of multi carrier system is easier. The fixed WiMAX standards define profiles using the 256-point FFT OFDM PHY layer specification. Furthermore, fixed WiMAX systems provide transmissions with a maximum data rate up to 70 Mbps in a 20 MHz channel bandwidth for service area up to 5 km, and offer the users a broadband connectivity without needing a direct line-of-sight to the base station.

The main features of fixed WiMAX are listed below:-

- Use of an OFDM modulation scheme, which allows the transmission of multiple signals using different subcarriers simultaneously. Because the OFDM waveform is composed of multiple narrowband orthogonal carriers, selective fading is localized to a subset of carriers that are relatively easy to equalize.
- Design of an adaptive modulation and coding mechanism that depends on channel and interference conditions. It adjusts the modulation method almost instantaneously for optimum data transfer, thus making a most efficient use of the bandwidth.
- Support of both time and frequency division duplexing formats, FDD and TDD, allowing the system to be adapted to the regulations in different countries.
- Robust FEC techniques, used to detect and correct errors in order to improve throughput. The FEC scheme is implemented with a Reed- Solomon encoder concatenated with a convolution one, and followed by an interleave. Optional support of block turbo coding (BTC) and convolution turbo coding (CTC) can be implemented.

- Use of flexible channel bandwidths, comprised from 1.25 to 20 MHz, thus providing the necessary flexibility to operate in many different frequency bands with varying channel requirements around the world. This flexibility facilitates transmissions over longer ranges and from different types of subscriber platforms. In addition, it is also crucial for cell planning, especially in the licensed spectrum.
- Optional supports of both transmit and receive diversity to enhance performance in fading environments through spatial diversity, allowing the system to increase capacity. The transmitter implements space time coding (STC) to provide transmit source independence, reducing the fade margin requirement, and combating interference. The receiver, however, uses maximum ratio combining (MRC) techniques to improve the availability of the system.
- Design of a dynamic frequency selection (DFS) mechanism to minimize interferences.
- Optional support of smart antennas, whose beams can steer their focus to a particular direction or directions always pointing at the receiver, and consequently, avoiding interference between adjacent channels, and increasing the spectral density and the SNR. There are two basic types of smart antennas, those with multiple beam (directional antennas), and those known as adaptive antenna systems (AAS). The first ones can use either a fixed number of beams choosing the most suitable for the transmission or an steering beam to the desired antenna. The second type works with multi-element antennas with a varying beam pattern. These smart antennas are becoming a good alternative for BWA deployments.
- Implementation of channel quality measurements which help in the selection and assignment of the adaptive burst profiles.
- Support of both time and frequency division multiplexing formats (TDM and FDM), to allow interoperability between cellular systems working with TDM, and wireless systems that use FDM.

The 2048- point FFT OFDMA PHY specification is used by mobile WiMAX which allow a coverage of 1.6 to 5 km with available transmission rates of 5 mbps in a 5 MHz channel and a user maximum speed below 100 km/h. Apart from the features of fixed WiMAX mobile WiMAX have additional features of handoffs and power saving mechanisms.

The table given below shows a comparison of features of both the standards:-

	802.16	802.16-2004	802.16e
Spectrum	10-66 GHz	<11 GHz	< 6 GHz
Maximum data rate	32-134 Mbps (28 MHz channel)	Up to 70 Mbps (20 MHz channel)	Up to 15 Mbps (5 MHz channel)
Alignment	Los	LoS & NLoS	LoS & NLoS
Coverage Range	Nearly 2-5 km	Nearly 5-10 km	2-5 km
Channel bandwidth	20,25 & 28 MHz	Flexible from 1.25 to 20 MHz	Equal to 802.16-2004
Modulation	2-PAM, 4-QAM, 16-QAM, 64-QAM	OFDM with 256 subcarriers, 2-PAM, 4-QAM, 16-QAM, 64-QAM	OFDMA with 2048 subcarriers 4-QAM, 16-QAM, 64-QAM
Mobility	Fixed	Fixed & Pedestrian	Vehicular (20-100 kmph)

Table 3.1 Comparision Table

3.2 Overall All System Model

The block diagram for Wi-MAX system (standard: 802.16e) is shown

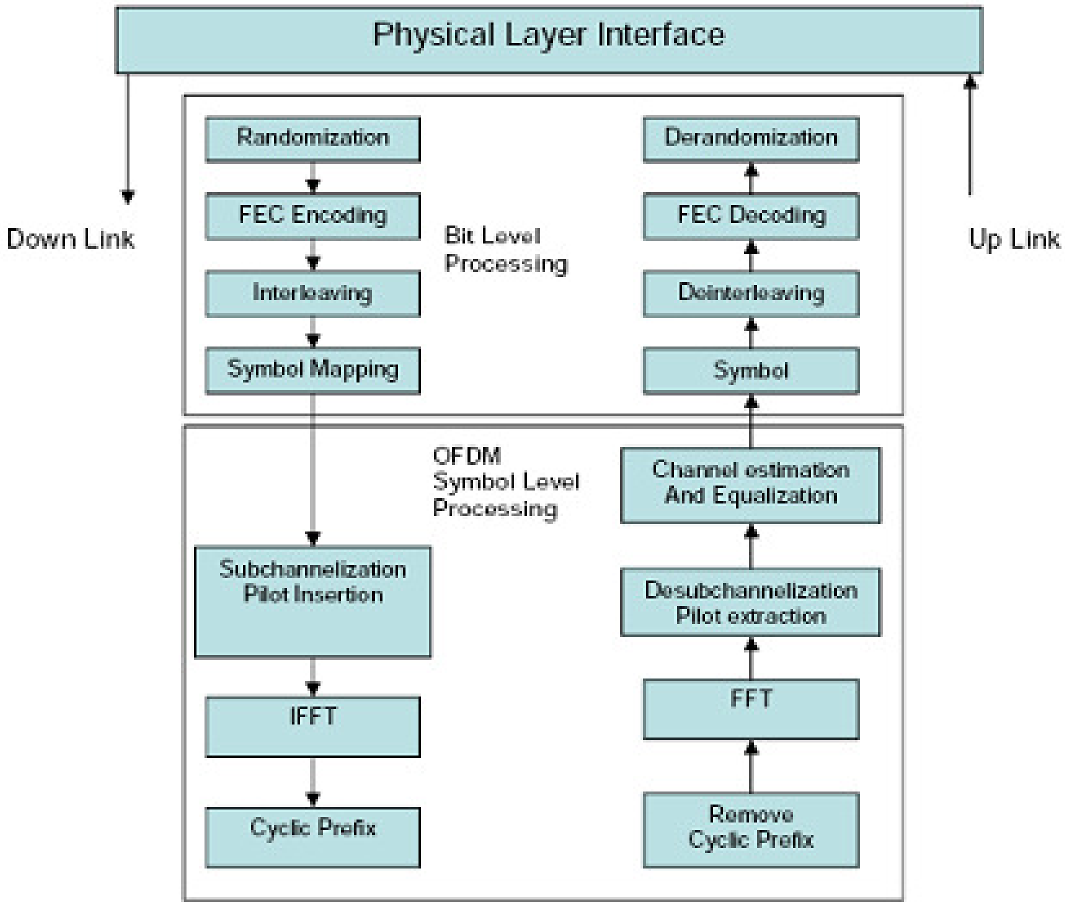


Figure 3.1 Physical layer (802.16e) [11]

Now we will take each block one by one in detail and in present paper simulation has been done for each block separately.

3.2.1 Randomization

Randomization is the first process carried out in layer after the data packet is received from the higher layers each burst in Downlink and Uplink is randomized. It is basically scrambling of data to generate random sequence to improve coding performance .Linear Feedback Shift Resister main component used for this purpose.[7]

3.2.2 Forward Error Correction (FEC)

Coding has the usefulness that it allows us to increase the rate at which information may be transmitted over a channel while maintaining a fixed error rate. Alternatively, coding allows us to reduce the information bit error rate while maintaining a fixed transmission rate. More generally, coding allows us, in principle (up to the Shannon limit) to design a communication system in which both information bit rate and error rate are independently and arbitratarily specified but subject to a constraint on bandwidth. The price we pay, for seeking to reach closer to Shannon limit, is increased hardware complexity both at the transmitter where encoding is done and at the receiver where decoding is affected. In principle, with ingenious enough coding and unlimited complexity we would be able to reach the Shannon limit. That is, we would be able to transmit at channel capacity and with an error rate which may be made as small as desired. One measure of the efficiency of a code is precisely the extent to which it allows us to approach the Shannon limit.[20] In this thesis CC and RS is used for the channel coding purpose.

3.2.2.1 Convolution Coding (CC)

In block coding, the encoder accepts a k-bit message block and generates an n bit code word. Thus, code words are produced on block by block basis. Clearly, provision must be made in the encoder to buffer an entire message block before generating the associated code word. There are applications, however, where the message bits come in serially rather than in large blocks, in which case the use of buffer may be undesirable. In such case the use of convolutional coding may be preferred method [2].

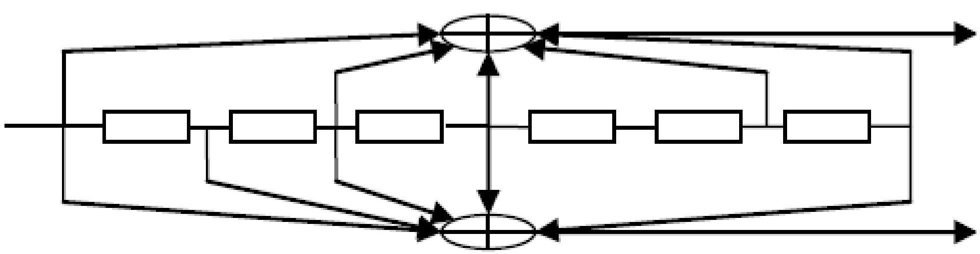


Figure 3.2: Convolutional encoder [6]

A convolution code introduces redundant bits into the data stream through the use of linear shift register. The information bits are input into shift register and the output encoded bits are obtained by modulo-2 addition of the input information bits and the contents of the shift register 802.11a physical layer uses Convolution code as the mandatory FEC. The convolutional encoder shall use the industry-standard generator polynomials, g0 = 1338 and g1 = 1718, of rate R = 1/2, [2] as shown in Figure 3.2. The bit denoted as “A” shall be output from the encoder before the bit denoted as “B.” Higher rates like 2/3 and 3/4, are derived from it by employing “puncturing.” Puncturing is a procedure for omitting some of the encoded bits in the transmitter (thus reducing the number of transmitted bits and increasing the coding rate) and inserting a dummy “zero” metric into the convolutional decoder on the receive side in place of the omitted bits. For Decoding the Viterbi algorithm is used.

To describe a convolution code, one needs to characterize the encoding function (m), so that given an input sequence m, one can readily compute the output sequence U. Several methods are used for representing a convolutional encoder, the most popular being the connection pictorial, connection vectors and polynomials, the state diagram, the tree diagram. The encoder can be represented in several different but equivalent ways representation. They are [4-7]

- Generator/connection Representation
- Tree Diagram Representation
- State Diagram Representation
- Trellis Diagram Representation

3.2.2.2 Connection Representation

We shall use the convolutional encoder, shown in fig. 3.3, as a model for discussing convolution encoders. The figure illustrates a (2,1) convolution encoder with constraints length$k$=3. There are n=2 modulo-2 adders; thus the code rate k/n is ½. At each input bit time, a bit is shifted into the left most stage and bits in the register are shifted one position to the right. Next, output switch samples the output of each modulo-2 adder (i.e., first the upper adder, then the lower adder), thus forming the code symbol pair making up the branch word associated with the bit just inputted.

The sampling is repeated for each inputted bit. The choice of connections between the adders and the stages of the register gives rise to the characteristics of the code. Any change of the choice of connections result in a different code. The connections are of course, not chosen or changed arbitrarily. The problem of choosing connections to yield good distance properties is complicated and has not been solved in general.

Unlike a block code that has a fixed word length n, a convolution code has no particular size. However, convolution codes are often forced into a block structure by periodic truncation. This requires a number of zero bits to be appended to the end of input data sequence, for the purpose of clearing of flushing the encoding shift register of the data bits. Since the added zeros carry no information, the effective code rate falls belowk/n. To keep the code rate close to k/n, the truncation period is generally made as long as practical.

3.2.2.3 Polynomial Representation

Sometimes, the encoder connections are characterized by generator polynomials. We can represent a convolution encoder with a set of n generator polynomials, one each of the n modulo-2 adders. Each polynomial is of degree $k-1$ or less and describes the connection of the encoding shift register to that modulo-2 adder, much the same way that a connection vector does. The coefficient of each term in the $k-1$ degree polynomial is either 0 or 1 depending on whether a connection exists or does not exist between shift register and modulo-2 adder in question.

$g1 = 1\ 1\ 1$

$g2 = 1\ 0\ 1$

$g1(X) = 1+X+X^2$

$g2\ (X) = 1+X^2$

The output sequence is found as follows:

$(X) = m(X)\ g1(X)$ Interlaced with $m\ X\ g2\ X$ (3.1)

3.2.2.4 State Representation and the State Diagram

A convolution encoder belongs to a class of devices known as finite state machine that means machine that have a memory of past signals. The adjective finite refers to the fact that there are only a finite number of unique states that the machine can encounter. What is meant by the state of a finite state machine? In the most general sense, the state consists of the smallest amount of information that, together with a current input to the machine, can predict the output of the machine. The state provides some knowledge of the past signaling events and the restricted set of possible output in the future. A future state is restricted by past state. For a state $1/n$ convolution encoder, the state is represented by the contents of the rightmost $k-1$ stages. Knowledge of the state together with knowledge of the next input is necessary and sufficient to determine the next output. Let the state of the encoder at time ti be defined as $Xi = mi-1$, $mi-2$,..., mi. The $it\square$ codeword branch Ui is completely determined by state Xi and the present input bit mi; thus the state Xi represents the past history of the encoder in determining the encoder output. The encoder state is said to be Markov, in the sense that the probabilities p(Xi+1|Xi, Xi−1…….X0) of being in state Xi+1,given all previous states, depends only on the most recent state Xi;that is the ,the probability is equal to p(Xi+1|Xi).

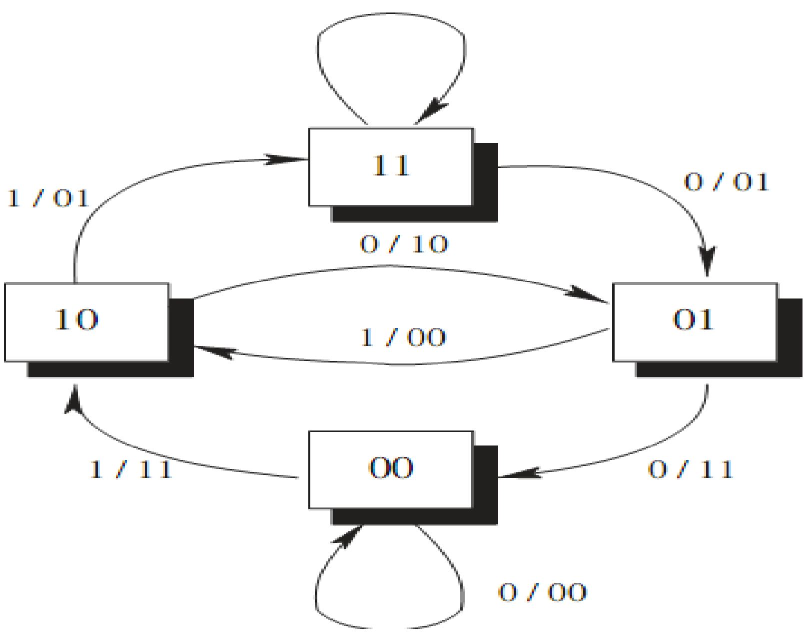


Fig 3.3 State diagram for encoder

One way to represent simple encoder is with a state diagram; such a representation for the encoder in fig. 3.3 is shown in fig. 3.5. The states, shown in the boxes of the diagram, represent the possible contents of the right most $k-1$ stages of the register and the path between the states represent the output branch words resulting from such state transition. The state of the register are designated a= 00, b=10, c=01, d=11; the diagram shown in fig.3.3 illustrates all the state transitions that are possible for the encoder in fig. 3.3. There are only two transitions emanating from each state, corresponding to the two possible input bits. Next to each path between states is written the output branch word associated with the state transition. In drawing the path, we use the convention that a solid line denotes a path associated with the input bit, zero and a dashed line denotes a path associated with an input bit one. Notice that it is not possible in single transition to move from a given state to any arbitrary state. As a consequence of shifting in one bit at a time, there are only two possible state transitions that the registers can make at each bit time. For example, the present encoder state is 00; the only possibilities for the state at the next shift are 00.

3.2.2.5 The Tree Diagram

Although the state diagram completely characterize the encoder, one cannot easily use it for tracking the the encoder transitions as a function of time since the diagram cannot represent time history. The tree diagram adds the diamension of time to the state of diagram. The tree diagram for the convolutional encoder shown in fig 3.3 is illustrated in fig. 3.5

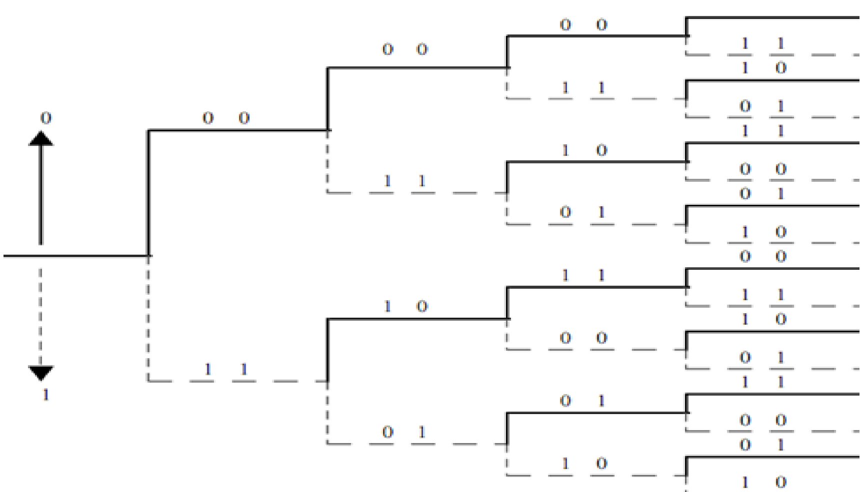


Fig 3.4 Tree diagram

At each successive input bit time the encoding procedure can be described by traversing the diagram from left to right, each tree branch describing an output branch word. The branching rule for finding a codeword sequence is as follows .If the input bit is zero, its associated branch word is found by moving to the next right most branch in the upward direction .If input bit is one, its branch word is found by moving to the next rightmost branch in the downward direction .assuming that the initial contents of the encoder is all zeros. The diagram shows that if input bit is zero, the output branch is 00 and if the first input bit is one, the output branch is 11. Similarly, if the first input bit is one and the second input bit is a zero , the second output branch word is 10.or the first input bit is a one and second input bit is one, second output branch is 01.Following procedure we see that the input sequence 11011 traces the heavy line drawn on the tree diagram in fig. 3.5.This path corresponds to the output code word sequence 1101010001.

3.2.2.5 The Trellis Diagram

Observation of the fig. 3.5 tree diagram shows that for this example, the structure repeats itself at time t4, after third branching (in general the tree structure repeats after k branching ,where k is the constraint length). We label each node in the tree of fig. 3.5 correspond to the four possible states in the shift register, as follows: a=00, b=10, c=01, and d=11.The first branching of the tree structure, at time t1, produces a pair of nodes labelled a and b. At each successive branching the number of nodes doubles. The second branching, at time t2, results in four nodes labelled a, b, c, and d. After the third branching, there are total of eight nodes: two are labelled a, two are labelled b, two are labelled c, and two are labelled d.

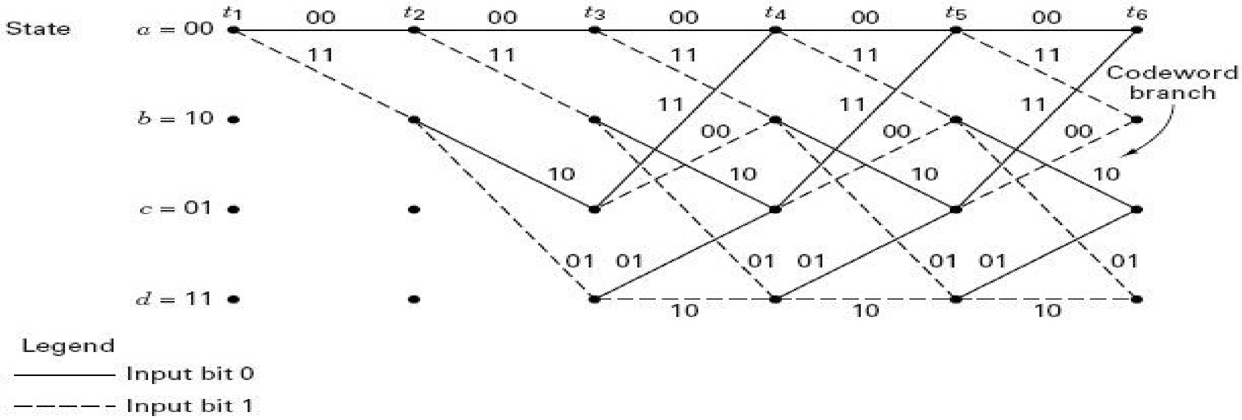


Fig
3.5 Trellis Diagram

We can see that all branches emanating from two nodes of the same state generate identical branch word sequences. From this point on, the upper and lower halves of the tree are identical. The reason for this should be obvious from examination of encoder in fig 3.3.As the fourth input bit enters the encoder on left, the first input bit is ejected on the right and no 23 longer influences the output branch words. Consequently, the input sequences 1 0 0 x y….and 0 0 0 x y…, where the left most bit is the earliest bit, generates the same branch words after the (k=3) rd branching. This means that any two nodes having the same state label at the same time can be merged, since all succeeding path will be indistinguishable. If we do this to tree structure of figure 3.3, we obtain another diagram, called the trellis diagram. The trellis diagram, by exploiting the repetitive structure, provides a more manageable encoder description than does the tree diagram. The trellis diagram for the convolutional encoder of fig.3.3 is shown in fig 3.6.

In drawing the trellis diagram, we use the same convention that we introduced with the state diagram – a solid line denotes the output generated by an input bit zero, and a dashed line denotes the output generated by an input bit one. The nodes of the trellis characterize the encoder states; the first row nodes correspond to the state a= 00, the second and subsequent rows correspond to the states b= 10, c= 01 and d= 11.At each unit of time, the trellis requires $2k-1$ nodes to represent the $2k-1$ possible encoder states. The trellis in our example assumes a fixed periodic structure after trellis depth 3 is reached (at time$t4$). I n general case, the fixed structure prevails after depth k is reached. At this point and thereafter, each of the state can be entered from either of two preceding states. Also each of the states can transition to one of the two states. Of the two outgoing branches, one corresponds to an input bit zero and the other corresponds to an input bit one. On fig 3.6, the output branch words corresponding to the state transitions appear as labels on the trellis branches.

3.2.2.7 The Viterbi Convolutional Decoding Algorithm

Viterbi decoder can be represented in three parts

1. Find out the Hamming distance of each path.
2. Add the distance of each path.
3. Compare and select the shortest path.

The Viterbi decoding algorithm was discovered and analyzed by Viterbi in 1967. The Viterbi algorithm essentially performs maximum likelihood decoding; however, it reduces the computational load by taking advantage of the special structure in the code trellis. The advantage of Viterbi decoding, compared with brute force decoding, is that the complexity of a Viterbi decoder is not a function of the number of symbols in the codeword sequence. The algorithm involves calculating a measure of similarity or distance, between the received signal, at time ti, and all the trellis paths entering each state at time ti. The Viterbi algorithm removes from consideration those trellis paths that could not possibly be candidates for the maximum likelihood choice. When two paths enter the same state, the one having the best metric is chosen; this path is called the surviving path. This selection of surviving path is performed for all the states. The decoder continues in this way to advance deeper into the trellis, making decision by eliminating the least likely path. The Viterbi algorithm is, in fact, maximum likelihood. Note that the goal of selecting the optimum path can be expressed, equivalently, as choosing the codeword with the maximum likelihood metric or as choosing the codeword with the minimum distance metric.

3.2.2.8 Reed Solomon Codes (RS codes)

The Reed-Solomon codes are nothing but the subclass of non binary BCH codes; they are usually abbreviated as RS codes. The encoder for RS code operates on multiple bits rather than individual bits unlike binary encoder which operates on single bit. Specifically, an RS $(n,)$ code is used to encode m-bit symbols by adding $n-k$ redundant symbols. When m an integer power of two, the m-bit symbols are called bytes. A popular value of m is 8; because, 8-bit RS codes are extremely powerful.[4]

Parameters of a t-error-correcting RS code

Block length: $n=2m-1$ $symbols$

Message size: k $symbols$

Parity-check size: $n-k=2t$ $symbols$

Minimum distance: $dmin=2t+1$ $symbols$

Reed-Solomon codes can be used as both error-correcting and erasure codes. In the error-correcting setting, we wish to transmit a sequence of numbers over a noisy communication

channel. The channel noise might cause the data sent to arrive corrupted. In the erasure setting, the channel might fail to send our message. For both cases, we handle the problem of noise by sending additional information beyond the original message. The data sent is an encoding of the original message. If the noise is small enough, the additional information will allow the original message to be recovered, through a decoding process.

3.2.2.9 Arithmetic in GF (2^r)

In practice, we want our Reed-Solomon codes to be very efficient. In this regard, working in $GF(p)$ for some prime is inconvenient, for several reasons. Let us suppose it is most convenient if we work in blocks of 8 bits. If we work in $GF(251)$, we are not using all the possibilities for our eight bits. Besides being wasteful, this is problematic if our data (which may come from text, compressed data, etc.) contains a block of eight bits which corresponds to the number 252!

It is therefore more natural to work in a field with $2r$ elements, or $GF(2^r)$. Arithmetic is this field is done by finding an irreducible (prime) polynomial $\pi(x)$ of degree r, and doing all arithmetic in $Z2[\pi(x)]$. That is, all coefficients are modulo 2, arithmetic is done modulo $\pi(x)$, and $\pi(x)$ should not be able to be factored over $GF(2)$.

For example, for $GF(28)$, an irreducible polynomial is $\pi(x) = x^8 + x^6 + x^5 + x + 1$. A byte can naturally be thought of as a polynomial in the field. For example, by letting the least significant bit represent $x0$, and the ith least significant bit represent xi, we have that the byte 10010010 represents the polynomial $x^7 + x^4 + x$. Adding in $GF(2^r)$ is easy: since all coefficients are modulo-2, we can just XOR two bytes together. For example

10010010+10101010 = 00111000

$$(x7 + x4 + x) + (x^7 + x^5 + x^3 + x) = x^5 + x^4 + x^3 \tag{3.2}$$

Moreover, subtracting is just the same as adding. Multiplication is slightly harder, since we work modulo $\pi(x)$. As an example

$$(x^4 + x)\,(x^4 + x^2) = x^8 + x^6 + x^5 + x + 1 \qquad (3.4)$$

However, we must reduce this so that we can fit it into a byte. As we work modulo π , we have that $\pi(x) = 0$, or $x^8 = x^6 + x^5 + x + 1$. Hence

$$(x^4 + x)\,.(x^4 + x^2) = x8 + x^4 + x^5 + x^3 = (x^6 + x^5 + x + 1) + x^6 + x^5 + x^3 = x^3 + x + 1 \qquad (3.5)$$

And hence 00010010×00010100=00001011.

Rather than compute these products on the fly, all possible 256× 256 pairs can precomputed once in the beginning, and then all multiplications are done by just doing a lookup in the multiplication lookup table. Hence by using memory and pre-processing, one can work in G (28) and still obtain great speed.

Reed-Solomon codes work exactly the same over $GF\,(2^r)$ as they do over $GF(p)$, since in both cases the main requirement, namely that a polynomial of degree $d - 1$ be uniquely defined by d points, is satisfied.

3.2.3 Interleaver

Interleaving can be employed in digital data transmission technologies to mitigate the effect of burst errors. When too many errors exist in one code word, due to a burst error, the decoding of a code word cannot be done correctly. To reduce the effect of burst error, the bits in one code word are interleaved before being transmitted. When interleaving occurs the place of bits will change, which means that a burst error can not disturb a huge part of one code word.[8] All encoded data bits shall be interleaved by a block interleaver with a block size corresponding to the number of bits in a single OFDM symbol N_{CBPS}. The interleaver is defined by a two-step permutation. The first permutation ensures that adjacent coded bits are mapped onto nonadjacent subcarriers. The second ensures that adjacent coded bits are mapped alternately onto less and more significant bits of the constellation and, thereby, long runs of low reliability (LSB) bits are avoided. We shall denote by k the index of the coded bit before the first permutation; i shall be the index after the first and before the second permutation, and j shall be the index after the second permutation, just prior to modulation mapping.

The first permutation is defined by the rule

$i = (N_{CBPS}/16)(k mod\ 16 + floor\ k16)$ where, k=0,1,…,N_{CBPS}−1

The function floor (.) denotes the largest integer not exceeding the parameter.

The second permutation is defined by the rule

$j = s * floor\ is + i + N_{CBPS} - floor\ 16 * iN_{CBPS}\ mod$, where i=0,1,…,N_{CBPS}−1

The value of s is determined by the number of coded bits per subcarrier, N_{BPSC}

s=max $(N_{CBPS}/2,1)$

The de-interleaver, which performs the inverse relation, is also defined by two permutations.

Here we shall denote by j the index of the original received bit before the first permutation; i shall be the index after the first and before the second permutation, and k shall be the index after the second permutation, just prior to delivering the coded bits to the convolutional (Viterbi) decoder.

The first permutation is defined by the rule

$i = s * floor(j/s) + (j + floor\ (16 * jN_{CBPS}))mods$ j=0,1,…,$NCBPS$−1.

The second permutation is defined by the rule $k = 16 * i - (N_{CBPS} - 1)\ floor\ 16 * i/N_{CBPS})$ i=0,1,… $NCBPS$−1

3.2.4 Subcarrier modulation and mapping [5-7]

The OFDM subcarriers shall be modulated by using BPSK, QPSK, 16-QAM, or 64-QAM modulation, depending on the RATE requested. The encoded and interleaved binary serial 28

input data shall be divided into groups of $NCBPS$ (1, 2, 4, or 6) bits and converted into complex numbers representing BPSK, QPSK, 16-QAM, or 64-QAM constellation points. The output values, d, are formed by multiplying the resulting $(I+jQ)$ value by a normalization factor$KMOD$, as described as

$d = I + jQ * KMOD$

The normalization factor, $Kmod$, depends on the base modulation mode, as prescribed in Table 3.1 The purpose of the normalization factor is to achieve the same average power for all mappings.

3.2.4.1 Binary Phase Shift Keying (BPSK)

The phase of the carrier is varied to represent binary ones and zeros which is used to transmit data via changing and modulating of carrier wave is called Phase Shift keying and if the phase shift uses two phases differing by 180 degree to represent binary digits, the modulation is called BPSK. The principle equation of BPSK as following:

S(t) = Acos (2πfct) :Binary 1, Acos (2πfct+π):Binary 0, Acos (2πfct): Binary 1,

Acos (2πfct):Binary 0,

Where

A = constant

fc= the carrier frequency

t = the bit duration 29

3.2.4.2 Quadrature Phase Shift Keying (QPSK)

QPSK is also known as quaternary PSK, 4-PSK or 4-QAM. Each symbol consists of two bits and signal transmits among the phases that are separated by 90 degrees but used only one bit per channel. The constellation contains four points but the decision make in two bits. The principle equation of QPSK as following: $(t) = Acos\left(2\pi fc + \frac{\pi}{4}\right): 11,\ Acos\left(2\pi fct + \frac{3\pi}{4}\right): 01, Acos\left(2\pi fct + \frac{3\pi}{4}\right): 00$,

$Acos\left(2\pi fct + \frac{3\pi}{4}\right): 10$. (3.6)

3.2.4.3 16-Quadrature Amplitude Modulation

This is also called 16-state quadrature amplitude modulation which means four different phases and 4 different amplitudes are used for 16 symbols because 24 = 16. In this mechanism every symbol represents 4 bits.

64 QAM as same as 16-QAM except it is 64 possible signal combinations with each symbol represent six bits(26 =64). 64 QAM is a complex modulation technique but gives high efficiency.

3.2.5 Pilot subcarriers

In each OFDM symbol, four of the subcarriers are dedicated to pilot signals in order to make the coherent detection robust against frequency offsets and phase noise. These pilot signals shall be put in subcarriers –21, –7, 7 and 21. The pilots shall be BPSK modulated by a pseudo binary sequence to prevent the generation of spectral lines. [4]

3.2.6 IFFT and FFT

An inverse Fourier transform converts the frequency domain data stream into the corresponding time domain [9]. Then a parallel to serial convertor is used to transmit time domain samples of one symbol. The Fast Fourier Transformation (FFT) is used to convert data in time domain to the frequency domain at the receiver. An inverse Fourier transform converts the frequency domain data input to time domain representing OFDM Subcarrier. IFFT is useful for OFDM because it generates samples of a waveform with frequency component satisfying orthogonality condition. It also removes the need of oscillator.

3.2.7 Cyclic Prefix Addition

1/2 or 1/4 or1/8 or 1/16 or 1/32 times of data symbol is added at beginning of the OFDM. [4]

3.3.1 BER Analysis for different channels and modulation schemes

Today's demand for communication of "anywhere any time" can be fulfilled by higher capacity and reliability, WiMAX (world-wide inter-operability for micro wave access) has taken place for above purpose. The enhanced feature of this standard is that it can provide wireless connectivity with interoperability with fixed, portable, nomadic users as well.[36].There is gradual advancement in this standard, starting with 802.16 , designed for LOS in frequency band 10 to 66GHz, then concept of NLOS was introduced for frequency range 2 to 11 GHz was named as fixed standard "802.16a".With little variation 802.16a was evolved in order to optimize the power consumption of the mobile device. Basically 802.16e standard is an advancement of 802.16-2004 standard by adding portability [37] (As 802.16-2004 was targeted only for stationary transmission).

For 802.16e MAC & PHY layers has been defined, but for present work only PHY has been taken into consideration.

PHY layer for mobile WiMAX (IEEE-802.16e) has scalable FFT size128-2048 point FFT with OFDMA, Range is from 1.6 to 5 Km. at 5Mbps in 5MHz. channel BW, it supports 100Km/hr speed.

In the preceding section physical layer will be discussed in the full detail. [38]

3.3.2. Mathematical Analysis for B.E.R

For quasi-static flat fading channel, after sampling and passing through FFT processor, k^{th} sub-carrier can be given by

$$r_k = \alpha X_k S_l + \alpha \sum_{l=1, l \neq k}^{N} S_{l-k+1} X_l + \eta_k \tag{3.7}$$

k = 1,2,3,………..N.

Where X_k symbol used for k^{th} sub-carrier and □ is a complex Gaussian variable zero mean and $\sigma^2{}_R$ as variance. η_k As AWGN channel with zero mean and σ^2 as variance and N is the number of sub-carriers .The coefficients of the ICI is given as

$$S_k = \sin(\pi[k-1+\varepsilon]) / N \sin(\pi[k-1+\varepsilon]) \exp\{j\pi(1-1/N)(k-1)\} \tag{3.8}$$

3.3.2.1 BPSK Modulation

In BPSK modulation 1 or 0 is directly represented by the phase 1 or 0 respectively. Hence wave form can be given as

$$X_k(t) = A\cos\{2\pi f_c t + \pi . d_k\} \tag{3.9}$$

3.3.2.2 AWGN Channel

Characteristic function of above equation using proper- trigonometric can be given as

$$\phi_{R(r1)}(\omega) = 1/2^{N-2} \exp\{j\omega R(S_1) - \sigma^2\omega^2/2\} \sum_{K=1}^{2^{N-1}} \cos[\omega R(S^T e_k)] \quad (3.10)$$

Using Euler's identity

$$\phi_{R(r1)}(\omega) = 1/2^{N-2} \sum_{k=1}^{2^{N-2}} (\exp\{j\omega\theta_k - \sigma^2\omega^2/2\} + \exp\{j\omega\beta_k - \sigma^2\omega^2/2\}) \quad (3.11)$$

Where $\theta_k = R(S + S^T)$ and $\beta_k = R(S - S^T)$

Above represents the kind of Gaussian density function as we are using binary data, so there will be error if and only if $R(r_1) \le 0$ so the probability of error

$$P_b(\varepsilon) = 1/2^{N-1} \sum_{k=1}^{2^{N-1}} \left\{ Q(\sqrt{2\gamma\theta_k}) + Q(\sqrt{2\gamma\beta_k} \right\} \quad (3.12)$$

Where

$\gamma = E_b / N_0$; Q(x) is the Gaussian function Q(x). Above is the general form for BER. Taking ε is equal to zero and θ_k = β_k = 1 (6) reduces for channel without offset .

3.3.2.3 Rayleigh Fading Channel

Raleigh channel can be obtained simply introducing Complex fading coefficient $|\alpha|$ in the above equation, so

$$P_b(\varepsilon) = 1/2^{N-1} \sum_{k=1}^{2^{N-1}} \left\{ Q(\sqrt{2\gamma\theta_k}\,|\alpha|) + Q(\sqrt{2\gamma\beta_k}\,|\alpha|) \right\} \tag{3.13}$$

Now bit error probability

$$P_b = \int_0^{\infty} P_b(\varepsilon\,|\alpha|)\, f\,|\alpha|\, d\,|\alpha| \tag{3.14}$$

where,

$$f(|\alpha|) = |\alpha| / \sigma^2_R \exp(|\alpha| / \sigma^2_R)$$

with proper mathematical manipulation, finally we get bit error probability as

$$P_b\{\varepsilon\} = 1/2 - 1/2^N \sum_{k=1}^{2^{N-2}} \left(\sqrt{2\sigma_R^2\theta_k^2\gamma/1+2\sigma_R^2\theta_k^2\gamma} + \sqrt{2\sigma_R^2\beta_k^2\gamma/1+2\sigma_R^2\beta_k^2\gamma} \right) \tag{3.15}$$

3.3.2.4 QPSK Modulation

QPSK signal can be generated by two BPSK signals. To distinguish between two signals orthogonality of sine and cosine has been used. We can represent QPSK by

$$x(t) = 1/\sqrt{2}d_I(t)\cos(2\pi f_c t) + 1/\sqrt{2}d_Q(t)\sin(2\pi f_c t) \tag{3.16}$$

3.3.2.5 AWGN Channel

Let 1st sub-carrier used to transmit symbol is X_1 = 1+j out of X_k, so sampled signal after sampling and taking FFT we have

$$u = X_k S_l + \sum_{l=1, l \neq k}^{N} S_{l-k+1} X_l \tag{3.17}$$

Now after addition of noise the probability of correct decision is

$$P(u+n \in D_1 / X_1 = 1+j, u) = Q(-R(u)/\sigma) \times Q(-I(u)/\sigma) \tag{3.18}$$

Where D_1 is the quadrant of complex plane with no probability of error and *I(u)* is the imaginary part of u.

From [1.final.ber], Corresponding SER is given as

$$P_S(\varepsilon) = 1 - 1/2^{2N-2} \sum_{k=1}^{2^{N-2}} \sum_{n=1}^{2^{N-2}} \left(\sum_{m=1}^{4} Q(-\sqrt{2\gamma\psi_{kn}[1,m]}) Q(-\sqrt{2\gamma\psi_{kn}[2,m]}) \right) \tag{3.19}$$

With ε equal to zero equation above expression can be reduce for channel without CFO.

3.3.2.6 Rayleigh Fading Channel

As explained in the previous section symbol error rate for Rayleigh channel can be deduced with AWGN channel simply introducing Fading coefficient. [39]

So SER for Raleigh channel is given by

$$P_s(\varepsilon) = 1 - 1/2^{2N-2} \sum_{k=1}^{2^{N-2}} \sum_{n=1}^{2^{N-2}} \left(\sum_{m=1}^{4} Q(-\sqrt{2\gamma\psi_{kn}[1,m]|\alpha|}) Q(-\sqrt{2\gamma\psi_{kn}[2,m]|\alpha|}) \right) \tag{3.20}$$

3.3.2.7 QAM Modulation

In QAM modulation two bits can be transmitted during symbol time. The waveform is

$$x(t) = d_I(t)\cos(2\pi f_c t) + d_Q(t)\sin(2\pi f_c t) \tag{3.21}$$

3.3.2.8 Rayleigh Channel

Now let v be the of mobile terminal, c be the speed of light, f_c be the carrier frequency, then Doppler frequency fm, is given by $f_m = f_c(v/c)$

The density function for M-ary QAM is given by

$$p_{\gamma s}(x) = 1/\overline{\gamma_s} \exp -(x/\overline{\gamma_s}), x \geq 0 \tag{3.22}$$

Where, γ_s and $\overline{\gamma_s}$ are symbol energy to noise ratio for received signal and average received symbol energy to noise ratio respectively

And average symbol error probability,

$$P_s = \int_0^{\infty} P_M(x) p_{\gamma s}(x) dx \tag{3.23}$$

Where average Es to noise ratio is

Now,

$$\overline{\gamma_s} = 1/\{1 - 1/N^2[N + 2\sum_{i=1}^{N-1} 9N - i) j_0(2\pi f_m T_m i)] + NT_s / E_s / N_0\} \tag{3.24}$$

average Eb to noise ratio $\overline{\gamma_b}$ can be calculated by following formula

$$\overline{\gamma_b} = \gamma_s / \log_2 M \tag{3.25}$$

Where M is the number of symbol for M-ary QAM from previous two equations.

$$\overline{\gamma_b} = 1/\{1 - 1/N^2[N + 2\sum_{i=1}^{N-1} 9N - i) j_0(2\pi f_m T_m i)] + (NT_s \log_2 MA)/(E_b / N_0)\} \tag{3.26}$$

3.3.2.9 AWGN Channel

Probability of bit error rate for AWGN can be simply deduce from above expression as [40]

$$P_b = Q(\sqrt{2\gamma_b}) \tag{3.27}$$

Finally BER is given as

$$P_b = \int_0^{\infty} p_b(x)\Upsilon_s(x)dx = 1/2[1 - sqrt(\frac{\Upsilon b}{(1+\Upsilon b)})] \tag{3.28}$$

3.4 Spectral-efficiency analysis [25-27]

Spectral efficiency is defined as the number of bits per second per hertz (b/s/Hz). There are only three techniques are available in literature. On the basis of these techniques there are various models in each. These are modulation techniques, coding techniques and spectrum shaping techniques. Each has been reviewed in brief in this paper. In theory, the number of modulation

levels proportional relation with the spectral efficiency. But an increase in modulation result the higher precision is required at the demodulator to detect the phase and frequency. Which require the higher S/N ratio for same BER.

Another way to achieve spectral efficiency is the FEC coding. In this paper different such models are taken and simulated showing significant improvement. Finally, spectrum shaping techniques are included with few different models.

CHAPTER 4 - FADING CHANNEL –BASICS

4.1 Fading-definition [40-42]

The most troublesome and frustrating problem in receiving radio signals is variations in signal strength, most commonly known as 'FADING'. There are several conditions that can produce fading. When a radio wave is refracted by the ionosphere or reflected from the Earth's

Surface, random changes in the polarization of the wave may occur. Vertically and horizontally mounted receiving antennas are designed to receive vertically and horizontally polarized waves, respectively. Therefore, changes in polarization cause changes in the received signal level because of the inability of the antenna to receive polarization changes. Fading also results from absorption of the RF energy in the ionosphere. Absorption fading occurs for a longer period than other types of fading, since absorption takes place slowly.

Usually, however, fading on ionospheric circuits is mainly a result of multipath propagation.

4.1.1 Multi path fading

Multipath fading is a feature that needs to be taken into account when designing or developing a radio communications system. In any terrestrial radio communications system, the signal will reach the receiver not only via the direct path, but also as a result of reflections from objects such as buildings, hills, ground, water, etc that are adjacent to the main path as shown below.

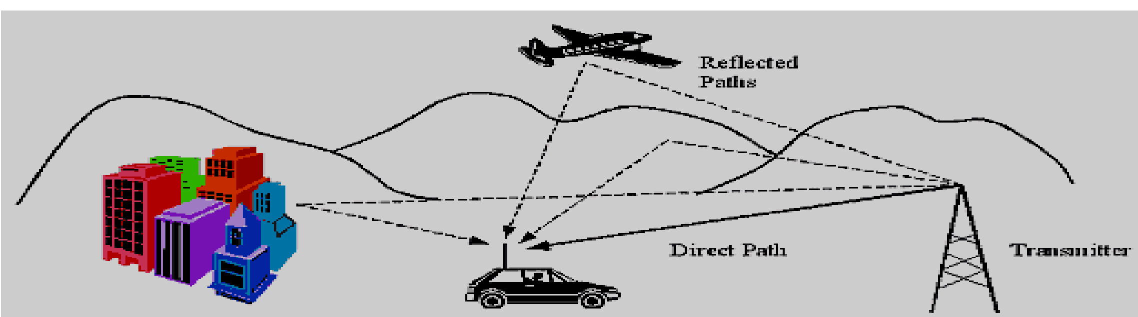


Figure 4.1: Multi path fading channel

The overall signal at the radio receiver is a summation of the variety of signals being received. As they all have different path lengths, the signals will add and subtract from the total dependent upon their relative phases.

At times there will be changes in the relative path lengths. This could result from either the radio transmitter or receiver moving, or any of the objects that provides a reflective surface moving. This will result in the phases of the signals arriving at the receiver changing, and in turn this will result in the signal strength varying as a result of the different way in which the signals will sum together. It is this that causes the fading that is present on many signals. This can cause problems with phase distortion and Inter symbol interference (ISI) when data transmissions are made. As a result, it may be necessary to incorporate features within the radio communications system that enables the effects of these problems to be minimized.

4.1.2 Types of multi path fading

Propagation models have traditionally focused on predicting the average received signal strength at a given distance from the transmitter, as well as the variability of the signal strength in close spatial proximity to a particular location. Depends upon it, the multi path fading is mainly classified into 2 types.

A. Large scale fading
B. Small scale fading

4.1.2.1 Large-Scale Fading

Fading, that concentrates on the *mean* signal strength for an arbitrary transmitter-receiver (T-R) separation distance is useful in estimating the radio coverage area of a transmitter and is called large scale fading. It is mainly due to the absorption of RF energy in the ionosphere and is calculate by keeping the transmitter and receiver in the fixed positions. That’s why; it is also referred to as absorption fading. The absorption occurs due to 3 types of mechanisms. They are Reflection, Diffraction and Scattering.

Different fading models are developed to estimate this large scale fading such as Longley-rice model, Durkin's model, Okumura model, Hata model, Ericsson multiple breakpoints Models, Walfish and Bertoni model, etc…

4.1.2.2 Small - Scale Fading

Small-scale fading refers to the dramatic changes in signal amplitude and phase that can be experienced as a result of small changes (as small as half wavelength) in the spatial position between transmitter and receiver. The type of fading experienced by a signal propagating through a mobile radio channel depends on the nature of the transmitted signal with respect to the characteristics of the channel.

The fig 4.2 illustrates the relation between small scale and large scale fading.

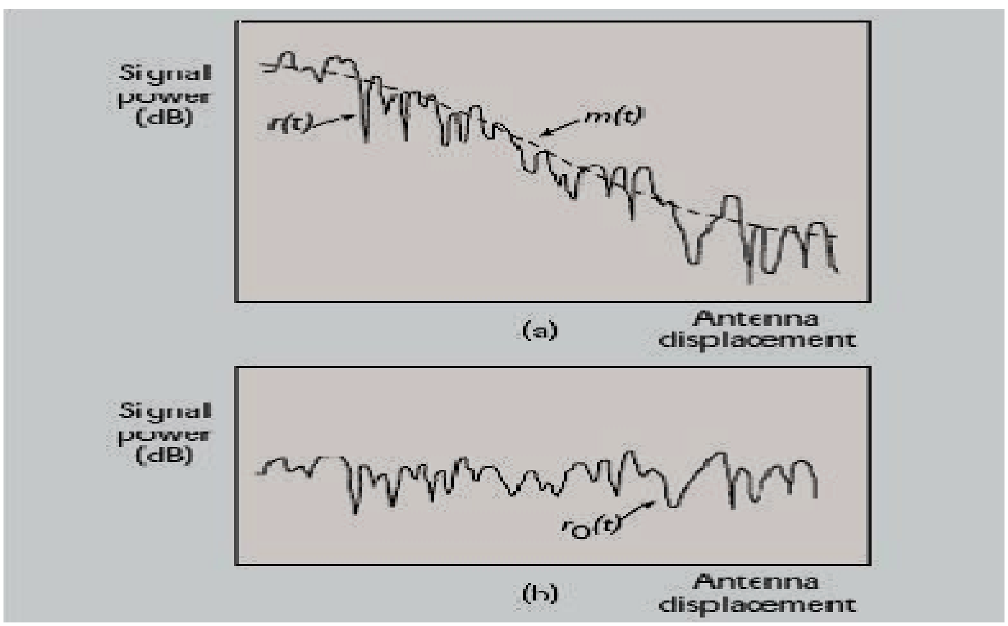


Figure 4.2: Large scale& small scale fading [5]

In fig 4.2a, Small scale fading is superimposed on large scale fading can be easily identified. In fig 4.2b large scale fading m(t) has been removed in order to view the small scale fading $r_o(t)$. Depending on the relation between the signal parameters (such as bandwidth, symbol period, etc.) and the channel parameters (such as rms delay spread and

Doppler spread), different transmitted signals will undergo different types of fading. The time dispersion and frequency dispersion mechanisms in a mobile radio channel lead to four possible distinct effects, which are manifested depending on the nature of the transmitted signal, the channel, and the velocity. While multipath delay spread leads to *time dispersion* and *frequency selective fading*, Doppler spread leads to *frequency dispersion* and *time selective fading*. The two propagation mechanisms are independent of one another. Table 2.1 shows the types of fading.

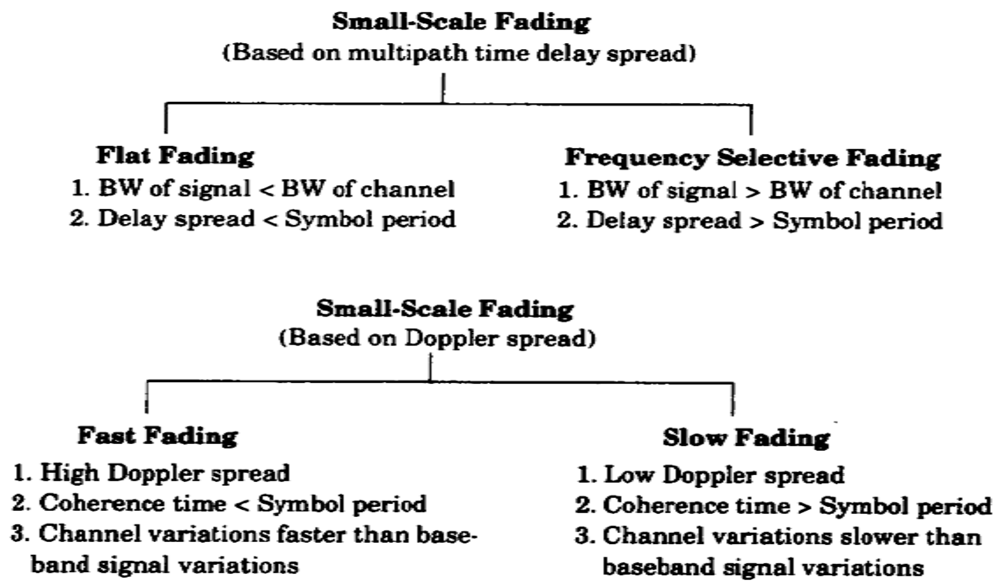


Table 4.1: Types of small scale fading [5]

Fading effects due to multi path time delay spread

As discussed above, multipath time delay spread can affect *radio communications channels* and it can be classified as

- Flat fading
- Frequency selective fading

Flat fading

If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal, then the received signal will undergo *flat fading*. In flat fading, the multipath structure of the channel is such that the spectral characteristics of the transmitted signal are preserved at the receiver. However the strength of the received signal changes with time, due to fluctuations in the gain of the channel caused by multipath.

It can be seen from Figure 2.3 that if the channel gain changes over time, a change of amplitude occurs in the received signal. Over time, the received signal $r(t)$ varies in gain, but the spectrum of the transmission is preserved. In a flat fading channel, the reciprocal bandwidth of the transmitted signal is much larger than the multipath time delay spread of the channel, and $h_b(t_\tau)$ can be approximated as having no excess delay (i.e., a single delta function with τ=0).

Flat fading channels are also known as *amplitude varying channels* and are sometimes referred to as *narrowband channels*, since the bandwidth of the applied signal is *narrow* as compared to the channel flat fading bandwidth. Typical flat fading channels cause deep fades, and thus may require 20 or 30 dB more transmitter power to achieve low bit error rates during times of deep fades as compared to systems operating over non-fading channels. The distribution of the instantaneous gain of flat fading channels is important for designing radio links, and the most common amplitude distribution is the Rayleigh distribution. The Rayleigh flat fading channel model assumes that the channel induces amplitude which varies in time according to the Rayleigh distribution.

To summarize, a signal undergoes flat fading if

$$B_S \ll B_C \text{ and } T_S \gg \sigma_\tau \quad (4.1)$$

Where Ts is the reciprocal bandwidth (e.g., symbol period) and B_s is the bandwidth, respectively, of the transmitted modulation, and σ_τ and B_C are the rms delay spread and coherence bandwidth, respectively, of the channel.

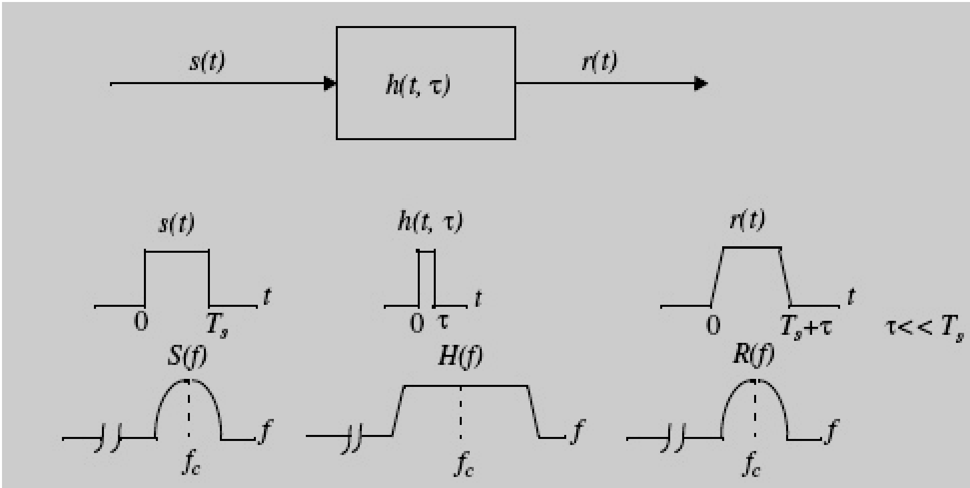


Figure 4.3 Flat fading channel characteristics. [5]

Frequency Selective Fading

If the channel possesses a constant-gain and linear phase response over a bandwidth that is smaller than the bandwidth of transmitted signal, then the channel creates frequency selective fading on the received signal. Under such conditions, the channel impulse response has a multipath delay spread which is greater than the reciprocal bandwidth of the transmitted message waveform. When this occurs, the received signal includes multiple versions of the transmitted waveform which are attenuated (faded) and delayed in time, and hence the received signal is distorted.

Frequency selective fading is due to time dispersion of the transmitted symbols within the channel. Thus the channel induces *inter symbol interference* (ISI). Viewed in the frequency domain, certain frequency components in the received signal spectrum have greater gains than others.

Frequency selective fading channels are much more difficult to model than flat fading channels since each multipath signal must be modeled and the channel must be considered to be a linear filter. It is for this reason that wideband multipath measurements are made, and models are developed from these measurements. When analyzing of the mobile communication systems, statistical impulse response models such as the two-ray Rayleigh fading model (which considers the impulse response to be made up of two delta functions which independently fade and have sufficient time delay between them to induce frequency selective fading upon the applied signal), or computer generated or measured impulse responses, are generally used for analyzing frequency selective small-scale fading. Figure 4.4 illustrates the characteristics of a frequency selective fading channel.

For frequency selective fading, the spectrum *S (f)* of the transmitted signal has a bandwidth which is greater than the coherence bandwidth B_c of the channel. Viewed in the frequency domain, the channel becomes frequency selective, where the gain is different for different frequency components. Frequency selective fading is caused by multipath delays which approach or exceed the symbol period of the transmitted symbol. Frequency selective fading channels are also known as *wideband channels* since the bandwidth of the signal $s(t)$ is wider than the bandwidth of the channel impulse response. As time varies, the channel varies in gain and phase across the spectrum of $s(t)$, resulting in time varying distortion in the received signal $r(t)$. To summarize, a signal undergoes frequency selective fading if

$$B_S > B_C \text{ and } T_S < \sigma_\tau . \tag{4.2}$$

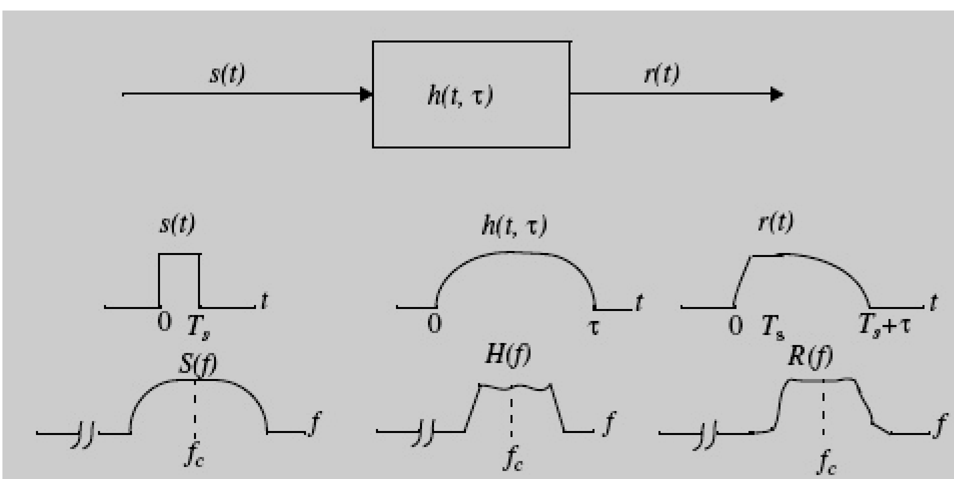


Figure 4.4 Frequency selective fading channel characteristics [5]

A common thumb rule is frequency selective if T_S modulation used.that a channel is flat fading if $T_S \geq 10\sigma_\tau < 10\sigma_\tau$ although this is dependent on the and a channel is specific type of modulation used.

Fading Effects Due to Doppler Spread

Cellular telecommunications is subject to this type of fading effects. There are a variety of reasons for this. The first is that the mobile station or user is likely to be moving, and as a result the path lengths of all the signals being received are changing. The second is that many objects around may also be moving. Automobiles and even people will cause reflections that will have a significant effect on the received signal. Accordingly multipath fading due to Doppler spread has a major bearing on cellular telecommunications. According to Doppler spread, fading can be classified as

- Fast fading
- Slow fading

Fast Fading

Depending on how rapidly the transmitted baseband signal changes as compared to the rate of change of the channel, a channel may be classified either as a *fast fading* or *slow fading* channel. In a *fast fading channel*, the channel impulse response changes rapidly within the symbol duration. That is, the coherence time of the channel is smaller than the symbol period of the transmitted signal. This causes frequency dispersion (also called time selective fading) due to Doppler spreading, which leads to signal distortion. Viewed in the frequency domain, signal distortion due to fast fading increases with increasing Doppler spread relative to the bandwidth of the transmitted signal. Therefore, a signal undergoes fast fading if

$$T_S > T_C \text{ and } B_S < B_D \qquad (4.3)$$

It should be noted that when a channel is specified as a fast or slow fading channel, it does not specify whether the channel is flat fading or frequency selective in nature. Fast fading only deals with the rate of change of the channel due to motion. In the case of the flat fading channel, we can approximate the impulse response to be simply a delta function (no time delay). Hence, a *flat fading, fast fading* channel is a channel in which the amplitude of the delta function varies faster than the rate of change of the transmitted baseband signal. In the case of a *frequency selective, fast fading* channel, the amplitudes, phases, and time delays of any one of the multipath components vary faster than the rate of change of the transmitted signal. In practice, fast fading only occurs for very low data rates.

Slow Fading

In a *slow fading channel*, the channel impulse response changes at a rate much slower than the transmitted baseband signal *s(t)*. In this case, the channel may be assumed to be static over one or several reciprocal bandwidth intervals. In the frequency domain, this implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signals. Therefore, a signal undergoes slow fading if

$T_S << T_C$ and $B_S >> B_D$ (4.4)

It should be clear that the velocity of the mobile (or velocity of objects in the channel) and the baseband signaling determines whether a signal undergoes fast fading or slow fading. The relation between the various multipath parameters and the type of fading experienced by the signal are summarized in Figure 2.5. Over the years, some authors have confused the terms fast and slow fading with the terms large-scale and small-scale fading. It should be emphasized that fast and slow fading deal with the relationship between the time rate of change in the channel and the transmitted signal, and not with propagation path loss models.

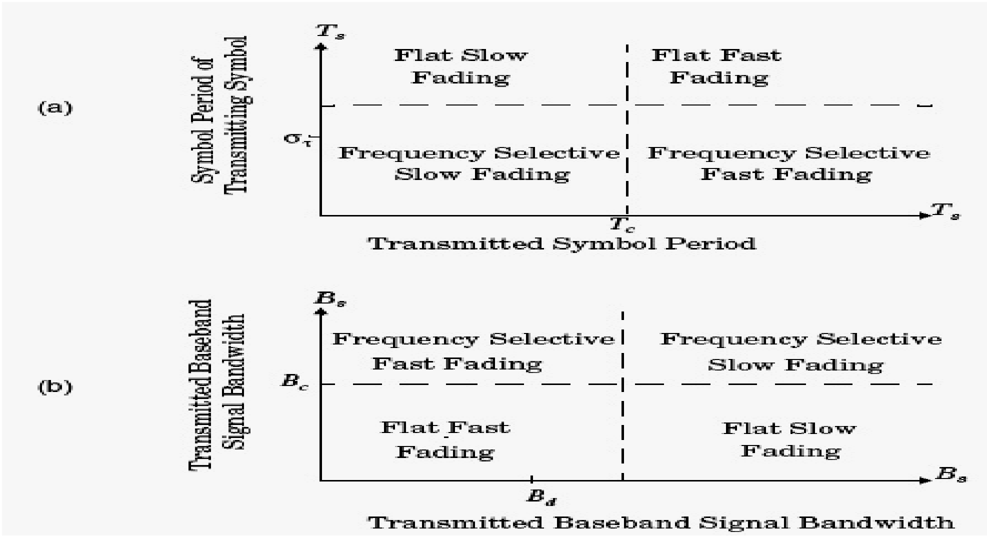


Figure 4.5 Type of fading experienced by a signal as a function T_s and B_s

4.2. Different Channel Models

There are basically two factors affecting the radio-wave propagation shadowing and multipath fading. Mathematical model (statistical modeling), has been developed based on the particular propagation environment and the underlying communication scenario. Which can be listed as envelope and phase fluctuations over time? For coherent modulation system phase distortion becomes a severe problem. In this section we will take different channel model and their basics. The coherence time and Doppler are related by formula Tc= 1/Fd. Now let Ts be symbol period then fading is said to be slow if Ts is smaller than the channel's coherence time Tc otherwise, it is considered to be fast. In slow fading a particular fade level will affect many successive symbols, which leads to burst errors, whereas in fast fading the fading deco relates from symbol to symbol. Again if bandwidth is much smaller than the channel's coherence bandwidth fc .the channel is known as frequency flat otherwise it frequency selective. For flat fading, channel is modeled by AWGN, which is typically assumed to be statistically independent of the fading amplitude $_{\alpha}$ and which is characterized by a one-sided power spectral density N0 (W/Hz). The amount of fading (AF), or "fading figure," associated with the fading PDF is defined as radio of var(α^2) to E^2 (α^2), where E(.),represent expectation. On other hand frequency selective channel can be classified as Rayleigh, Rician Nakagami. In Rayleigh channel signal is passed through fading. In simulation it is model as following Rayleigh channel object h with an input sample period Ts, a maximum Doppler shift fd, a vector of path delays tau, and a vector of average path power gains pdb (in dB). It then uses the Doppler Gaussian function to create a Gaussian Doppler object with a standard deviation sig mag, and assigns it to the Doppler Spectrum property of the constructed channel object. However Rayleigh distribution pdf is given as [42]

$$\Pr(R) = (2r/\Omega)\exp\,(r^2/\,\Omega) \qquad (4.5)$$

and used if no line of sight between transmitter and receiver. Next channel model is Rician distribution in which some line of sight signal may be presented however in this case severity is less than that of Rayleigh distribution. in this channel dominant component of signal is much stronger than other multipath component. The pdf is giaven as following

$$P(r) = \left(\frac{r}{\sigma 2}\right) exp - \frac{r^2 + A^2}{2\sigma} \times Io\left(\frac{Ar}{\sigma^2}\right) \qquad (4.6)$$

Where Io(.) is the modified Bessel function of the first kind and zero order . K parameter is given by : $K = A^2/2\sigma$

Above described fading models are not suitable for long distance. For long distance purpose M..Nakagami formulated a parametric gamma function based on high Frequency and long distance. The Nakagami distribution, also known as the “m-distribution,” provides greater flexibility in matching experimental data. Experimental results have shown that the Nakagami distribution fits experimental data collected in a variety of fading better than Rayleigh, Rician, or Lognormal distributions. The Nakagami family distributions span from the one-sided Gaussian distribution (m=1/2) to the non fading channel case (m=infinite) Rayleigh fading (m=1) as a special case; along with the cases of fades that are more severe than Rayleigh (1/2≤m≤1) and fades that are less severe than Rayleigh (1< m). They can also be used as an approximation to lognormal and Rician distributions for a certain range of average SNRs. The probability density function (PDF) for this distribution is given by Nakagami[43]

$$P_R(r) = \frac{m^m r^{2m-1}}{\Omega^m \Gamma m} exp - \{mr^2/\Omega\} \qquad (4.7)$$

Rayleigh fading is particular case of Rician and Nakagami distributions. The Nakagami model can be used to approximate the PDF of the power of a Rician fading signal. The Rician K-factor is defined as the ratio of signal power in dominant component to the scattered power. In indoor channels with an unobsuucted line-of-sight between transmit and receive antenna, the K-factor is between, say, 4 and 12 dB. Rayleigh fading is the special case of Rice fading for which K = 0 (-infinity dB). Nakagami distribution is a good model for outdoor in urban areas where line of-sight (LOS) does not exists (Rayleigh, m=1). Nakagami distribution with large m is a good model for indoor propagation where LOS exists and in mobile satellite systems.[44]

4.3 Simulation Model of Nakagami Channel

As Nakagami fading is generalized channel model representing various fading condition in wireless channel, so we have taken simulation and analysis of the Nakagami fading channel. In this section. Following are required steps for MATLAB simulation of Nakagami fading channel.

1. Generating the received signal using equation

$$s_{ray}(t) = \sum_{i=1}^{N} a_i \cos(\omega_c t + \omega_{di} t + \varphi_i) \tag{4.8}$$

Using Toolbox of MATLAB, we can easily obtained the distribution of path amplitudes a_i phase φ_i and the ω_{di} Doppler shift,of i^{th} wave where a_i, taken to be phases has been generated with function uniform (statistics tool-box) and Doppler shift is taken symmetrical.[44]

2. The command demon from the Signal Processing Toolbox has been taken to demodulated signal and hence to get the in phase and quadrature components,
3. With above two steps the Rayleigh and Ricean fading envelop can be obtain and then superimposed according to

$$R_{Nakga} = R_{ray} e^{1-m} + R_{rice}(1 - e^{1-m}) \tag{4.9}$$

to get Nakagami fading channel.

CHAPTER 5 - RESULTS AND DISCUSSION

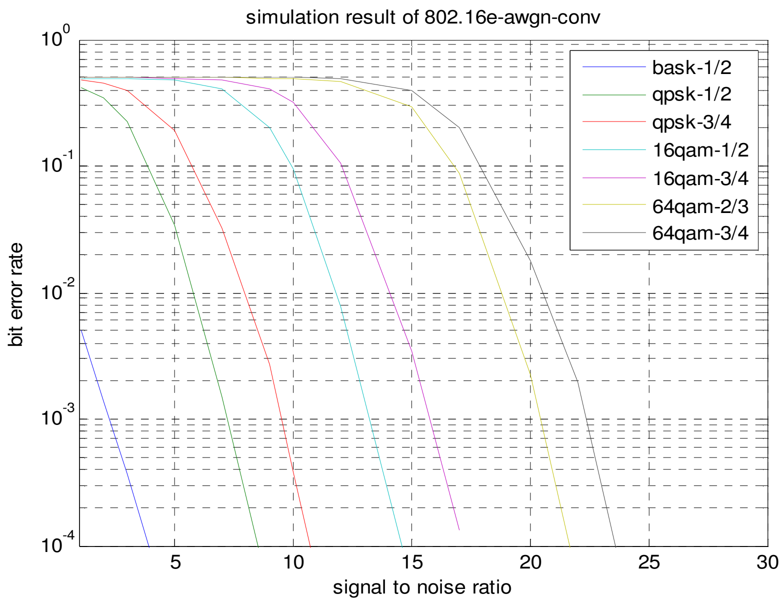


Fig: 5.1a BER for AWGN with CC

Above fig. shows the SNR vs BER graph for different modulation techniques, for AWGN channel.

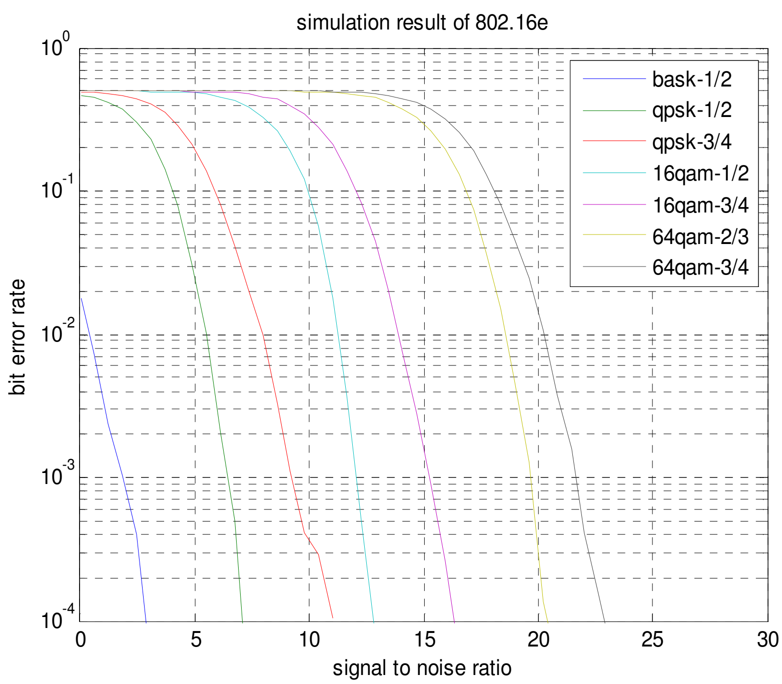


Fig: 5.1b BER for AWGN with CC-RS

From fig 5.1a and 5.1b it is very clear that using FEC,RS over CC improve the BER more or less for all the modulation techniques.

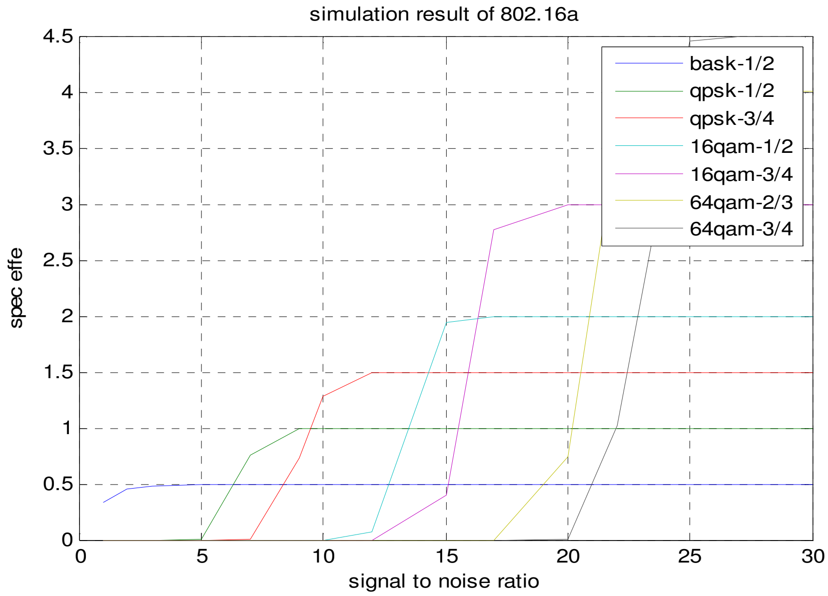


Fig: 5.2a Spectral efficiency for AWGN with CC

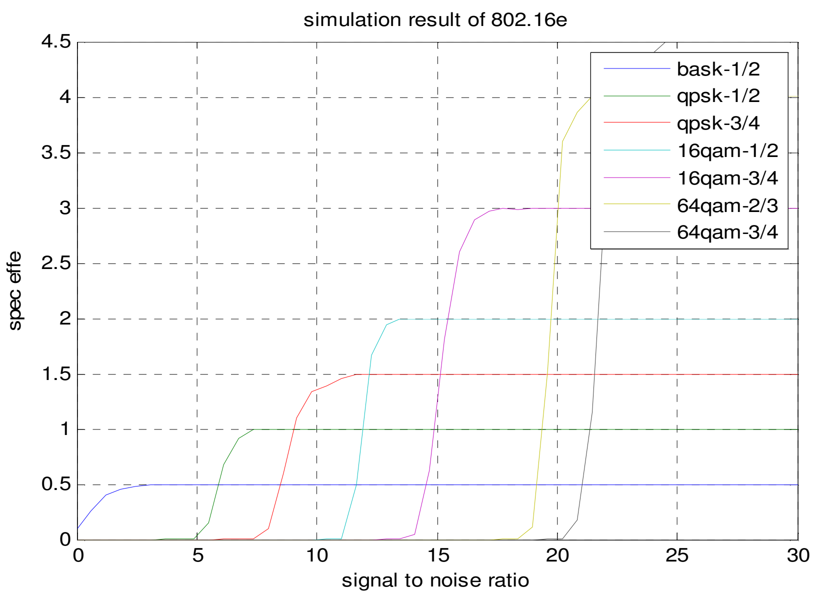


Fig: 5.2b Spectral efficiency for AWGN with CC-RS

Fig. 5.2a and 5.2b, shows the spectral efficiency vs SNR analysis. For AWGN channel, with CC alone and CC-RS respectively.

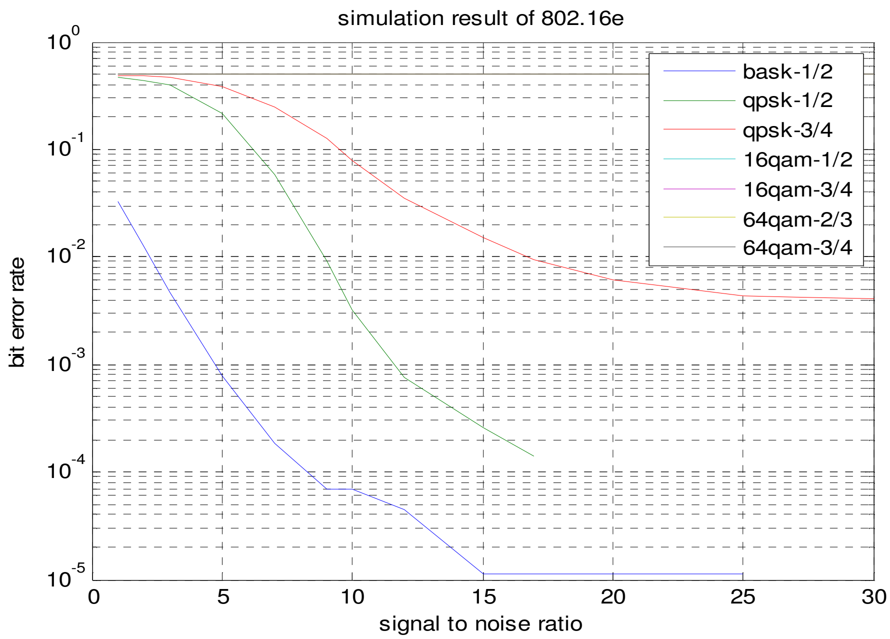


Fig: 5.3a BER for Rayleigh with CC-RS

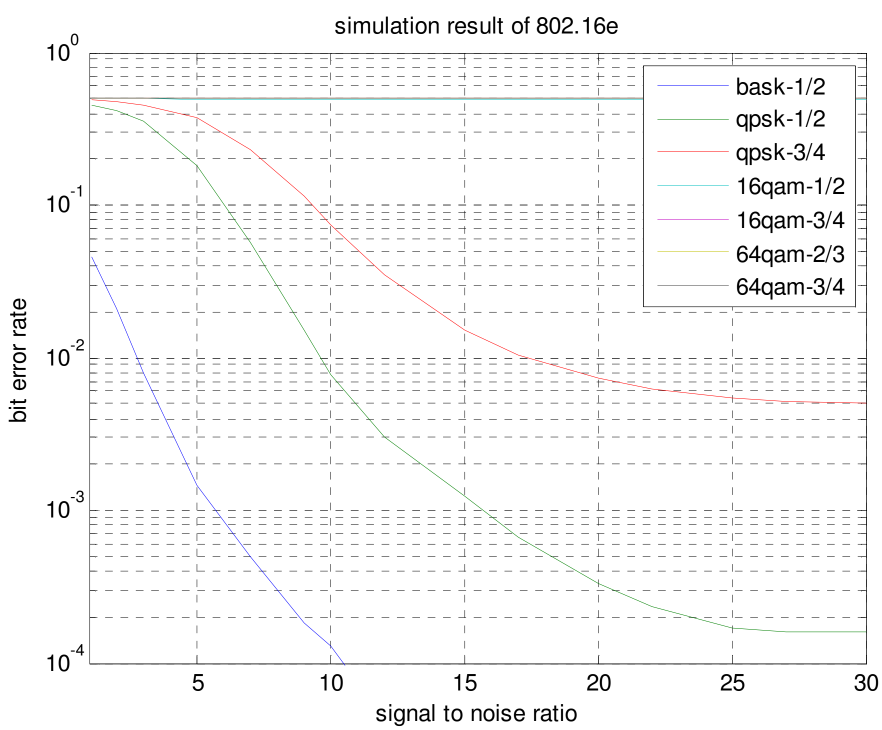


Fig: 5.3b BER for Rayleigh with CC

From above two figures it is clear there is at SNR 10 dB for ,BPSK, BER decreases FROM 0.001242 to 0.0002609. also for other modulations it shows similar results. **59**

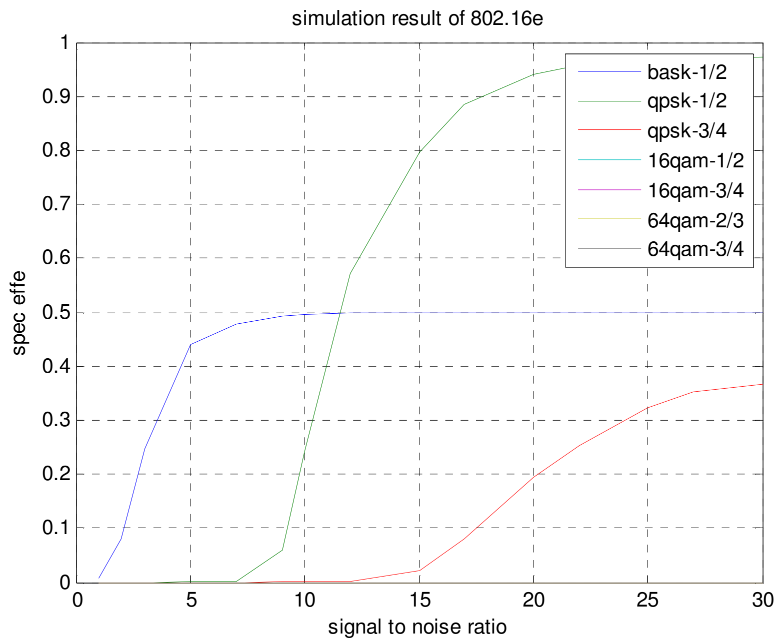


Fig: 5.4a Spectral efficiency for Rayleigh

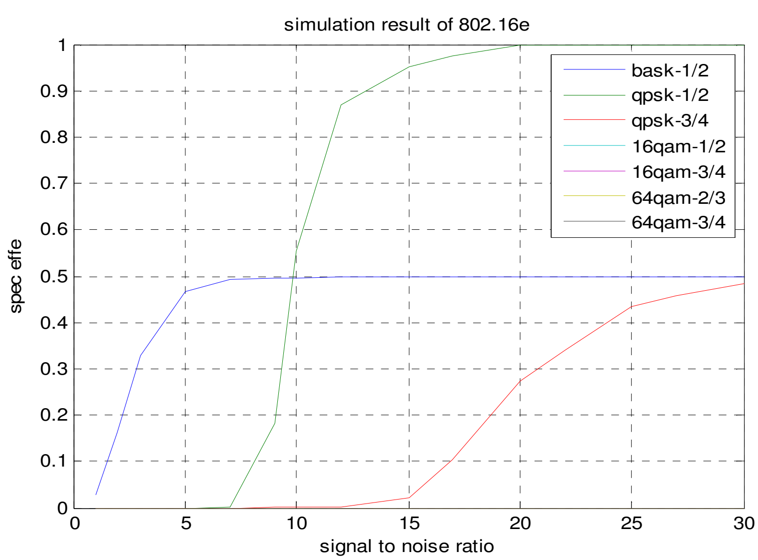


Fig: 5.4b Spectral efficiency for Rayleigh CC-RS

From above fig. it is noted that for BPSK, at SNR 10 dB there is 60% loss of spectral efficiency as a result of CC-RS ,FEC coding.

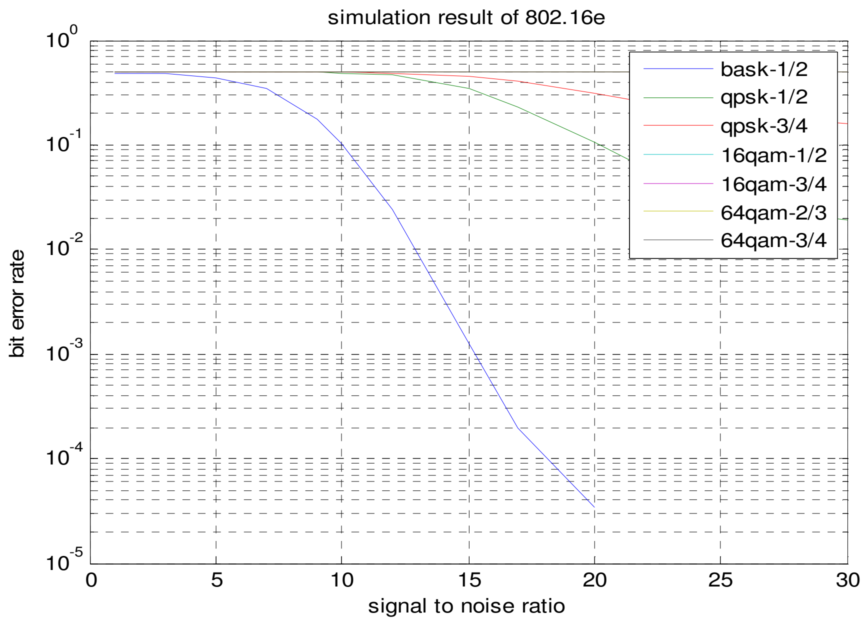


Fig: 5.5a BER for Nakagami with CC

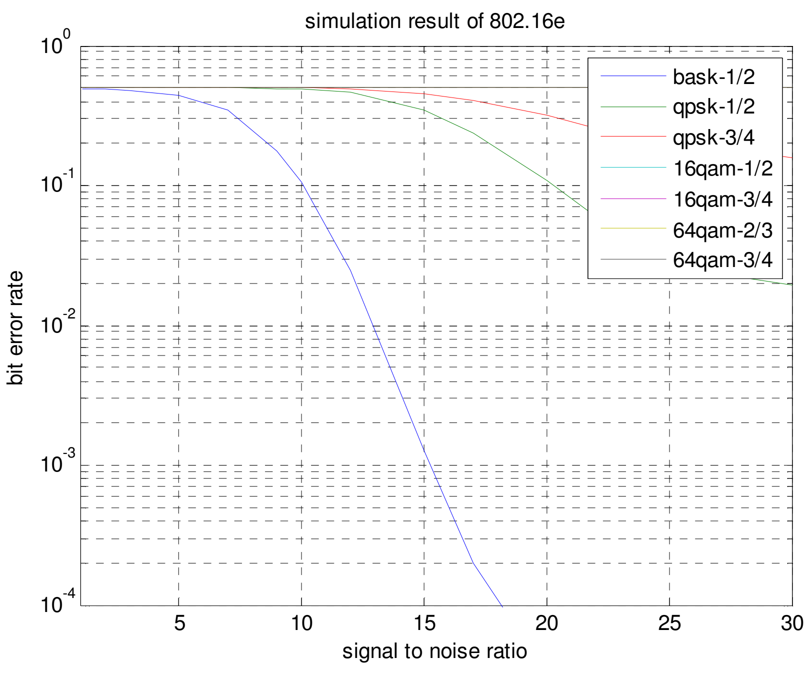


Fig: 5.5b BER for Nakagami with CC-RS

Above fig. shows the SNR vs BER graph for most practical channel Nakagami for different FEC coding, it can be recorded that CC-RS over improves the BER 61

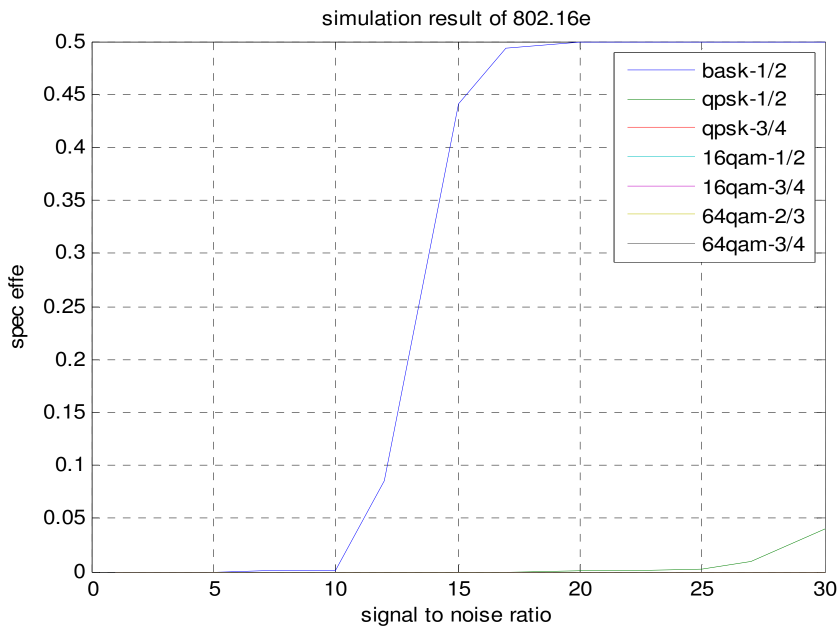


Fig: 5.6 a Spectral efficiency for Nakagami

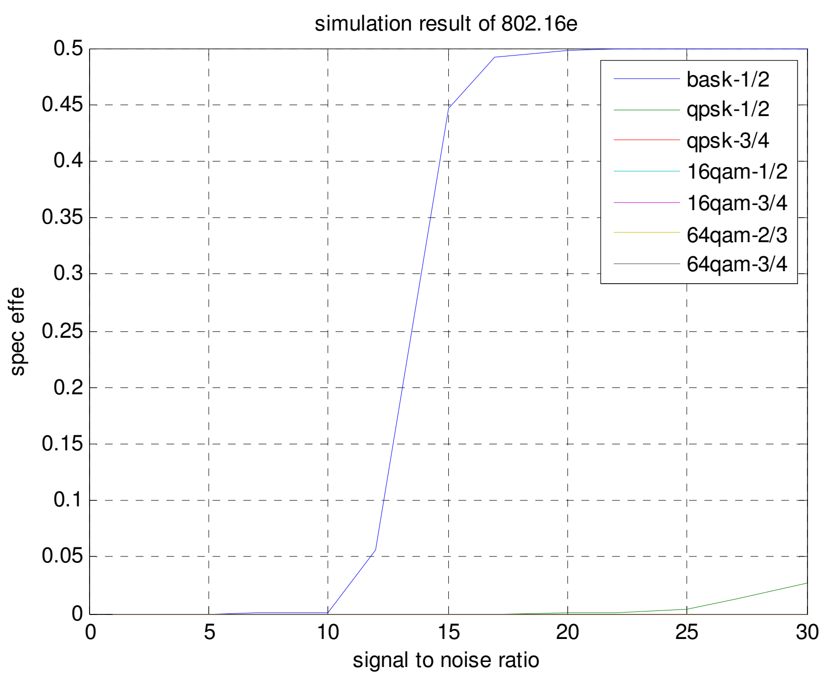


Fig: 5.6b Spectral efficiency for Nakagami CC-RS

From above two figures for Nakagami FEC CC-RS over CC has negligible effect on spectral efficiency. **62**

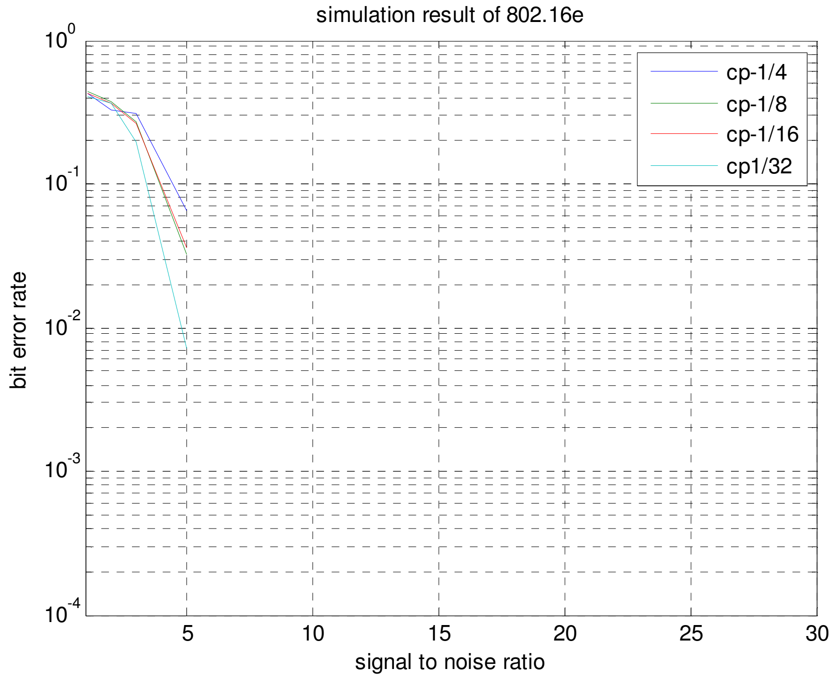


Fig: 5.7a Effect of CP on BASK modulation.

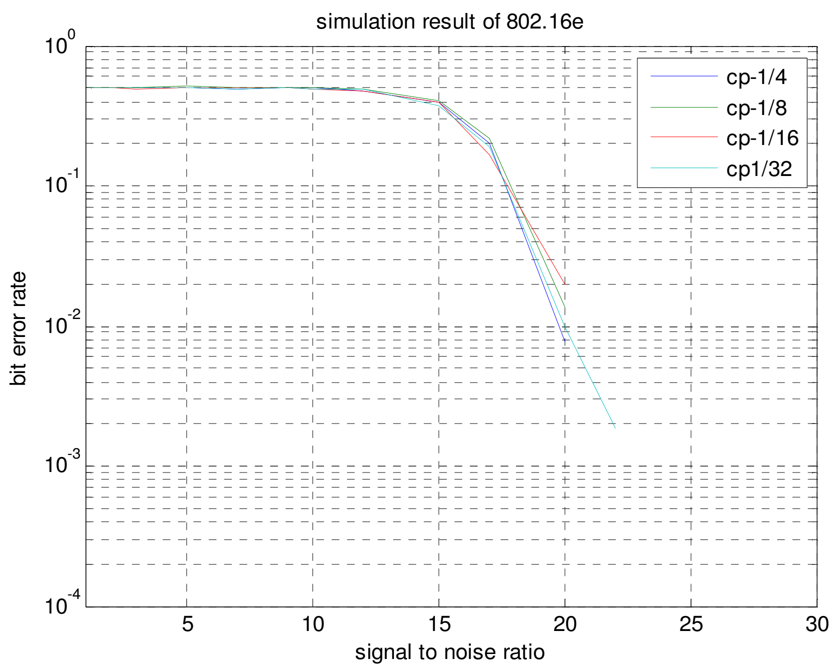


Fig 5.7b Effect of CP on 64QAM-3/4

Above fig. shows how CP affect the BER performance. In fig 5.7a it is shown for BASK and in 5.7b same for 64QAM-3/4 63

CHAPTER 6: CONCLUSION

Standard 802.16e, WiMAX is most suitable candidate as a standard for coming generation mobile system. Practically, ICI, PAPR and Frequency-Offset are major issues for performance evaluation of a wireless communication system. In analysis, these factors are evaluated in terms of BER. At the same time crises of “spectrum” cannot be ignored. In this thesis, the key contribution was the implementation of the IEEE 802.16e OFDM PHY layer using MATLAB in order to evaluate the PHY layer performance under different channel model, like AWGN, Rayleigh and Nakagami. The Nakagami is best fading channel for practical environment. There are two reasons behind it. First, Nakagami channel is closest to the experimental results compare to all fading distribution in literature (like Ryaleigh, Ricean and Lognormal). Secondly Nakagami distribution is a good model for outdoor (NLOS) and indoor (LOS) both. Hence gradual effect starting from ideal AWGN to practical one Nakagami has been targeted.

For every channel model BER analysis has been done, for this purpose different modulation technique has been taken. But as it is known that an increased in modulation level results increase in BER. For example in case of AWGN the BER at SNR 10 dB BER increases 0.00064 to 0.09418 as modulation changes from QPSK-1/2 to 16QAM-1/2 RS channel coding over CC has been implemented in order to reduce the BER for a given modulation level. For same sets of data BER get improved to 0.01196. The Similar results have found for other channels too. So the BER analysis, as function of modulation techniques, as function of different channel coding and also as function of different channel models was first goal of this thesis.

On the basis of above parameters Spectral Efficiency analysis was the second goal of thesis. For example in case of Rayleigh channel for QPSK-1/2, of 6-8% of spectral efficiency saving, has been recorded at SNR value 20dB.

Hence, it has been concluded that there is 87% improvement in BER at the cost of 6-8% spectral efficiency, for above example. A similar conclusion can be drawn for all other cases too.

In future, there is great scope to increase the spectral efficiency, of WiMAX system. The spectral shaping may be the most efficient tool for this purpose.

REFERENCES

[1] Masahiro Umehira and Takatoshi Sugiyama "**OFDM/CDMA Technologies for Future Broadband Mobile Communication Systems"** *IEICE TRANSACTIONS* on Fundamentals of Electronics, Communications and Computer Sciences Vol.E85-A No.12 pp.2804-2812, January 2002

[2] Richard Van Nee, R. Prasad "OFDM for Wireless Multimedia Communications", *Artech. House, Inc. Norwood*, MA, USA Boston,2000.

[3] White Paper: High-speed Wireless OFDM communication Systems, Wi-LAN *Inc.*, *IEEE trans. Theory*, vol.13, No.11, pp.2135-2137, February 2001.

[4] P.H. Moose, "A Technique for OFDM Frequency Offset Correction", *IEEE Transactions on Communications*, vol.42, no.10, pp. 1009-1010, October 1994.

[5] B.P. Lathi, Modern Digital and Analog Communication systems, *CBS College Publishing*, ISBN:0-03-058969-X, 1983

[6] Martin S. Roden, Analog and Digital Communication Systems*, Prentice Hall International, Inc*, 4th edition, ISBN:0-13-399965-3, 1996

[7] Yoshihiko Akaiwa, "Introduction to Digital Mobile Communication", *John Wiley & sons, Inc*, ISBN: 0-471-17545-5, pp. 3109-311,1997

[8] "*IEEE Standard 802.16 for Global Broadband Wireless Access*, "http://ieee802.org/16/docs/03/C8021603_14.pdf"

[9] Roger B. Marks, Brian G. Kiernan, *and* Carl J. Bushue, IEEE Std. 802.16-2001, "Standard for Local and Metropolitan area networks Part 16: Air

Interface for Fixed Broadband Wireless Access Systems", *IEEE Std. 802.16-2001*, December 2001

[10] Roger B. Marks, Ken Stanwood and Chang IEEE Standard for Local and Metropolitan Area Networks Part16: Air Interface for Fixed Broadband Wireless Access Systems", *IEEE 802.16-2004*, 1 October, 2004

[11] Kenneth Stanwood and Lars Lindh, "IEEE Standard for Local and metropolitan area networks Part 16: Air Interface for Fixed and Mobile Broadband Wireless Access Systems Amendment 2: Physical and Medium Access Control Layers for Combined Fixed and Mobile Operation in Licensed Bands and Corrigendum 1" *IEEE Std 802.16e-2005and IEEE Std 802.162004/Cor. 12005(Amendment and Corrigendum to IEEE Std 802.162004)* , February 2006

[12] Roger B. Marks, Carl Eklund and Dean Chang, "IEEE Standard for Local and metropolitan area networks Part 16: Air Interface for Fixed Broadband Wireless Access Systems Amendment 2: Medium Access Control Modifications and Additional Physical Layer Specifications for 2-11GHz", *IEEE Std802.16a2003 (Amendment to IEEE Std. 802.162001),* January 2003

[13] **http://www.wimaxindustry.com/wp/papers/intel_80216_wimax.pdf,** *IEEE* 802.16 and WIMAX

[14] Mukul Kabra and S.L. Maskara, "Enhancing Data Rates using OFDM Techniques for Various Applications" : Dhirubhai Ambani Institute of Information and Communication Technology *IET computer society,* vol. 534, pp. 223-226, April 2005

[15] Leonard J.and Cimini, JR, "Analysis and Simulation of a Digital Mobile Channel Using Orthogonal Frequency Division Multiplexing", *IEEE Transactions on Communications* vol. Comm.-33,no-7,july-1985

[16] A. Batra et al. "Multi-band OFDM physical layer proposal for IEEE 802.15 task group 3a", *tech. rep., IEEE P802.15-03/268r3*, March 2004.

[17] R.W. Chang, and R.A. Gibby, "Theoretical Study of Performance of an Orthogonal Multiplexing Data Transmission Scheme," *IEEE Transactions on Communications*, **16**, 4, pp. 529-540, April 1968

.[18] Hikmet Sari, "Characteristics and Compensation of Multipath Propagation in Broadband Wireless Access System", *ECPS 2005 Conference,* pp15-18 March, 2005

[19] V. Erceg, K.V.S. Hari, M.S. Smith and D.S. Baum "Channel Models for Fixed Wireless Applications", *IEEE 802.16.3 Task Group Contributions*,Feb 2001

[20] J. Jang and K.B. Lee, Transmit power adaptation for multiuser OFDM systems, *IEEE J. Select. Areas Commun.,* 21(2), 171, 2003.

[21] D. Tse and P. Viswanath, "Fundamentals of Wireless Communication," *Cambridge University Press, New York,* 2005.

[22] H. Liu and G. Li, "OFDM-Based Broadband Wireless Networks," *John Wiley & Sons, New York*, 2005

[23] Koffman, I. and Roman, V.,Broadband wireless access solutions based on OFDM access in IEEE 802.16" *Communications Magazine, IEEE, Vol.40, Iss.4*, Apri 2002,Pages:96-103

[24] ETSI Broadband Radio Access Networks (BRAN); HIPERMAN; Physical (PHY) Layer. *Standard TS 102 177,* 2003

[25] Alan Barbieri, Dario Fertonani, Giulio Colavolpe, "Spectrally-Efficient Continuous Phase Modulations,*"IEEE* trans. wireless communications, vol. 8, no. 3, March 2009.

[28] M.M.K Howlader and Haroon quereshi, "Spectral efficient OFDM system using M-ary PSK modulation," *IEEE trans.*2003.

[29] John H.Gass, Christopher and J.Langford " An Application of Turbo Trellis-Coded Modulation to Tactical Communication" *ITT Industries Aerospace/Communications Division,*2001

[30] Hesham El Gamal, Bassel F. Beidas, and Stan Kay, "Turbo Decoding for High Spectral Efficiency Satellite Communications," *IEEE* trans. 2000 PP. 440-445.

[31] Noboru Izuka, Yoshimasa, Shoji Mizuoo, Yuichi Yamakita, " Spectrum Shaping Method to improve the spectral efficiency of OFDM systems," *IEEE trans*. 2002 .

[32] Keukjoon Rang, Namshin Cho, Kilnam Oh, Jaehong Park, and Daesik Hong,“Improve the spectral efficiency for the OFDM system using the time and frequency domain equalization ,” *IEEE trans.,* PP. 468-468,1996

[33] Masanori Hamamural Shin'ichi Tachikawa', “Bandwidth Efficiency Improvement for Multi-carrier Systems” *IEEE Communication,* PP 48-52, 2004

[34] Saleem Ahmed, Razi Ur Rehman, Humor Hwang, “New Techniques to Reduce Sidelobes in OFDM System,” *IEEE*, *Third International Conference on Convergence and Hybrid Information Technology.* 2008

[35] Venkatesh Vadde, Steven Gray, “ Partial Response Signaling for Enhanced Spectral Efficiency and RF Performance in OFDM Systems,” *IEEE* 2001, pp 3120-3124.

[36] M.C. Valenti, “Channel coding for IEEE 802.16e mobile WiMAX,” a tutorial presented at *International Conference on Communications (ICC)* , June 18,2009

[37] Nathan J. Muller, “Wireless A to Z,” *McGraw-Hill, New York Chicago 2003.*

[38] Shih-Chang Huang, Chien Chen and Rong-Hong Jan “Draft IEEE Standard for Local and metropolitan area networks,” Part 16: Air Interface for Fixed Broadband Wireless Access Systems, Amendment for Physical and Medium Access Control *IEEE 802.16e/D7*,28 Feb2006.

[39] Parthia singe, Dharmawansa, Nangana Rajatheva and Hlaing Minn, “An Exact Error Probability Analysis of OFDM Systems with Frequency Offsets,”, ***IEEE Transactions on Communications*** ,Volume 57 , Issue 1, Pages:26-31,January 2009

[40] Luca Rugini, Member IEEE, and Paolo Banelli, "BER of OFDM Systems Impaired by Carrier Frequency Offset in Multipath Fading Channels," *IEEE Trans.* on Wireless Communications, vol. 4, no. 5, September 2005.

[41] Marvin K. Simon , Mohamed-Slim and Alouini "Digital Communication over Fading Channels A Unified Approach to Performance Analysis" *A Wiley-Interscience Publication JOHN WILEY & SONS, INC. New York*

[42] M. Patzold, "*Mobile fading channels-modeling, analysis& synthesis*", John Wiley, 2002.

[43] S. Elnoubi, S. Abou Chahine and H. Abdallah "BER Performance of GMSK in Nakagami Fading Channels" *2lSt National Radio Science Conference (NRSC2004 (NTI)* March 16-18,2004

[44] Cyril-Daniel and Iskander "A MATLAB based Object-Oriented Approach to Multipath Fading Channel Simulation", *IEEE Trans* 2006