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**Analysis of channel estimation for OFDM system in
WiMAX application**

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Analysis of channel estimation for OFDM system in WiMAX application

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Jerusalem/Palestine

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Dedication

I dedicate this work to my parents, fiance and my brothers.

Hala M. Mahmoud

Declaration:

I Certify that this thesis submitted for the degree of Master, is the result of my own research, except where otherwise acknowledged, and that this study (or any part of the same) has not been submitted for a higher degree to any other university or institution.

Signed:

Hala M. Abed Al Kareem Mahmoud

Date:

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ABSTRACT

With the rapid growth of digital communication in recent years the need for high speed data transmission has been increased. Moreover, future wireless systems are expected to support a wide range of services such as video, data and voice. OFDM is a transmission technique based on many orthogonal carriers that transmit simultaneously. OFDM system achieves high data rate in mobile environment, due to its resistance to Inter-Symbol Interference (ISI), which is a common problem in the high speed data communication. This master thesis analyzes different channel estimation techniques in OFDM system with parameter settings according to IEEE 802.16e and WiMAX. Channel estimation technique for OFDM systems over frequency selective Rician fading and Rayleigh fading channel, based on comb type pilot arrangement is investigated. Channel estimation methods may be classified as blind, semi-blind or pilot-aided. Blind algorithms do not require any training data and exploit statistical or structural properties of communication signals. Pilot-aided methods, on the other hand, rely on a set of known symbols interleaved with data in order to acquire the channel estimate. Semi-blind methods combine a blind criterion with limited amount of pilot data, which improves both effective data rates and convergence speed. Pilot-aided methods, can be performed by many ways, either inserting pilot tones into all of the subcarriers of OFDM symbols with a specific period or by inserting pilot tones into each OFDM symbol. In this thesis, we explored comb pilot arrangements in details. The advantage for comb type pilots arrangement in channel estimation is the ability to track the variation of the channel caused by Doppler frequency. It is observed that the Doppler effect can be reduced, which will increase the system mobility. Doppler frequency is the main reason for Inter-Carrier Interference (ICI). It is observed that ICI increases the noise level. Hence, one way to compensate ICI is to increase the number of pilots inserted. The estimation of the channel at the pilots frequency is based on Least Square (LS), and Kalman estimation methods. Kalman estimation outperforms LS estimation, the estimators perform about the same for SNR lower than 10 dB. This is an interesting property which means that the choice of channel estimator is not that important in terms of symbol errors for low Signal to Noise Ratio (SNR). When choosing a channel estimating method for low SNR the focus should instead be on how much information the estimating methods needs and also how high its complexity is. Three interpolation methods have been used to interpolate the channel response at the data frequency; linear interpolation; spline interpolation and low pass interpolation. The performance of the interpolation methods are compared, low pass interpolation has better performance than other methods. We have compared the performances of all schemes by measuring Bit error Rate (BER) with M-Quadrature Amplitude Modulation (M-QAM), and M-Phase Shift keying (M-PSK) as modulation schemes. The of Doppler frequency effect and pilots spacing are evaluated. The proposed algorithm of using comb type pilots arrangement and Kalman estimator achieves good performance and high mobility system with reasonable complexity compared with other systems.

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Abbreviations

3G	Third Generation
4G	Fourth Generation
AAA	Authorization, Authentication and Accounting
ADSL	Asynchronous Digital Subscriber Line
AMPS	Advanced Mobile Phone Services
APSB	Aided Pilot Symbol Blinded
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BW	Bandwidth of Transmitted Signal
BWA	Broadband Wireless Access
CDMA	Code Division Multiple Access
CFO	Carrier Frequency Offset
CFR	Channel Frequency Response
CIR	Channel Impulse Response
CP	Cyclic Prefix
DAB	Digital Audio Broadcasting
DFT	Discreet Fourier Transform
DPSK	Deferential Phase Shift Keying
DVB	Digital Video Broadcasting
EKF	Extended Kalman Filter
FDMA	Frequency Division Multiple Access
FFT	Fast Fourier Transform
GPRS	General Packet Radio Services
GSM	Global system Mobile
HDTV	High Definition Television
IBI	Inter Block Interference
ICI	Inter Carrier Interference
IDFT	Inverse Discreet Fourier Transform
IFFT	Inverse Fast Fourier Transform

IN	Intelligent Network
ISDN	Integrated Services Digital Network
ISI	Inter Symbol Interference
LMMSE	linear Minimum Mean-Square Error
LMS	Least Mean Squares
LS	Least Squares
MAI	Multi Access Interference
MC-CDMA	Multi-Carrier CDMA
MIMO	Multi Input Multi Output
NLOS	Non Line Of Sight
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
P/S	Parallel to Serial
PSA	Pilot Symbol Aided
PSAM	Pilot Symbol Aided Modulation
PSK	Phase Shift Keying
QAM	Quadrature Amplitude Modulation
QOS	Quality Of service
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
SNR	Signal to Noise Ratio
SOFDMA	Scalable Orthogonal Frequency Division Multiple Access
TDD	Time Division Duplex
TDMA	Time Division Multiple Access
TIA	Telecommunications Industry Association
UMTS	Universal Mobile Telecommunication System
VoIP	Voice Over IP
WCDMA	Wide-band Code Division Multiple Access
Wi-Fi	Wireless-Fidelity
WiMAX	World Interoperability for Microwave Access
WLANs	Wireless Local Area Networks

WSS	Wide-Sense Stationary
WSSUS	Wide Sense Stationary Uncorrelated Scattering
Mbps	Mega Bits per seconds

Publications

1.Hala Mahmoud, Allam Mousa, Rashid Saleem, “Kalman Filter Channel Estimation Based On Comb-Type Pilots for OFDM System in Time and Frequency –Selective Fading Environments”, Mosharaka International Conference on Communications, Computers and Applications, MIC-CCA- Aug 2008.

Problem Statement

In wireless mobile communications, the channels are time and frequency selective, the transmitted signal propagating via multiple paths experiences various delays due to different lengths of the paths. This makes the channel frequency selective and causes ISI. The ISI may distort the received signal so severely that the transmitted symbols cannot be recovered. Coherence bandwidth measures the frequency selectivity of the channel. Multicarrier modulation, such as OFDM, is a powerful technique to turn the frequency selective wireless channel into a set of frequency flat narrowband channels. Mobility causes the channel impulse response to be time-varying. Hence, it needs to be tracked over time. To achieve this the, knowledge of the Channel Impulse Response (CIR) is needed at the receiver in order to recover the transmitted data.

we concentrate on the following problem regarding to design channel estimation in pilot-aided OFDM wireless systems:

- The choice of how the pilot information should be transmitted. Pilot symbols along with the data symbols can be transmitted in a number of ways, and different patterns yielding different performances.
- CIR at the pilots subcarriers can be estimated by different methods such as LS; LMS; MMSE; and Kalman filter and each method has its advantages and disadvantages
- The design of an interpolation filter with both low complexity and good performance.

So, our proposed algorithm is suggested to increase the overall performance for OFDM system over time varying channel using comb type pilots arrangement, Kalman filter and the mean of interpolation .

The motivation is to :

- Estimate the effect of the Multipath fading channel in OFDM system in order to decrease the BER using Kalman filter .
- Reduce the effect of Doppler shift, and so increase the system mobility.
- Reduce the complexity of the interpolation by using one dimensional interpolator.

Justification

Fourth Generation (4G) is the next generation of wireless networks that will replace third Generation (3G) networks sometime in future. 4G is intended to provide high speed, and capacity. WiMAX allows interoperability and combines the benefits that other wireless networking technologies offer individually and leads a path towards 4G and become the 4G wireless technology in the future. The 4G/WiMAX spectrum will be used for voice, video and data. WiMAX can provide users with connectivity wherever they are and deliver the promise of 4G. WiMAX is based on OFDM technology paired with Multiple-Input Multiple-Output (MIMO) smart antenna technology which is best suitable for 4G.

OFDM is a transmission technique that is built-up by many orthogonal carriers that transmitted simultaneously, the ISI which caused by the multipath, in single carrier is a limiting factor in the performance of mobile wireless communication systems. The solution is the multicarrier systems based on OFDM, which transmit low data rate (large symbol time) on several overlapping orthogonal subcarriers. In addition a guard time is provided at the start of each symbol. By doing so, the symbol time is made large enough so that the system becomes less sensitive to multipath. A disadvantage of OFDM system is that usually the subcarriers will not be orthogonal when received at the receiver due to Doppler shift. Hence, this frequency offset has to be estimated. In a mobile fading channel, where the channel varies fast, the performance is highly degraded, and hence channel estimation is to be done to overcome the effect of fading. For this, an OFDM system has pilot symbols (on pilot subcarriers) embedded in between the data symbols (on data subcarriers), which provides the channel information at the receiver. This channel estimated values at the receiver, are interpolated over the data subcarriers and the data symbols are decoded. The basic idea with pilot symbols is that there is a strong correlation between the pilot symbol fading and the fading of information data symbols that are sent close to the pilot symbol in time and subcarrier. The estimation of the channel at the pilot frequencies will be based on LS and Kalman Filter.

Thesis outlines

In chapter 1, an introduction to wireless mobile communication systems is presented. An evolution of the mobile generations toward the fourth Generation (4G) is presented.

In chapter 2, an over view on fading in wireless environment, Radio waves propagate from a transmitting antenna, and travel through free space undergoing absorption, reflection, refraction, diffraction, and scattering.

In chapter 3, an introduction to OFDM principles, and the different between OFDM system and single carrier system.

In chapter 4, a description of an OFDM system over time-varying channel.

In chapter 5 This chapter considers channel estimation in OFDM transmissions. In communication systems, channel estimation methods may be classified as blind, semi-blind or pilot-aided. Our approach is to enhance the performance of the estimation process using Kalman estimator to estimate the CFR at the pilots, and comb pilots arrangement to track the time varying channel and so decrease Doppler effects.

In chapter 6, Simulation results.

In chapter 7, conclusion and future works.

Chapter One

Introduction

1.1 introduction

The mobile communication systems are often categorized as different generations depending on the services offered and the technologies used. The first generation comprises the analog Frequency Division Multiple Access (FDMA) systems such as AMPS (Advanced Mobile Phone Services). The second generation consists of the first digital mobile communication systems such as the Time Division Multiple Access (TDMA) based GSM (Global System for Mobile Communication), D-AMPS (Digital AMPS), Code Division Multiple Access (CDMA) based systems such as IS-95. These systems mainly offer speech communication, but also data communication limited to rather low transmission rates. The third generation started operations on 1st October 2002 in Japan. During the past few years, there has been an explosion in wireless technology. This growth has opened a new dimension to future wireless communications, whose ultimate goal is to provide universal personal and multimedia communication without regard to mobility or location with high data rates, [1-3]. To achieve such an objective, the next generation personal communication networks will need to support a wide range of services which will include high quality voice, data, still pictures and streaming video. Wireless broadband technologies promise to make all kinds of information available anywhere, anytime, at a low cost to a large portion of the population. More user devices than ever are going wireless for mobility and flexibility. WiMAX is a 4G technology that is fairly well accepted and will offer broadband data, voice and video services [4]. WiMAX (IEEE 802.16) with Wi-Fi (IEEE 802.11) will allow operators to deliver high quality voice, video and data services on a metropolitan scale, while Wi-Fi is able to provide high speed, localized, wireless Internet access, the emerging WiMAX standard is a wide area technology, supplying wireless coverage over an area of several kilometers. At the moment WiMAX technology will build upon Wi-Fi in a very complimentary way, hence WiMAX with

Wi-Fi can be called as a migration path to 4G. These future services are likely to include applications which require high transmission rates of several Mega bits per seconds (Mbps). In the current and future mobile communications systems, data transmission at high bit rates is essential for many services such as video, high quality audio and mobile integrated service digital network. When the data is transmitted at high bit rates, over mobile radio channels, the channel impulse response can extend over many symbol periods, which leads to ISI. OFDM is one of the promising candidate to mitigate the ISI. In an OFDM signal the bandwidth is divided into many narrow subchannels which are transmitted in parallel. Each subchannel is typically chosen narrow enough to eliminate the effect of delay spread [5]. OFDM has proven to be a modulation technique well suited for high data rates on time dispersive channels [6]. There are some specific requirements when designing wireless OFDM systems, for example, how to choose the bandwidth of the subchannels used for transmission, and how to achieve reliable synchronization. The synchronization is especially important in packet-based systems since it has to be achieved within a few symbols. In order to achieve good performance the receiver has to know the impact of the channel. The problem is how to extract this information in an efficient way. Conventionally, known symbols are multiplexed into the data sequence in order to estimate the channel. From these symbols, all channel attenuations are estimated.

1.2 The wireless channel

The radio propagation channel exhibits many forms of channel impairments, notably multipath delay spread, Doppler spread, fading ambient noise, Interference, distortion and noise can be differentiated into multiplicative and additive types as follows:

- Multiplicative interference and distortion are normally signal-dependant, and include fading, inter symbol interference. It causes attenuation, of the transmitted signal. The net result is a reduction in usable frequency spectrum. This form of disturbance cannot be suppressed by using filtering.
- Additive noise is not as severe as multiplicative noise, but it still reduces signal detectability. Out of band noise can be suppressed by filtering, but in band noise will penetrate through the filter [7].

Effective and efficient transmitters and receivers are needed to combat interference and distortion. Transmitter/receiver design requires a good knowledge of the channel characteristics. Thus, a good understanding of the propagation channel is essential for the design of effective transmitter, receivers, and communication protocols [7].

1.3 Wireless communication Standards

1.3.1 Second-generation standards:

Various digital cellular standards were developed in several regional standards bodies during the late 1980s and early 1990s. The first-generation standards had been developed some ten years earlier. The development of the new European digital cellular standard has been since 1985. GSM has since evolved into the leading global second-generation standard, GSM is an eight-slot, TDMA system with 200 kHz carrier spacing. In terms of service, GSM is mobile ISDN (Integrated Services Digital Network), with support for a wide variety of services. Intelligent Network (IN) support in the mobile environment has also been defined for GSM for example, the virtual home environment as well as many advanced data services, and General Packet Radio Services (GPRS), packet access can also be integrated into GSM. The TDMA specification, which was defined in the USA, in 1988, by the Telecommunications Industry Association (TIA), was developed with the aim of digitizing the AMPS. To maintain compatibility with AMPS, the TDMA specification stipulates 30 kHz carrier spacing in a three-slot TDMA solution. The narrowband Code-Division Multiple Access (CDMA) IS-95 specification stipulates 1.25 MHz carrier spacing for telephony services. TIA began defining this specification in 1991. Each of the second-generation standards essentially defines a mobile telephony system that is, a system that provides mobile end-users with circuit-switched telephony services. Apart from voice services, these systems support supplementary services and some low-bit-rate data services [8].

1.3.2 Third Generation Wireless Systems:

Third generation (3G) mobile systems such as the Universal Mobile Telecommunication System (UMTS) and CDMA2000. These systems are striving to provide higher data rates than 2G systems such as GSM, and IS-95. Second generation systems are mainly targeted at providing voice services, while 3G systems will shift to more data oriented services such as Internet access. Third generation systems use Wide-band Code Division Multiple Access (WCDMA) as the carrier modulation scheme. This modulation scheme has

- A high multipath tolerance.
- Flexible data rate.
- Allows a greater cellular spectral efficiency than 2G systems.

3G systems will provide a significantly higher data rate (64 kbps – 2 Mbps) than second-generation systems (9.6 – 14.4 kbps). The higher data rate of 3G systems will be able to support a wide range of applications including Internet access, voice communications and mobile videophones. In addition, a large number of new applications will emerge to utilize the permanent network connectivity, such as wireless appliances, notebooks with built in mobile phones, remote logging, wireless web cameras, car navigation systems, and so forth. In fact most of these applications will not be limited by the data rate provided by 3G systems, but by the cost of the service [9].

1.3.4 Forth Generation wireless Systems:

1G and 2G systems were voice communications, and digitized voice communications with some data communications, respectively, where a major difference was roaming between regions. 3G systems provide multimedia and wireless Internet at relatively high data rates, by utilizing packet switched services. However, significant paradigm shift should be taken into account for 4G systems, since wireless LAN, wireless MAN (WiMAX), wireless ad-hoc and sensor networks are becoming popular.

4G is the next generation of wireless networks that will replace 3G networks. 4G is intended to provide high speed, high capacity, low cost per bit, IP based services for video, data and

VoIP. 4G is all about an integrated, global network that is based on an open system approach. At the moment we have several technologies each capable of performing some of the functions like broadband data access in mobile or nomadic environments, supporting voice traffic using VoIP etc. The emerging IEEE 802.16 Broadband Wireless Access (BWA) technology WiMAX, allows interoperability and combines the benefits that other wireless networking technologies offer individually, and leads a path towards 4G and become the 4G wireless technology in the future. The 4G/WiMAX spectrum will be used for voice, video and data all BWA applications.

4G networks should encompass broadband wireless services, such as High Definition Television (HDTV) (4 - 20 Mbps) and computer network applications (1 - 100 Mbps). This will allow 4G networks to replace many of the functions of WLAN systems. However, to cover this application, cost of service must be reduced significantly from 3G networks. The spectral efficiency of 3G networks is too low to support high data rate services at low cost. As a consequence one of the main focuses of 4G systems will be to significantly improve the spectral efficiency. In addition to high data rates, future systems must support a higher Quality of Service (QoS) than current cellular systems, which are designed to achieve 90 - 95% coverage [9], i.e. network connection can be obtained over 90 - 95% of the area of the cell. This will become inadequate as more systems become dependent on wireless networking. As a result 4G systems are likely to require a QoS closer to 98 - 99.5%. In order to achieve this level of QoS it will require the communication system to be more flexible and adaptive. In many applications it is more important to maintain network connectivity than the actual data rate achieved. If the transmission path is very poor, e.g. in a building basement, then the data rate has to drop to maintain the link. Thus the data rate might vary from as low as 1 kbps in extreme conditions, to as high as 20 Mbps for a good transmission path. Alternatively, for applications requiring a fixed data rate, the QoS can be improved by allocating additional resources to users with a poor transmission path. A significant improvement in spectral efficiency will be required in order for 4G systems to provide true broadband access. This will only be achieved by significant advances in multiple aspects of cellular network systems, such as network structure, network management, smart antennas, RF modulation, user allocation, and general resource allocation [9].

4G Mobile and wireless communication systems should support the following functions:

- Higher transmission rate up to 100Mbps.
- Flexible to advanced Internet, QoS control.
- Enhanced security.
- Seamless operation across networks.
- Multiple broadband access options in combined with public and private networks. including wireless LAN, wireless home link and ad-hoc network.

4G systems are also characterized by the bandwidth to be allocated in 2-5 GHz band, propagation loss is higher resulting in smaller cell size. Also, due to higher Doppler shift, more complex and robust synchronization and channel estimation techniques are needed.

1.4 Mobile WiMAX Key Advantages

Mobile WiMAX, as a 4G technology, meets all the requirements for Personal Broadband access. It supports high data rates, high sector throughput, multiple handoff mechanisms, power-saving mechanisms for mobile devices, advanced QoS and low latency for improved support of real-time applications, advanced Broadband Wireless Access (AAA) functionality. Unlike the CDMA-based 3G systems, which have evolved from voice-centric systems, WiMAX is designed to meet the requirements necessary for the delivery of broadband data services as well as voice. UMTS, CDMA2000 and TD-SCDMA are all optimized for voice applications. The new technologies employed in mobile WiMAX result in lower equipment complexity and simpler mobility management due to the all-IP core network that provides with many other additional advantages over CDMA based 3G systems including [10];

- Tolerance to multipath and self-interference.
- Scalable Channel Bandwidth.
- Orthogonal Uplink Multiple Access.
- Support for Spectrally-Efficient TDD.

- Frequency-Selective Scheduling.
- Fractional Frequency Reuse Fine QoS.
- Advanced Antenna Technology.

OFDM technology provides operators with an efficient modulation technique to overcome Non Line Of Sight (NLOS) propagation. The WiMAX OFDM waveform offers the advantage of being able to operate with the large delay spread of the NLOS environment. By virtue of the OFDM symbol time and use of a cyclic prefix, the OFDM waveform eliminates the ISI problems and the complexities of adaptive equalization [11]. 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. OFDM is the basis for Orthogonal Frequency Division Multiple Access (OFDMA) with the advantages of OFDM carrying over to OFDMA [10],[12].

1.5 OFDM

The demand for multimedia wireless communications is growing today at an extremely rapid pace and this trend is expected to continue in the future. The common feature of many current wireless standards for high-rate multimedia transmission is the adoption of a multicarrier air interface based on OFDM. OFDM is an alternative wireless modulation technology to CDMA. OFDM has the potential to surpass the capacity of CDMA systems and provide the wireless access method for 4G systems. OFDM is a modulation scheme that allows digital data to be efficiently and reliably transmitted over a radio channel, even in multipath environments. OFDM transmits data by using a large number of narrow bandwidth carriers. These carriers are regularly spaced in frequency, forming a block of spectrum. The frequency spacing and time synchronization of the carriers is chosen in such a way that the carriers are orthogonal, meaning that they do not cause interference to each other. This is despite the carriers overlapping each other in the frequency domain.

The idea behind OFDM is to convert a frequency selective channel into a collection of frequency-flat subchannels with partially overlapping spectra. This goal is achieved by

splitting the input high-rate data stream into a number of sub streams that are transmitted in parallel over orthogonal subcarriers [13-15]. So in OFDM system a serial data stream is split into parallel streams that modulate a group of orthogonal subcarriers. Compared to single carrier modulation, OFDM symbols have a relatively long time duration, but a narrow bandwidth. Consequently, OFDM is robust to channel multipath dispersion and results in a decrease in the complexity of equalizers for high delay spread channels or high data rates [16-17]. However, the increased symbol duration makes an OFDM system more sensitive to the time variations of mobile radio channels. Furthermore, it provides larger flexibility by allowing independent selection of the modulation parameters over each subcarrier.

OFDM is a method to pack subcarriers, together into a symbol using as little bandwidth as possible see Fig.1.1 [18]. The subcarriers of a symbol are densely packed but at the center frequency of a subcarrier there is no overlap. In the time domain, orthogonality of the subcarriers translates into subcarriers all having an integer number of cycles in the OFDM symbol duration, and adjacent subcarriers having a number of cycles that differs by exactly one cycle as shown in Fig.1.2 [18]. The OFDM is done by passing the subcarriers through an Inverse Discrete Fourier Transform (IDFT) that takes the subcarriers from distinct points in the frequency domain into the time domain. The name OFDM is derived from the fact that the digital data is sent using many carriers, each of a different frequency (Frequency Division Multiplexing) and these carriers are orthogonal to each other, hence Orthogonal Frequency Division Multiplexing. The idea of using a Discrete Fourier Transform (DFT) for implementation of the generation and reception of OFDM signals, eliminating the requirement for banks of analog subcarrier oscillators [19]. This presented an opportunity for an easy implementation of OFDM, especially with the use of Fast Fourier Transforms (FFT), which are an efficient implementation of the DFT.

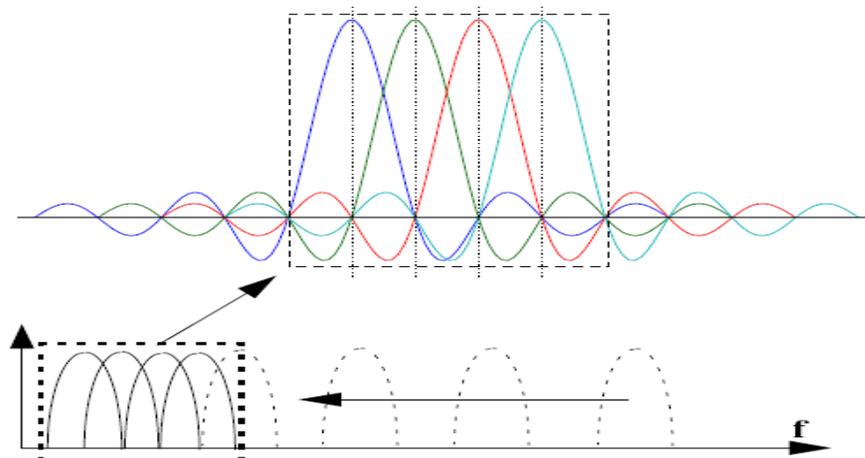


Fig.1.1 Orthogonal Frequency Division Multiplexing.

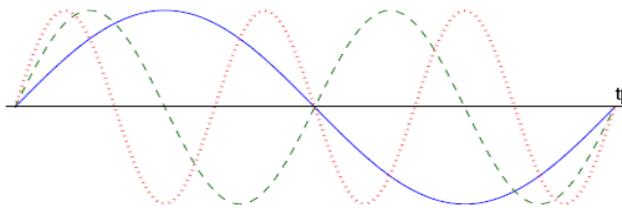


Fig1.2 Subcarriers within an OFDM Symbol.

OFDM has been adopted in some commercial systems such as Digital Audio Broadcasting (DAB) [20], terrestrial Digital Video Broadcasting (DVB-T) [21], and the IEEE 802.11a Wireless Local Area Network (WLAN) [22].

Orthogonal Frequency Division Multiple Access (OFDMA) technology, which results from a combination of OFDM with a FDMA protocol. This scheme was originally suggested by Sari and Karam for cable TV (CATV) networks [23], and later adopted in the uplink of the Interaction Channel for Digital Terrestrial Television (DVB-RCT) [24], it has become part of the emerging IEEE 802.16 standards for wireless metropolitan area networks (WMANs) [25], and is currently attracting vast researches attention from both academic and industrial point view, as a promising candidate for next generation broadband wireless networks.

1.6 OFDMA

OFDMA has recently attracted vast research attention from both academia and industry and has become part of new emerging standards for broadband wireless access. Even though the OFDMA concept is simple in its basic principle, the design of a practical OFDMA system is far from being a trivial task. Synchronization represents one of the most challenging issues and plays a major role in the physical layer design. In OFDMA systems, the available subcarriers are divided into several mutually exclusive subchannels that are assigned to distinct users for simultaneous transmission. The OFDMA system in downlink is essentially equivalent to an OFDM system. The only difference is that in OFDMA each transmitted block conveys simultaneous information for multiple subscribers while in OFDM it carries data for a single specific user. To fix the ideas, assume that the BS communicates with M users by exploiting N available subcarriers. The latter are evenly divided into R subchannels, each consisting of $P = N/R$ subcarriers. Without loss of generality, we consider the situation in which different subchannels are assigned to distinct users, even though in practice more subchannels may be allocated to the same user depending on its requested data rate [26].

The orthogonality among subcarriers guarantees intrinsic protection against Multiple Access Interference (MAI). Furthermore, OFDMA inherits from OFDM the ability to compensate channel distortions in the frequency domain without the need of computationally demanding time domain equalizers. Despite its appealing features, the design of an OFDMA system poses several technical challenges. One of the major problems with an OFDMA system is to synchronize the uplink transmission, because every user has to transmit its frame so that they avoid interfering the other users. For example as in Fig.1.3, if user 2 transmits too early it will disturb some of the user 1 transmission and if it transmits too late it will disturb user 4. On the downlink side this problem will not arise since the signal originates from a single source. So similarly to OFDM, OFDMA is extremely sensitive to timing errors and carrier frequency offsets between the incoming waveform and the local references used for signal demodulation. Inaccurate compensation of the frequency offset destroys orthogonality among subcarriers and produces Inter Channel Interference (ICI) as well as MAI. Timing errors result in Inter Block Interference (IBI) and must be counteracted to avoid severe error rate degradations.

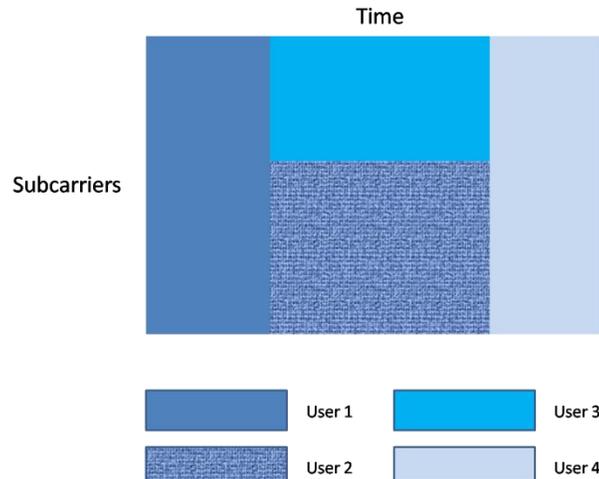


Fig 1.3 OFDMA system

In OFDMA, closely spaced and overlapped subcarriers are divided into groups and assigned to multiple users for simultaneous transmissions. Unlike traditional FDMA, where any overlapping of the frequency spectrum of different users introduces MAI, the orthogonality of subcarriers guarantees that there is no ICI, which prevents MAI among users in OFDMA systems. OFDMA inherits from OFDM the weakness of being sensitive to inaccurate frequency references. Carrier-Frequency Offset (CFO) between the transmitter and the receiver causes the loss of orthogonality among subcarriers and introduces ICI. In OFDMA, CFO will further cause MAI, which degrades the system performance [27]. CFO estimation is relatively simple in the broadcast link (downlink), where different user's signals are multiplexed by the same transmitter, and the orthogonality among all subcarriers is maintained. Each user can perform the frequency synchronization by estimating a single CFO between itself and the transmitter, and compensate accordingly. Many CFO-estimation algorithms proposed for OFDM are applicable to the OFDMA downlink. The real challenge exists in the uplink of OFDMA, where a number of users share the total number of subcarriers, and each user has its own CFO. CFO estimation, in this case, becomes a multiple-parameter estimation problem. CFO estimation in the OFDMA uplink is closely related to the subcarrier-assignment scheme adopted by the system [28].

1.7 IEEE 802.16e

The IEEE 802.16 standards are intended to offer wireless broadband technology for the long-range connection back to the service provider. They are also known as WiMAX standards and are supported by the WiMAX Forum. In January 2006, the WiMAX Forum announced the first products for IEEE802.16-2004-compliant certification [29]. The IEEE 802.16 standard ensures compatibility and interoperability between broadband wireless access components. IEEE 802.16 technology provides speeds comparable to wired systems, like cable and Digital Subscriber Line (DSL) links. End users can connect it to their internal wired Ethernet or wireless LANs. A summary of the IEEE 802.16 standards is shown in Table 1.1 [30]. These new technology of robust wireless communication has become available, and it is having a significant impact on how industrial operations are conducted.

Table 1.1 IEEE 802.16 standards

Standard	802.16e	802.16a/802.16d	802.16
Completed	Dec 2005	802.16a: Jan 2003 802.16d: Sep 2004	Dec 2001
Spectrum	2 to 6 GHz	2 to 11 GHz	10 to 66 GHz
Channel conditions	NLOS	NLOS	LOS
Maximum data rate	75 Mb/s (20 MHz channel)	75 Mb/s (20 MHz channel)	134Mb/s (28 MHz channel)
Modulation	Same as 802.16d	OFDM, OFDMA,QPSK, 16-QAM,64- QAM,BPSK	QPSK, 16-QAM, 64-QAM
Mobility	Mobile and roaming	Fixed and portable	Fixed
Channel bandwidths	Same as 802.16d	Scalable 1.5 to 20 MHz	20,25,and28 MHz
Typical cell radius	1 to 3 miles	3 to 5 miles	1 to 3 miles

IEEE 802.16e uses SOFDMA (Scalable Orthogonal Frequency Division Multiple Access) as transmission technique. SOFDMA is an OFDMA version where the bandwidth is scalable; in 802.16e it is scalable between 1.25 to 20 MHz. The scalability is achieved by changing the FFT size, while keeping fixed subcarrier spacing [31]. Mobile WiMAX is intended for the 2.3 GHz, 2.5 GHz and 3.5 GHz spectra [15]. The system is defined so that the user can travel at speeds between 0-120 km/h. The theoretical upper limit for the bit rate in WiMAX, given a

bandwidth of 10 MHz, is 31 Mbps in downlink and 23 Mbps in uplink [31]. The base stations have a typical coverage up to an 8 km radius in a NLOS environment [32]. One of the more interesting features of 802.16e is that it supports MIMO (Multiple Input Multiple Output) devices. The market that 802.16e is focused on is typical city area network. But in the future wireless devices such as laptops and cell phones will probably be able to access different types of wireless networks depending on availability.

1.8 Comparison between Wi-Fi and WiMAX:

Table 1.2 provides a strong comparison between Wireless-Fidelity (Wi-Fi) and WiMAX technologies in terms of various aspects [4].

Table 1.2 Comparison between Wi-Fi and WiMAX

	Wi-Fi (802.11)	WiMAX (802.16)
Modulation	OFDM However, the two implementations are not identical.	OFDM-WiMAX radios require higher power because they must transmit over much longer distances than Wi-Fi radios
Spectrum	Operated in the license free bands (2.4GHz, 5.1 GHz).	WiMAX operates in licensed as well as unlicensed swaths of Spectrum
Frequency	802.11 standard dictates one channel width and one frequency (either 2.4 or 5 GHz),	802.16a/REVd can work in a number of frequency bands, and the available frequency can be sliced into a variety of channel widths
Range	Wi-Fi in the LAN - Long-range Wi-Fi has complications.	WiMAX in the MAN / WAN
Channel bandwidth	Fixed 20 MHz bandwidth with 52 sub carriers	Variable bandwidths from 1 to 28 MHz with 256 sub carriers (192 data sub carriers)
Guard interval	Fixed at $\frac{1}{4}$ symbol	Time Variable length guard interval at the beginning of the packet to compensate for multi-path interference: ranges from $\frac{1}{32}$ to $\frac{1}{4}$ symbol time
Error vector magnitude (EVM)	-25dB which is required to achieve 10% packet error rate.	-31dB which is based on a 1% packet error rate

Chapter Two

Fading In Wireless Environment

Radio waves propagate from a transmitting antenna, and travel through free space undergoing absorption, reflection, refraction, diffraction, and scattering. They are greatly affected by the ground terrain, the atmosphere, and the objects in their path, like buildings, bridges, hills, trees. These multiple physical phenomena are responsible for most of the characteristic features of the received signal.

2.1 Mobile Radio Propagation :Large Scale and Small Scale Fading

Fig.2.1 represents an overview of fading channel manifestations. It starts with two types of fading effects that characterize mobile communications [33-34].

- large-scale fading.
- small-scale fading.

Large-scale fading represents the average signal power attenuation or path loss due to motion over large areas. The large-scale fading manifestation is shown in blocks 1, 2, and 3, [33]. This phenomenon is affected by prominent terrain contours between the transmitter and receiver like hills, forests, billboards, clumps of buildings. The receiver is often represented as being shadowed by such prominences. The statistics of large-scale fading provide a way of computing an estimate of path loss as a function of distance. This is described in terms of a mean-path loss and a log-normally distributed variation about the mean.

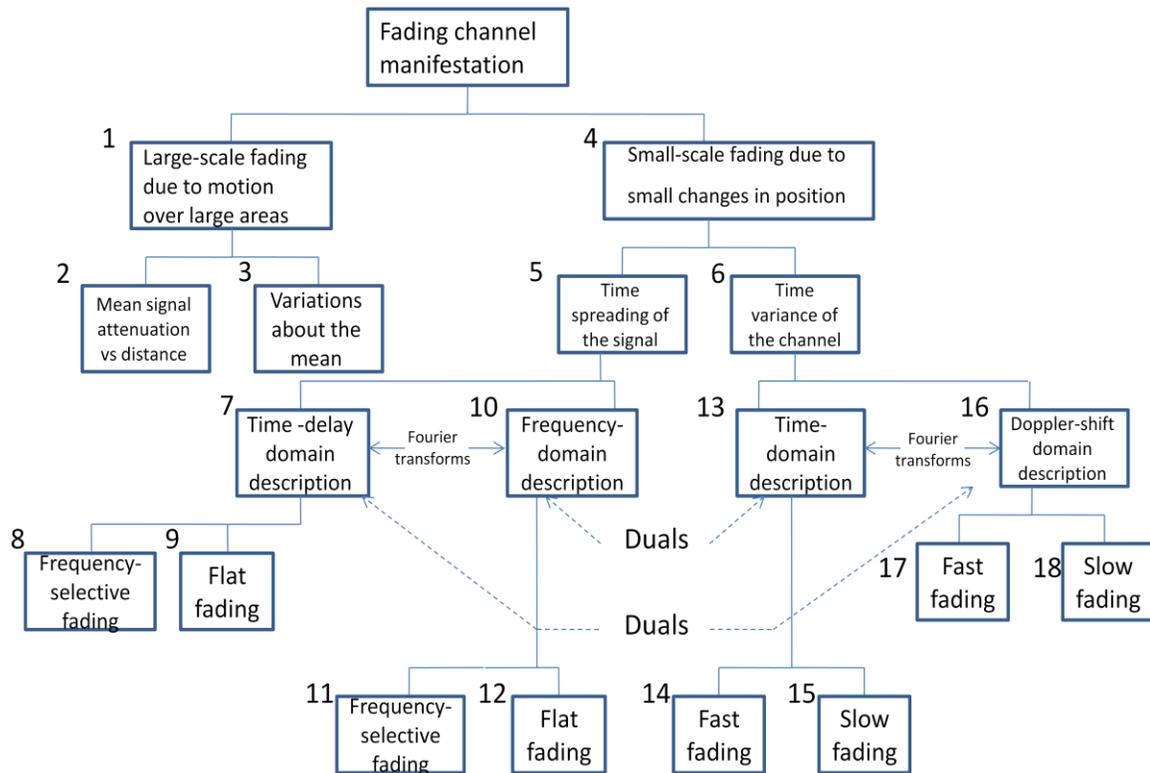


Fig.2.1 fading channel manifestations

Shadowing of the signal can occur whenever there is an obstruction between the transmitter and receiver. It is generally caused by buildings and hills, and is the most important environmental attenuation factor. Shadowing is most severe in heavily built up areas, due to the shadowing from buildings. However, hills can cause a large problem due to the large shadow they produce. Radio signals diffract off the boundaries of obstructions, thus preventing total shadowing of the signals behind hills and buildings. However, the amount of diffraction is dependent on the radio frequency used, where low frequencies diffracting more than high frequency signals. Thus high frequency signals, especially, Ultra High Frequencies (UHF), and microwave signals require line of sight for adequate signal strength. To overcome the problem of shadowing, transmitters are usually elevated as high as possible to minimize the number of obstructions.

Small-scale fading refers to the dramatic changes in signal amplitude and phase that can be experienced as a result of small changes in the spatial separation between a receiver and transmitter. The transmitted signal arrives at the receiver via several paths with different time delays creating a multipath situation as shown in Fig.2.2. At the receiver, these multipath waves with randomly distributed amplitudes and phases combine to give a resultant signal that is small-scale fading which is classified as flat or frequency selective or slow or fast fading. As indicated in Fig.2.1 small-scale fading manifests itself into two mechanisms, time spreading of the signal and time-variant behavior of the channel.

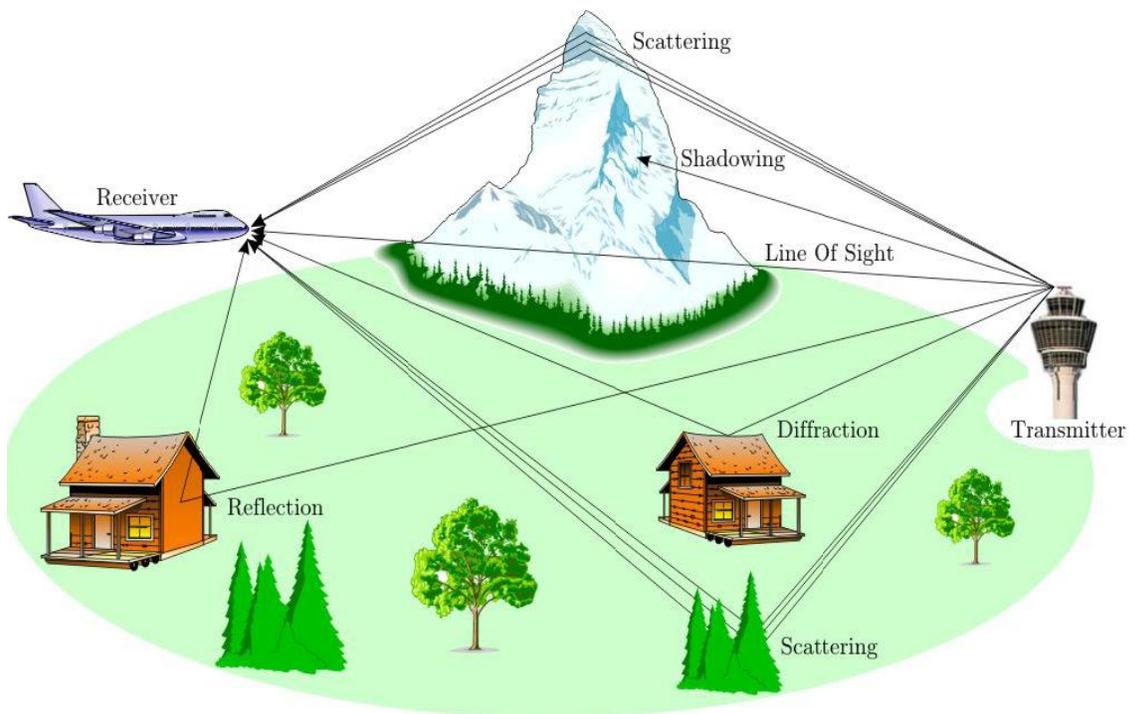


Fig.2.2 Multipath fading

For mobile radio applications, the channel is time-variant because motion between the transmitter and receiver results in propagation path changes. The rate of change of these propagation conditions accounts for the fading rapidity. So when there is a relative motion between the transmitter and the receiver, Doppler spread is introduced in the received signal spectrum, causing frequency dispersion such that;

- If the Doppler spread is significant relative to the bandwidth of the transmitted signal, the received signal is said to undergo fast fading. This form of fading typically occurs for very low data rates.
- If the Doppler spread of the channel is much less than the bandwidth of the baseband signal then signal is said to undergo slow fading.

Small-scale fading is also called Rayleigh fading because if the multiple reflective paths are large in number and there is no line-of-sight signal component, the envelope of the received signal is statistically described by a Rayleigh Probability Density Function (PDF). When there is a dominant non fading signal component present, such as a line-of-sight propagation path, the small scale fading envelope is described by a Rician PDF [34].

There are three basic mechanisms that impact signal propagation in a mobile communication system. They are:

- Reflection: Reflection occurs when a propagating electromagnetic wave impinges on a smooth surface with very large dimensions compared to the RF signal wavelength (λ)
- Diffraction occurs when the radio path between the transmitter and receiver is obstructed by a dense body with large dimensions compared to the propagation wave length (λ), causing secondary waves to be formed behind the obstructing body. Diffraction is a phenomenon that accounts for RF energy traveling from transmitter to receiver without a line-of-sight path between the two. It is often termed shadowing because the diffracted field can reach the receiver even when shadowed by an impenetrable obstruction.
- Scattering occurs when a radio wave impinges on either a large rough surface or any surface whose dimensions are on the order of λ or less, causing the reflected energy to spread out (scatter) in all directions. In an urban environment, typical signal obstructions that yield scattering are lampposts, street signs, and foliage.

2.2 Flat and Frequency selective Fading channel

A channel is referred to as frequency-selective if coherence bandwidth f_0 less than the symbol rate ($1/T_s$) where $f_0=1/\tau_{max}$ and τ_{max} is the maximum excess delay, W nominally taken to be equal to the signal bandwidth W . In practice, W may differ from $1/T_s$ due to system filtering or data modulation type (quaternary phase shift keying, QPSK, minimum shift keying, MSK, etc.) [33],[34].

Frequency-selective fading distortion occurs whenever a signal's spectral components are not all affected equally by the channel. Some of the signal's spectral components, falling outside the coherence bandwidth, will be affected differently compared to those components contained within the coherence bandwidth. This occurs whenever $f_0 < W$ and, [33].

Frequency-nonselective, or flat fading degradation, occurs whenever $f_0 > W$. Hence, all of the signal's spectral components will be affected by the channel in a similar manner. Flat-fading does not introduce channel-induced ISI distortion, but performance degradation can still be expected due to loss in SNR whenever the signal is fading. In order to avoid channel induced ISI distortion, the channel is required to exhibit flat fading by ensuring that

$$f_0 > W \cong \frac{1}{T_s} \quad (2.1)$$

For the flat-fading case, where $f_0 > W$ (or $\tau_{max} < T_s$). So if the mobile radio channel has a constant gain and a linear phase response over a bandwidth larger than the bandwidth of the transmitted signal under these conditions, the received signal has amplitude fluctuations due to the variations in the channel gain over time caused by multipath [33].

2.3 Fast and Slow Fading Channel

A channel is referred to as fast fading if the symbol rate, $1/T_s$ is less than the fading rate $1/T_0$ which is approximately equal to maximum Doppler shift F_d , where $F_d = \frac{v}{\lambda}$, and v is relative

velocity, λ is the signal wavelength, and T_0 denudated as coherence time. So fast fading is characterized by Eq.2.2 and Eq.2.3

$$W < F_d \quad (2.2)$$

Or

$$T_s > T_0. \quad (2.3)$$

Conversely, a channel is referred to as slow fading if the signaling rate is greater than the fading rate. Thus, in order to avoid signal distortion caused by fast fading, the channel must be made to exhibit slow fading by ensuring that the signaling rate must exceed the channel fading rate. That is,

$$W > F_d \quad (2.4)$$

Or

$$T_s < T_0. \quad (2.5)$$

In Eq.2.1, it was shown that due to signal dispersion, the coherence bandwidth, f_0 , sets an upper limit on the signaling rate which can be used without suffering frequency-selective distortion. Similarly, Eq.2.4 shows that due to Doppler spreading, the channel fading rate, f_D , sets a lower limit on the signaling rate that can be used without suffering fast fading distortion [33],[34].

A better way to state the requirement for mitigating the effects of fast fading would be that we desire $W \gg F_d$ (or $T_s \ll T_0$). If this condition is not satisfied, the random frequency modulation due to varying Doppler shifts will limit the system performance significantly, [33].

2.4 Statistical modeling of Fading:

Many models for the probability distribution function of the signal amplitude exposed to mobile fading have been given. Out of these models Rayleigh fading, Rician Fading Nakagami, fading log normal model, and Suzuki model.

2.4.1 Rayleigh Fading:

The mobile antenna, instead of receiving the signal over one line-of-sight path, receives a number of reflected and scattered waves, as shown in Fig.2.2. Because of the varying path lengths, the phases are random, and consequently, the instantaneous received power becomes a random variable. In the case of an un-modulated carrier, the transmitted signal at frequency f_c reaches the receiver via a number of paths, the i^{th} path having an amplitude a_i , and a phase ϕ_i . If we assume that there is no direct path or Line-Of Sight (LOS) component, the received signal $s(t)$ can be expressed as:

$$s(t) = \sum_{i=1}^L a_i \cos(2\pi f_c t + \phi_i) \quad (2.6)$$

where L is the number of paths. The phase ϕ_i depends on the varying path lengths changing by 2π , [35].

Eq.2.6 must be modified to include the effects of motion induced frequency and phase shifts. Let the i^{th} reflected wave with amplitude a_i and phase ϕ_i arrive at the receiver from an angle α_i relative to the direction of motion of the antenna. The Doppler shift of this wave is given by,

$$f_{di} = \frac{f_c v}{c} \cos(\alpha_i) \quad (2.7)$$

where v is the velocity of the mobile, c is the speed of light (3×10^8 m/s), and the α_i s are uniformly distributed over $[0, 2\pi]$. The received signal $s(t)$ can now be written as;

$$s(t) = \sum_{i=1}^L a_i \cos(2\pi(f_c + f_{di})t + \phi_i) \quad (2.8)$$

Expressing the signal in inphase and quadrature form Eq.2.8 can be written as;

$$s(t) = I(t) \cos(2\pi f_c t) - Q(t) \sin(2\pi f_c t) \quad (2.9)$$

Where

$$I(t) = \sum_{i=1}^L a_i \cos(2\pi f_{di} t + \phi_i) \quad (2.10)$$

And

$$Q(t) = \sum_{i=1}^L a_i \sin(2\pi f_{di} t + \phi_i) \quad (2.11)$$

The envelope r is given by;

$$r = \sqrt{I^2(t) + Q^2(t)} \quad (2.12)$$

The PDF of the received signal envelope, $f(r)$, can be shown to be Rayleigh [7], given by;

$$f(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) \quad r > 0 \quad (2.13)$$

Where r is the envelope amplitude of the received signal, and $2\sigma^2$ is the mean power of the multipath signal.

2.4.2 Rician Fading:

The model behind Rician fading is similar to that for Rayleigh fading, except that in Rician fading a strong dominant component is present [7],[34],[35]. This dominant component can, for instance, be the LOS wave. Refined Rician models also consider;

- The dominant wave can be a phasor sum of two or more dominant signals e.g. the LOS, plus a ground reflection.
- The dominant wave can also be subject to shadow attenuation. This is a popular assumption in the modeling of satellite channels.

Besides the dominant component, the mobile antenna receives a large number of reflected and scattered waves.

The Rician factor K is defined as the ratio of signal power in dominant component to the (local-mean) scattered power

$$K = \frac{\text{Direct Power}}{\text{Scattered Power}}$$

In the presence of such a path, the received signal can be written as; [7].

$$s(t) = \sum_{i=1}^L a_i \cos(2\pi(f_c + f_{di})t + \phi_i) + K \cos(2\pi(f_c + f_d)t) \quad (2.14)$$

Where f_d is the Doppler shift along the LOS path.

The envelope in this case has a Rician density function given by Eq.2.15 [7].

$$f(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + 2K^2}{2\sigma^2}\right) J_0\left(\frac{2rK}{\sigma^2}\right) \quad r > 0 \quad (2.15)$$

Where J_0 is the zero-order Bessel function of the first kind.

2.4.3 Nakagami Fading:

Nakagami fading occurs for instance for multipath scattering with relatively large delay-time spread, with different clusters of reflected waves. Within any one cluster, the phases of individual reflected waves are random, but the delay times are approximately equal for all waves. As a result, the envelope of each cumulated cluster signal is Rayleigh distributed. The average time delay is assumed to differ significantly between clusters. If the delay times also significantly exceed the bit time of a digital link, the different clusters produce serious ISI, so the multipath self-interference then approximates the case of co-channel interference by

multiple incoherent Rayleigh-fading signals. Following are some important facts related to Nakagami fading [35].

- If the envelope is Nakagami distributed, the corresponding instantaneous power is gamma distributed.
- The parameter m is called the 'shape factor' of the Nakagami or the gamma distribution.
- In the special case $m = 1$, Rayleigh fading is recovered, with an exponentially distributed instantaneous power
- For $m > 1$, the fluctuations of the signal strength reduce compared to Rayleigh fading.

The fading model for the received signal envelope, proposed by Nakagami, has the PDF given by Eq.2.17 [35].

$$f(r) = \frac{2m^m r^{2m-1}}{\Gamma(m)\Omega^m} \exp\left(-\frac{mr^2}{\Omega}\right) \quad r > 0 \quad (2.17)$$

where $\Gamma(m)$ is the Gamma function, m is the shape factor (with the constraint that $m \geq 1/2$) given by Eq.2.18

$$m = \frac{E^2\{r^2\}}{E\{[r^2 - E(r^2)]\}^2} \quad (2.18)$$

Where $E[X]$ is the expected value of X , the parameter Ω control the spread of the distribution and is giving as $E[r^2]$.

2.4.4 Lognormal Distribution:

The fading over large distances causes random fluctuations in the mean signal power. Evidence suggests that these fluctuations are lognormal distributed. A heuristic explanation for encountering this distribution is as follows: The transmitted signal undergoes multiple

reflections at the various objects in its path, before reaching the receiver. Then it splits up into a number of paths, which finally combine at the receiver. The expression for the transmitted signal is the same as given in Eq.2.8, except that the path amplitudes a_i are themselves the products of the amplitudes due to the multiple reflections [36], as in Eq.2.19, [35].

$$a_i = \prod_{j=1}^{M_i} a_{ji} \quad (2.19)$$

where M_i is the number of multiple reflections per path. Multiplication of the signal amplitude leads to a lognormal distribution [36], in the same manner that an addition results in a normal distribution by virtue of the central limit theorem [37]. A study of mobile radio propagation modeling reveals that there is no direct reference to the global statistics of path amplitudes [35].

However, the fact that the mean of the envelope is lognormal seems to be well established in the literature. The lognormal PDF is given by Eq.2.20.

$$f(r) = \frac{1}{r\sigma\sqrt{2\pi}} \exp\left\{-\frac{(\ln r - \mu)^2}{2\sigma^2}\right\} \quad r > 0 \quad (2.20)$$

where μ is the mean of $\log(r)$, and σ^2 is the variance of $\log(r)$. With this distribution, $\log r$ has a normal distribution.

2.4.5 Suzuki Distribution:

Another approach used to describe the statistical fluctuations in the received signal combines the Rayleigh and lognormal in a single model. Suzuki [38] suggested that the envelope statistics of the received signal envelope could be represented by a mixture of Rayleigh and lognormal distributions in the form of a Rayleigh distribution with a log normally varying mean [38]. He suggested the formulation in Eq.2.21,

$$f(r) = \int_0^{\infty} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) \frac{1}{\sigma\zeta\sqrt{2\pi}} \exp\left(-\frac{(\ln\sigma - \mu)^2}{2\zeta^2}\right) d\sigma \quad r > 0 \quad (2.21)$$

where σ is the mode or the most probable value of the Rayleigh distribution, ζ is the shape parameter of the lognormal distribution, [35].

Chapter Three

OFDM principles

3.1 Introduction

In a conventional serial data system, the symbols are transmitted sequentially, with the frequency spectrum of each data symbol allowed to occupy the entire available bandwidth. A parallel data transmission system offers possibilities for alleviating many of the problems encountered with serial systems. A parallel system is one in which several sequential streams of data are transmitted simultaneously, so that at any instant many data elements are being transmitted. In such a system, the spectrum of an individual data element normally occupies only a small part of the available bandwidth. In the OFDM scheme, shown in Fig.3.1 the serial data stream of a traffic channel is passed through a Serial-to-Parallel (S/P) convertor, which splits the data into a number of parallel channels. The data in each channel is applied to a modulator, such that for N channels there are N modulators whose carrier frequencies are f_0, f_1, \dots, f_{N-1} . The difference between adjacent channels is Δf and the overall bandwidth W , of the N modulated carriers is $N\Delta f$. In the more conventional serial transmission approach, the traffic data is applied directly to the modulator, transmitting at a carrier frequency positioned at the center of the transmission band and the modulated signal occupies the entire bandwidth W , [39].

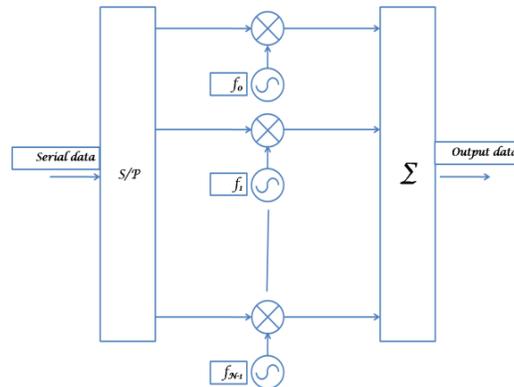


Fig.3.1 OFDM Scheme

OFDM is based on multicarrier communication techniques. The idea of multicarrier communications is to divide the total signal bandwidth into number of subcarriers and information is transmitted on each of the subcarriers. Unlike the conventional multicarrier communication scheme in which spectrum of each subcarrier is non-overlapping and band pass filtering is used to extract the frequency of interest, in OFDM the frequency spacing between subcarriers is selected such that the subcarriers are mathematically orthogonal to each other [40]. The spectra of subcarriers overlap each other but individual subcarrier can be extracted by baseband processing. This overlapping property makes OFDM more spectral efficient than the conventional multicarrier communication scheme. Fig.3.2 [11], shows the serial transmission of symbols $S_0; S_1; \dots; S_{N-1}$, during the N -symbol period of the conventional serial system, each OFDM modulator carries only one symbol, and the error burst causes severe signal degradation of the duration of k -serial symbols. In parallel stream the error burst is only a small fraction of the symbol period than each of the OFDM symbols and it is slightly affected by the fade. Accordingly they can be correctly demodulated. Thus while the serial system exhibits an error burst, no errors or few errors may occur using the OFDM approach. The principal advantage of OFDM is less sensitive to channel induced dispersion, this is because the symbol period has been increased, the channel's delay spread becomes a significantly shorter fraction of a symbol period than in the serial system.

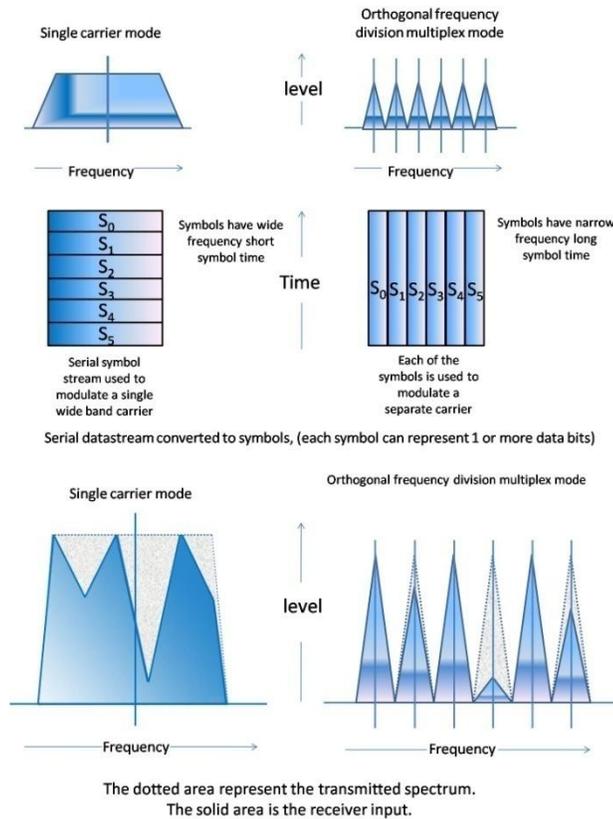


Fig.3.2 Effect of fade on serial and parallel transmission

3.2 Generation of OFDM Symbols

A baseband OFDM symbol can be generated in the digital domain before modulating a carrier for transmission. To generate a baseband OFDM symbol, a serial digitized data stream is first modulated using common modulation schemes such as the Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM). Then these data symbols are converted to parallel streams before modulating subcarriers. Subcarriers are sampled with sampling rate NT_s , where N is the number of subcarriers and T_s is the OFDM symbol duration. Finally, samples on each subcarrier are summed together to form an OFDM sample [41], as illustrated in Fig.3.3. This is equivalent to the N -point Inverse Discrete Fourier Transform (IDFT). It is well known that IDFT can be implemented efficiently using Inverse Fast Fourier Transform (IFFT) [42]. Therefore, in practice, the IFFT is performed on the data sequence at an OFDM transmitter for baseband modulation and the FFT is performed at an OFDM receiver for baseband

demodulation. Size of FFT and IFFT is N , which is equal to the number of subchannels available for transmission. The subchannel bandwidth is given by ;

$$f_{sc} = \frac{1}{T_s} = \frac{f}{N} \tag{3.1}$$

where f is the sample rate and T_s is the symbol time. Finally, a baseband OFDM symbol is modulated by a carrier to become a band pass signal and transmitted to the receiver.

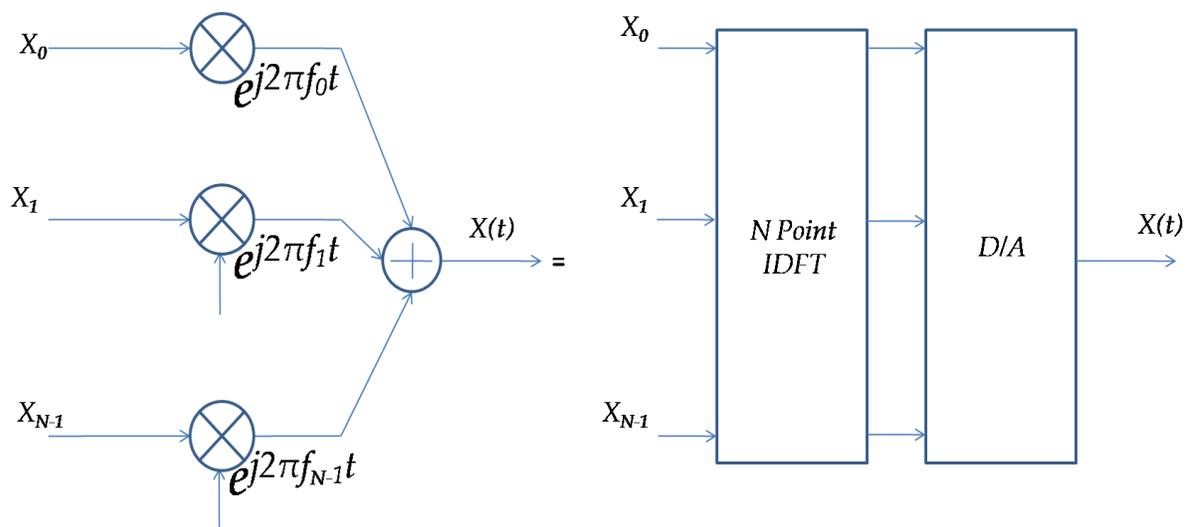


Fig.3.3 DFT implementation of transmitted wave form

3.3 Calculation of OFDM Parameters

For a given bit rate R and the delay spread of a multipath channel τ , the parameters of an OFDM system can be determined as follows [41]:

- The guard time G should be at least twice the delay spread, i.e.

$$G \geq 2\tau .$$

- To minimize the SNR loss due to the guard time, the symbol duration should be much larger than the guard time. However, symbols with long duration are susceptible to

Doppler spread, phase noise, and frequency offset. As a rule of thumb, the OFDM symbol duration T_s should be at least five times the guard time, i.e.

$$T_s \geq 5G$$

- The frequency spacing between two adjacent subcarriers is

$$\Delta f = 1/T_s$$

- For a given data rate R , the number of information bits per OFDM symbol B_{info} is

$$B_{info} = RT_s$$

- For a given B_{info} and the number of bits per symbol per subcarrier R_{sub} , the number of subcarriers N is

$$N = B_{info} / R_{sub}$$

where $R_{sub} = 2$ bits/symbol/subcarrier for QPSK

$R_{sub} = 4$ bits/symbol/subcarrier for 16-QAM.

The OFDM signal bandwidth is defined as

$$BW = N\Delta f$$

Two observations are made from the above calculations:

- Increasing the symbol duration decreases the frequency spacing between subcarriers. Thus, for a given signal bandwidth, more subcarriers can be accommodated. On the other hand, for a given number of subcarriers, increasing the symbol duration decreases the signal bandwidth.

- Increasing the number of subcarriers increases the number of samples per OFDM symbol. However, it does not necessarily imply that the symbol duration increases. If the OFDM symbol duration remains the same, the duration between two samples decreases as a result. This implies the increase in the OFDM signal bandwidth. On the other hand, if the OFDM signal bandwidth is fixed, then increasing the number of subcarriers decreases the frequency spacing between two subcarriers, which in turn increases the symbol duration. The duration between two samples remain the same in this case [41].

3.4 Inter-Symbol and Inter-Carrier Interference:

In a multipath environment, a transmitted symbol takes different times to reach the receiver through different propagation paths, in this case the channel introduces time dispersion in which the duration of the received symbol is stretched, this will cause the current received symbol to overlap previous received symbols and results in ISI, [34]. In OFDM, ISI usually refers to interference of an OFDM symbol by previous OFDM symbols.

In OFDM, the spectra of subcarriers overlap but remain orthogonal to each other. This means that at the maximum of each subcarrier spectrum, all the spectra of other subcarriers are zero [41]. The receiver samples data symbols on individual subcarriers at the maximum points and demodulates them free from any interference from the other subcarriers. Interference caused by data symbols on adjacent subcarriers is referred to ICI. The orthogonality of subcarriers can be viewed in either the time domain or in frequency domain. From the time domain perspective, each subcarrier is a sinusoid with an integer number of cycles within one FFT interval, however from the frequency domain perspective, this corresponds to each subcarrier having the maximum value at its own center frequency and zero at the center frequency of each of the other subcarriers. The orthogonality of a subcarrier with respect to other subcarriers is lost if the subcarrier has nonzero spectral value at other subcarrier frequencies. From the time domain perspective, the corresponding sinusoid no longer has an integer number of cycles within the FFT interval.

ICI occurs when the multipath channel varies over one OFDM symbol time. When this happens, the Doppler shifts on each multipath component causes a frequency offset on the subcarriers, resulting in the loss of orthogonality among them. ICI also occurs when there is a frequency offset due transmitter/ receiver oscillator mismatch, phase noise, and/or the non-linear power amplifier effect. The oscillator mismatch or the phase noise cause the received signal to be sampled at incorrect positions, and thereby taking the effect of all the subcarriers that is, orthogonality loss, [43],[44]. When ICI left without compensation, this effect reduces the performance of channel estimation methods, especially those based on fixed channel statistics [45]. ICI needs to be compensated either due to the frequency offset or the fast-varying nature of the Channel Impulse Response (CIR) taps, these two effect will be presented independently.

3.4.1 ICI Due to Frequency Offset:

ICI due to frequency offset mostly occurs due to the loss of synchronization of the subcarriers or the phase noise of the oscillators. In WLAN and WiMAX standards, in the preamble, two short duration OFDM symbols are provided for the synchronization purposes. These short symbols can also be used for the frequency offset estimation. The compensation of ICI due the frequency offset is relatively less challenging compared to the compensation of the ICI due to fast channel variation since the value of the frequency offset parameter is constant over all the subcarriers.

3.4.2 ICI Due to Fast Fading Channel:

When the CIR taps vary over the duration of OFDM symbols, for an accurate channel estimation, the CIR tap values corresponding to each sampling instance need to be obtained so that the corresponding CFR is estimated

The characteristics of ICI are similar to Gaussian noise, hence it leads to degradation of the SNR. The amount of degradation is proportional to the fractional frequency offset which is equal to the ratio of frequency offset to the carrier spacing. Frequency offset can be estimated

by different methods e.g.(using pilot symbols, the statistical redundancy in the received signal, or transmitted training sequences).

3.5 Current ICI reduction methods

Currently a few approaches for reducing ICI have been developed. Three solutions to combat ICI have been presented in [46].

- self-cancellation scheme in which redundant data is transmitted onto adjacent sub-carriers such that the ICI between adjacent sub-carriers cancels out at the receiver.
- Maximum likelihood (ML) estimation
- Extended Kalman filter (EKF) method, statistically estimate the frequency offset and correct the offset using the estimated value at the receiver.

A pulse shape is suggested to decrease the ICI in OFDM due to frequency offset ,such as rectangular, raised cosine and improved sinc power pulse were used in [47],[48],[49].

3.6 Guard Time Insertion

One of the most important reasons to employ OFDM modulation is the efficient way it deals with multipath delay spread. By dividing the input data stream in N subcarriers, the symbol duration is made N time longer, which reduces the relative multipath delay spread, relative to the symbol time, by the same factor.

To eliminate ISI almost completely, a guard time is introduced for each OFDM symbol. The guard interval is chosen larger than the expected delay spread, such that multipath components from one symbol cannot interfere with the next symbol. The method is best explained with reference to Fig 3.4. Every block of N samples as obtained by IFFT is quasi-periodically extended by a length N_g simply repeating N_g samples of the useful information block. The total sequence length becomes $N + N_g$ samples, corresponding to a duration of $T_s + T_g$. [39]. Trailing and leading samples of this extended block are corrupted by the channel transient

response, hence the receiver should demodulate only the central N number of samples, essentially unaffected by the channel's transient response. A guard time is inserted at the beginning of each OFDM symbol before transmission and removed at the receiver before the FFT operation. In order to preserve orthogonality among subcarriers. Guard time insertion can be introduced in many ways, but the most effective way of inserting guard period is to extract a portion of an OFDM symbol at the end and append it to the beginning of the OFDM symbol as shown in Fig.3.4.

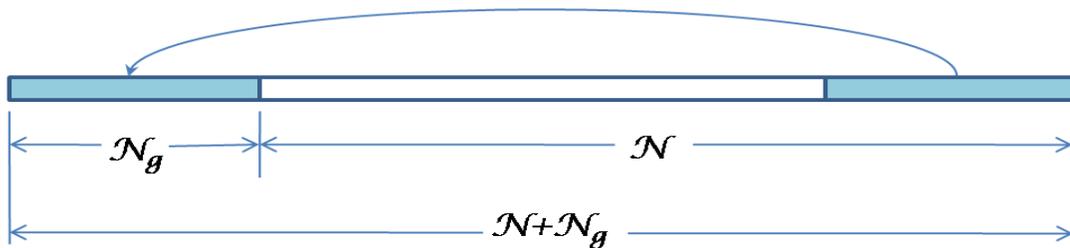


Fig.3.4 Guard interval by cyclic extension

The cyclic extension actually wastes channel capacity as well as transmitted power; however, if the useful information blocks are long, the extension length can be kept low relative to the useful information block length. If the guard time is chosen such that its duration is longer than the delay spread, the ISI can be completely eliminated. Fig 3.5 and Fig 3.6 illustrates the concept of guard time insertion in an OFDM system.

Fig 3.7 and Fig3.8 demonstrates the idea of eliminating ISI from OFDM symbols. In Fig.3.7 an OFDM symbol received is interfered from the previous OFDM symbol. On the other hand, Fig.3.8 shows that the OFDM symbol received is no longer interfered from the previous OFDM symbol. However, the received symbol is still interfered by its replicas and we refer to this type of interference as self-interference. In order to preserve orthogonality among subcarriers, the guard time is inserted by cyclically extending an OFDM symbol [41].

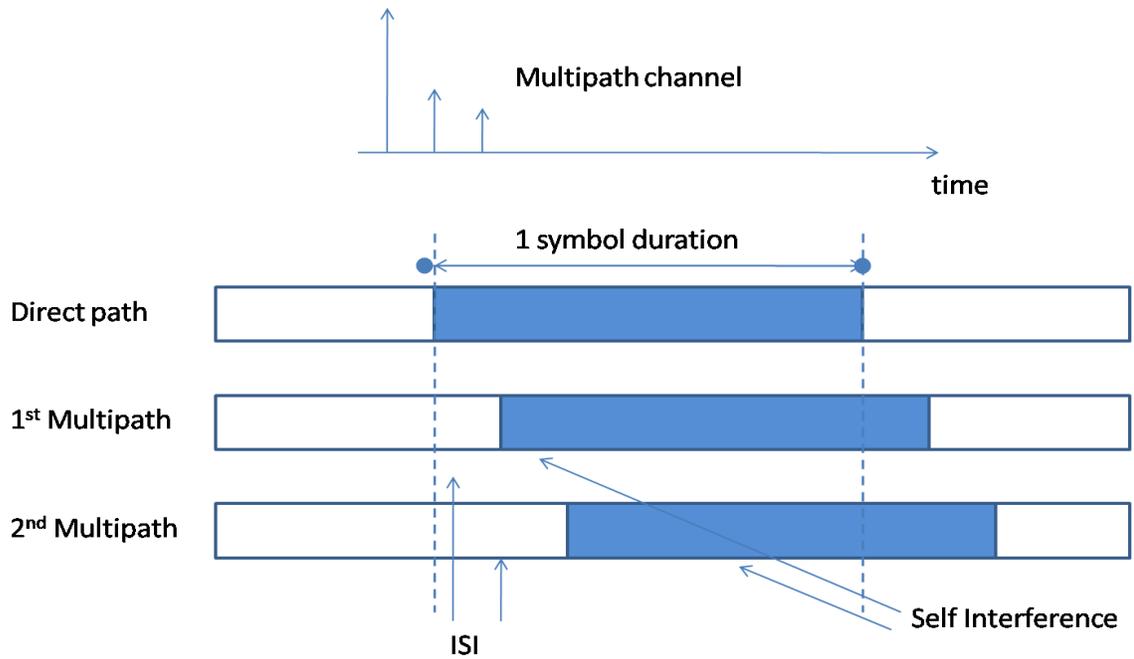


Fig 3.5 Received OFDM Symbol Components after passing through a multipath channel without guard interval.

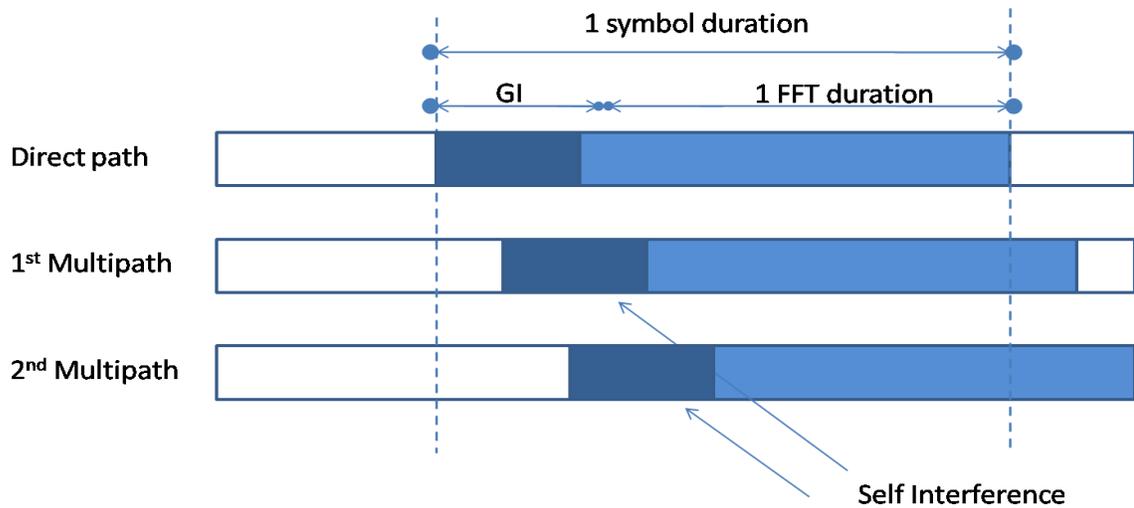


Fig 3.6 Received OFDM Symbol Components after passing through a multipath with guard interval

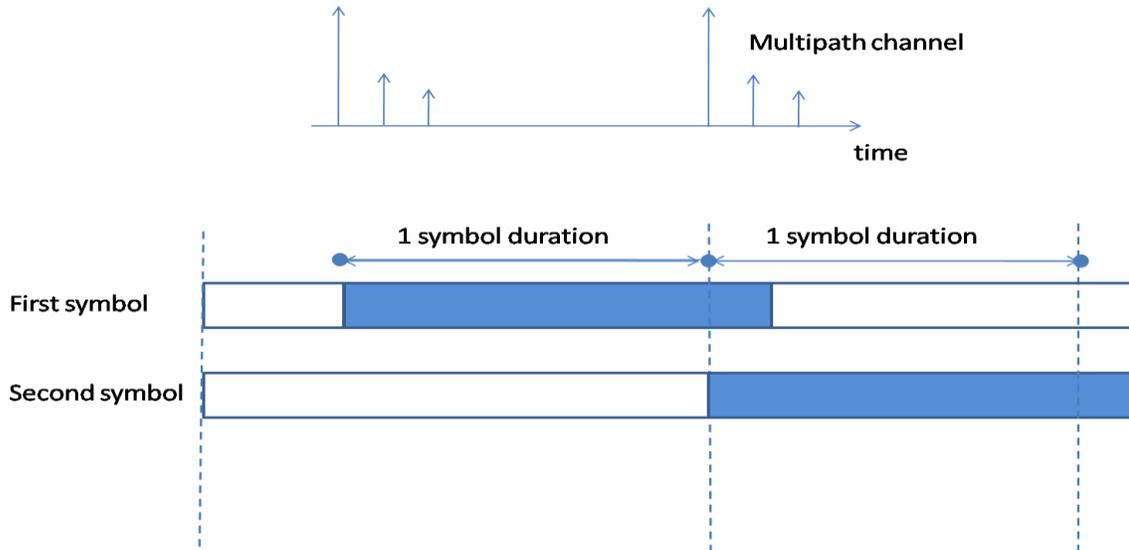


Fig.3.7 Received OFDM Symbols after passing through a multipath channel without guard interval

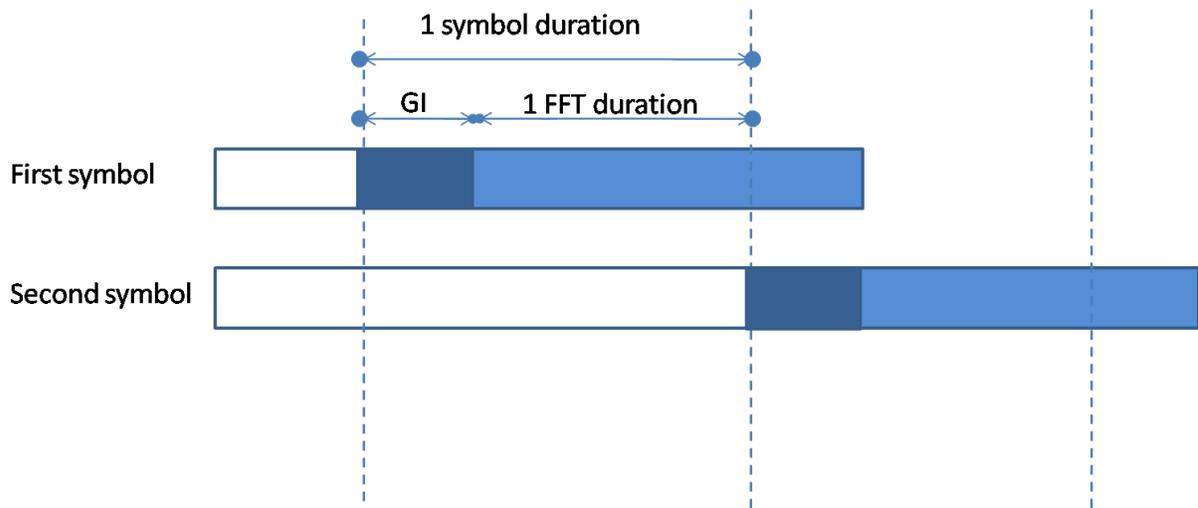


Fig.3.8 Received OFDM Symbols after passing through a multipath channel with guard interval

3.7 OFDM System design considerations

OFDM system design issues aim to decrease the data rate at the subcarriers, hence, the symbol duration increases and as a result, the multipath effects are reduced effectively. The insertion of higher valued CP will bring good results against combating multipath effects but at the

same time it will increase loss of energy. Thus, a tradeoff between these two parameters must be done to obtain a reasonable system design.

3.7.1 System design requirements:

OFDM system depends on the following requirements:

1. Available bandwidth: The bandwidth limit will play a significant role in the selection of number of subcarriers. Large amount of bandwidth will allow obtaining a large number of subcarriers with reasonable CP length.
2. Required bit rate: The system should be able to provide the data rate required for the specific purpose.
3. Doppler values: The effect of Doppler shift due to user movement should be taken into account.

3.7.2 System design parameters:

The design parameters are derived according to the system requirements. The design parameters for an OFDM system are as follows:

- Number of subcarriers: Large number of subcarriers will help to combat multipath effects. But, at the same time, this will increase the synchronization complexity at the receiver side .
- Symbol duration and CP length: A perfect choice of ratio between the CP length and symbol duration should be selected, so that multipath effects are combated and not significant amount bandwidth is lost due to CP.
- Modulation type per subcarrier: The performance requirement will decide the selection of modulation scheme. Adaptive modulation can be used to support the performance requirements in changing environment.

Chapter Four

OFDM system

4.1 System Model

In standard OFDM system, the information symbols are grouped into blocks and IDFT is performed on each block then a proper cyclic prefix (CP) extension is added before they are fed into the modulator and transmitted. At the receiver, DFT is performed on each received OFDM symbol after the CP is removed [50]. The main process is shown in Fig.4.1.

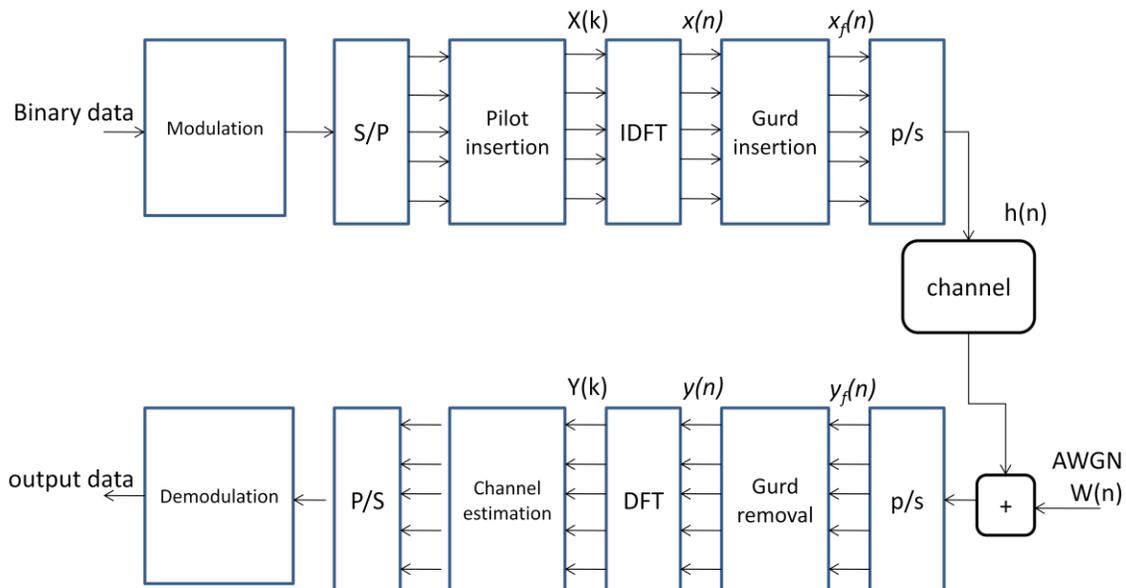


Fig.4.1: OFDM System Model

The OFDM system based on pilot channel estimation is given in Fig.4.1, [51-52]. The binary information is first grouped and mapped according to the modulation in signal mapper. After inserting pilots either to all sub-carriers with a specific period or uniformly between the

information data sequence, IDFT block is used to transform the data sequence of length N $\{X(k)\}$ into time domain signal $\{x(n)\}$ as in Eq.4.1.

$$x(n) = \sum_{k=0}^{N-1} X(k) e^{j\left(\frac{2\pi kn}{N}\right)} \text{ where } n = 0, 1, 2, \dots, N - 1 \quad (4.1)$$

Where N is the DFT length. Following IDFT block, guard time, which is chosen to be larger than the expected delay spread, is inserted to prevent inter-symbol interference. This guard time includes the cyclically extended part of OFDM symbol in order to eliminate inter-carrier interference (ICI) [53]. The resultant OFDM symbol is given by;

$$x_f(n) = \begin{cases} x(N + n), & n = -N_g, -N_g + 1, \dots, -1 \\ x(n), & n = 0, 1, \dots, N - 1 \end{cases} \quad (4.2)$$

Where N_g is the length of the guard interval. The transmitted signal $x_f(n)$ will pass through the time varying frequency selective fading channel with additive noise. The received signal is given by;

$$y_f(n) = x_f(n) \otimes h(n) + w(n) \quad (4.3)$$

Where $w(n)$ is Additive White Gaussian Noise (AWGN) and $h(n)$ is the channel impulse response. The channel response can be represented by Eq.4.4

$$h(n) = \sum_{i=0}^{L-1} h_i e^{j\left(\frac{2\pi}{N}\right) f_{di} T_n} \delta(\tau - \tau_i) \text{ where } 0 \leq n \leq N - 1 \quad (4.4)$$

Where $h(n)$ is the time varying channel impulse response, L is the total number of propagation paths, h_i is the complex impulse response of the i^{th} path, f_{di} is the i^{th} path Doppler frequency shift, τ is delay spread index, T_n is the sample period and τ is the i^{th} path delay normalized by the sampling time.

At the receiver, after passing to discrete domain through A/D and low pass filter, guard time is removed:

$$y_f(n) \text{ for } -N_g \leq n \leq N - 1 \quad (4.5)$$

$$y(n) = y_f(n + N_g) \quad n = 0, 1, \dots, N - 1 \quad (4.6)$$

$$y(n) = \sum_{i=0}^{L-1} h(n)x(n - i) + w(n) \quad (4.7)$$

Where $w(n)$ is a white Gaussian noise with zero mean and variance σ^2 . Then $y(n)$ is sent to DFT block.

$$Y(k) = \frac{1}{N} \sum_{n=0}^{N-1} y(n)e^{-j\left(\frac{2\pi kn}{N}\right)} \quad \text{where } k = 0, 1, 2, \dots, N - 1 \quad (4.8)$$

The received signal can be written in the form $Y(k) = DFT(y(n)) = X(k)H(k) + I(k) + W(k)$, [51].

Where :

The channel frequency response is,

$$H(k) = \sum_{i=0}^{L-1} h(n)e^{-j\left(\frac{2\pi ik}{N}\right)} \quad \text{where } k = 0, 1, 2, \dots, N - 1 \quad (4.9)$$

The FFT value of the noise is,

$$W(k) = \frac{1}{N} \sum_{n=0}^{N-1} w(n)e^{-j\left(\frac{2\pi nk}{N}\right)} \quad \text{where } n = 0, 1, 2, \dots, N - 1 \quad (4.10)$$

$I(k)$ in Eq.4.11, represent the ICI value caused by time-variant nature of the channel when the Doppler frequency is high, which is assumed to be a Gaussian random variable according to

central limit theory [54]. Both $I(k)$ and $W(k)$ are having bad effects on the useful signal, when the AWGN approach zeros (high SNR) the total noise becomes dominantly ICI [54].

$$I(k) = \frac{1}{N} \sum_{m=0, m \neq k}^{N-1} X(m) \sum_{n=0}^{N-1} H(m) e^{j\left(\frac{2\pi n(m-k)}{N}\right)} \quad (4.11)$$

Following DFT block the pilots signals are extracted and the estimated channel $H_e(k)$ for the data sub channels is obtained in channel estimation block then the transmitted data is estimated by Eq.4.12.

$$X_e = \frac{Y(k)}{H_e(k)} \quad (4.12)$$

4.2 WSSUS Channel:

A time-varying frequency-selective wireless channel is usually modeled as a Wide-Sense Stationary Uncorrelated Scattering (WSSUS) process [55-57]. The impulse response of a WSSUS channel is expressed as in Eq.4.13.

$$H(t, \tau) = \sum_{k=1}^L \alpha_k(t) \delta(\tau - \tau_k) \quad (4.13)$$

which describes the propagation of waves through multiple paths of different delays τ_k and attenuation $\alpha_k(t)$, $\alpha_k(t)$ are Wide-Sense Stationary (WSS) complex Gaussian processes, and are uncorrelated for different paths. Therefore, the autocorrelation function of the impulse is given by;

$$E[H(t, \tau), H^*(t - \Delta t, \tau - \Delta \tau)] = P_H(\Delta t, \tau) \delta(\Delta \tau) \quad (4.14)$$

The function $P_H(\Delta t, \tau)$ is the autocorrelation of the impulse response at the delay τ with the time difference Δt .

The Fourier transform of $P_H(\Delta t, \tau)$ with respect to the time difference Δt is the scattering function $S(f, \tau)$ of the channel which is given by;

$$S(f, \tau) = \int_{-\infty}^{\infty} P_H(\Delta t, \tau) e^{-j2\pi f \Delta t} d\Delta t \quad (4.15)$$

where f can be explained as the Doppler frequency. The scattering function is a measure of the average power output as a function of the time delay τ . The delay power spectrum is defined as in Eq.4.16

$$\rho_H(\tau) = \int_{-\infty}^{\infty} S(f, \tau) df \quad (4.16)$$

and the Doppler power spectrum is given by;

$$S(f) = \int_{-\infty}^{\infty} S(f, \tau) d\tau \quad (4.17)$$

The width of the delay power spectrum is referred to as the maximum delay spread, and the width of the Doppler power spectrum the maximum Doppler frequency.

A typical approximation for the delay power spectrum is exponential as shown in Eq.4.18 [58].

$$\rho_H(\tau) = \frac{1}{\tau_{max}} e^{\frac{-\tau}{\tau_{max}}} \quad (4.18)$$

where τ_{max} is the maximum delay spread of the channel. A typical approximation for the Doppler power spectrum is shown in Eq.4.19 [7],[34].

$$S(f) = \begin{cases} \frac{1}{\pi F_d} \cdot \frac{1}{\sqrt{1 - (f/f_d)^2}} & |f| < f_D \\ 0 & else \end{cases} \quad (4.19)$$

The autocorrelation of the different transmission paths $\alpha_k(t)$ is a zero order Bessel function that is depending on the time difference τ , and the Doppler shift F_d , $J_0(2\pi F_d \tau)$. The factor that controls how much the channel varies between two successive symbols is $F_d \tau$. The channel Autocorrelation Function (ACF) is shown in Fig.4.2, for maximum Doppler frequencies $F_d=70$ Hz and $F_d=150$ Hz, the higher the mobile speed the higher the fading rate, and accordingly the faster the time-variation of the channel. The Doppler power spectrum can therefore be obtained by taking the Fourier transform of the ACF. The resulting Doppler power spectrum of $H(t)$ is band-limited and U-shaped. Moreover, it exhibits twin peaks at F_d . Eq.4.19 is often referred to as Jake's Doppler power spectrum [58], as shown in Fig.4.2

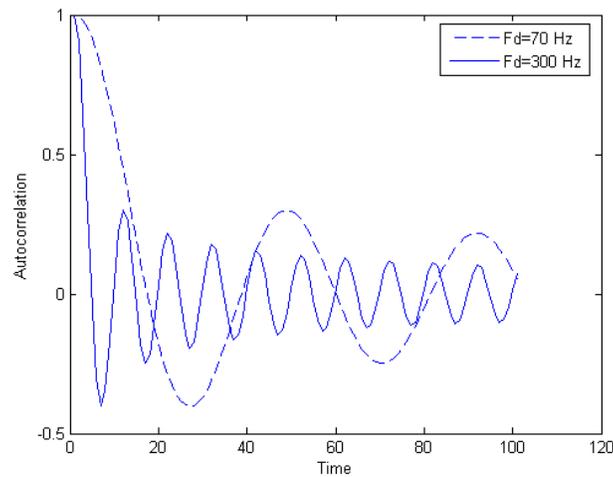


Fig.4.2 Channel ACF

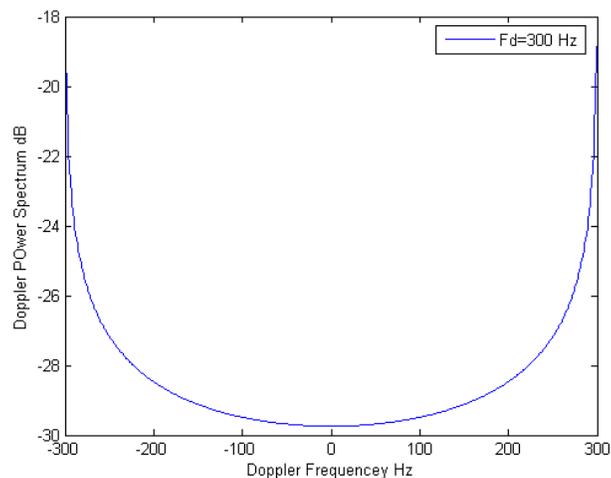


Fig.4.3 Jake's Doppler Power Spectrum.

4.3 AR Modeling of WSSUS Fading Channels:

To exploit the statistical properties of the fading channel given by Autocorrelation Function (ACF), the fading process over the m^{th} carrier can be modeled by a p^{th} order AR process, denoted by AR(p) and defined as shown in Eq.4.20 [58].

$$h(n) = - \sum_{i=0}^{p-1} a_i h(n-i) + u(n) \tag{4.21}$$

Where $\{a_i\} \rightarrow i = 0, 1, \dots, p-1$ are the AR model parameters and $u(n)$ is an Additive White Gaussian Noise. The relationship between the AR parameters and the fading process ACF is given by the well-known Yule-Walker equation (YW).

$$R_{hh} \mathbf{a} = -r_h \tag{4.22}$$

where R_{hh} is the fading channel autocorrelation matrix of size p by p , defined as in Eq.4.23.

$$R_{hh} = \begin{bmatrix} R_{hh}(0) & \dots & R_{hh}(-p+1) \\ \vdots & \ddots & \vdots \\ R_{hh}(p-1) & \dots & R_{hh}(0) \end{bmatrix} \tag{4.23}$$

and \mathbf{a} is a $p \times 1$ vector storing the AR parameter and given in Eq.4.24.

$$\mathbf{a} = [a_1 a_2 \dots a_p]^T \tag{4.24}$$

In addition, $r_h = [R_{hh}(1) R_{hh}(2) \dots R_{hh}(p)]^T$ is the $p \times 1$ channel autocorrelation vector where R_{hh} is the ACF which is given by Eq.4.25.

$$R_{hh}(n) = J_0(2\pi F_d T_s n) \tag{4.25}$$

where $J_0(\cdot)$ is the zero-order Bessel function of the first kind, T_s is the symbol period, and $(F_d T_s)$ denotes the Doppler rate. The corresponding PSD of the AR(p) process has the rational form as given in Eq.4.26 [59].

$$\Psi(f_n)_{AR} = \frac{\sigma^2}{|1 + \sum_{i=1}^p a_i \exp[j2\pi i f_n]|^2} \quad (4.26)$$

where f_n is the normalized frequency, and σ the variance of the driving process can be expressed as in Eq.4.27.

$$\sigma_u^2 = R_{hh}(0) + \sum_{i=1}^p a_i R_{hh}(-i) \quad (4.27)$$

Once the AR(p) parameters of the fading process are estimated, the autocorrelation function of the resulting AR(p) process has the form which is shown in Eq.4.28 [59].

$$\hat{R}_{hh}(n) = \begin{cases} R_{hh}(n), & 1 < n < p \\ -\sum_{i=1}^p a_i \hat{R}_{hh}(n-i), & n > p \end{cases} \quad (4.28)$$

Chapter Five

OFDM Channel Estimation

5.1 Introduction

Modulation can be classified as differential or coherent. When using differential modulation there is no need for a channel estimate, since the information is encoded in the difference between two consecutive symbols [60]. This is a common technique in wireless communication system and European DAB standard [61]. In differential modulation since no channel estimates is needed, this will reduces the complexity of the receiver. An interesting alternative of DPSK is differential amplitude phase shift keying [62], where a spectral efficiency greater than DPSK is achieved by using a differential coding of amplitude as well. Obviously, this requires a non uniform amplitude distribution. However, in wired systems, where channel is not changing with time, coherent modulation is an obvious choice. But, in wireless systems, the efficiency of coherent modulation makes it an ideal choice when the BER is high, such as in DVB [63]. Channel estimation in wired systems is straightforward, channel is estimated at startup, and since channel remains the same, therefore no need to estimate it continuously. Hence, in this thesis, we concentrate on channel estimation, regarding wireless OFDM. There are several basic techniques to estimate the radio channel in OFDM systems. The estimation techniques can be performed using time or frequency domain samples. These estimators differ in terms of their complexity, performance, practicality in applications to a given standard, and the a priori information they use. The a priori information can be subcarriers correlation in frequency, time and spatial domains.

Modulation of the OFDM subcarriers is analogous to the modulation in conventional serial systems. The modulation schemes of the subcarriers are generally QAM or PSK, in

conjunction with both coherent and non-coherent detection. If coherently detected modulation schemes are employed, then the reference phase of the OFDM symbol must be known, which can be acquired with the aid of pilot tones embedded in the spectrum of the OFDM symbol [64],[65]. For differential detection the knowledge of the absolute subcarrier phase is not necessary, and differentially coded signaling can be invoked either between neighboring subcarriers or between the same subcarriers of consecutive OFDM symbols.

Channel estimation in OFDM systems is a critical problem when the channel undergoes frequency selective and fast time variation. Iterative channel estimation is an adequate solution for this problem as it improves system performance and reduces pilot overhead when using Pilot-Symbol Aided (PSA) techniques. Comb-type channel estimation technique of inserting pilot has been considered in this thesis, since it has the ability to decrease the Doppler effect, and so enhance the mobility of the system. Accordingly it is good choice in channel estimation for OFDM in WiMAX application, because the mobility is an important requirement.

5.2 Channel estimation for OFDM system

Channel estimation has a long and rich history in single carrier communication systems. In these systems, the CIR is typically modeled as an unknown time-varying FIR filter, whose coefficients need to be estimated [66]. Many of the channel estimation approaches of single carrier systems can be applied to multi-carrier systems. However, the unique properties of multi-carrier transmission bring about additional perspectives that allow the development of new approaches for channel estimation of multi-carrier systems. In OFDM based systems, the data is modulated onto the orthogonal frequency carriers. For coherent detection of the transmitted data, these subchannel frequency responses must be estimated. In single carrier systems, the time domain channel can be modeled as a FIR filter, where the delays and coefficients can be estimated from time domain received samples, which are then transformed to frequency domain for obtaining the Channel Frequency Response (CFR). Alternatively, radio channel can also be estimated in frequency domain using the known data on frequency domain subchannels, instead of estimating FIR coefficients. The direct estimation of the channel for subcarriers treats each subcarrier as if the channels are independent [67]. If

coherent OFDM system is adopted, channel estimation becomes a requirement and usually pilot tones are used for channel estimation. Which known Pilot Symbol Assisted Modulation, (PSAM). The main idea of PSAM channel estimation is to multiplex known data streams with unknown data as shown in Fig.5.1. Conventionally the receiver firstly obtain tentative channel estimates at the positions of the pilot symbols by means of demodulation and then compute final channel estimates by means of interpolation.

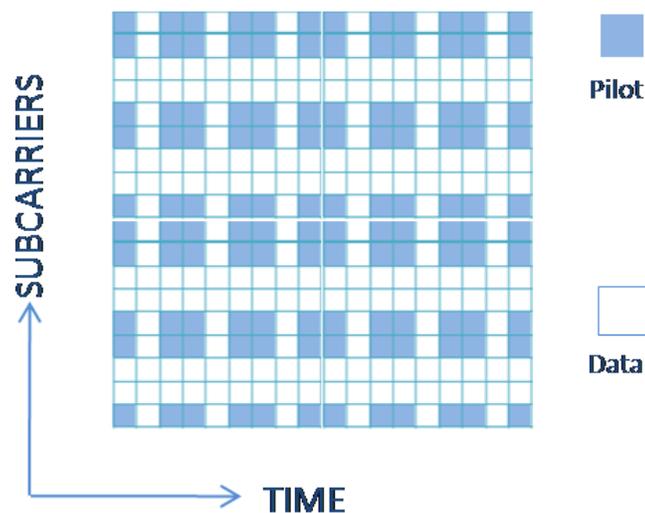
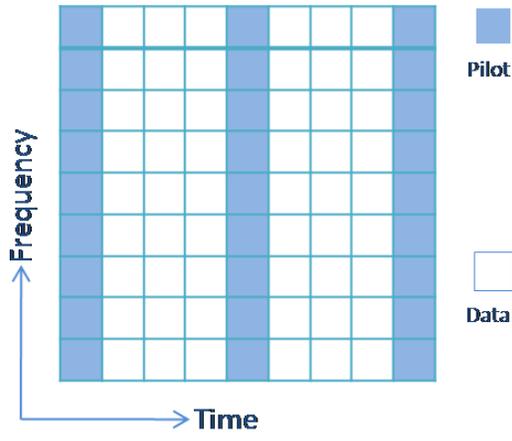


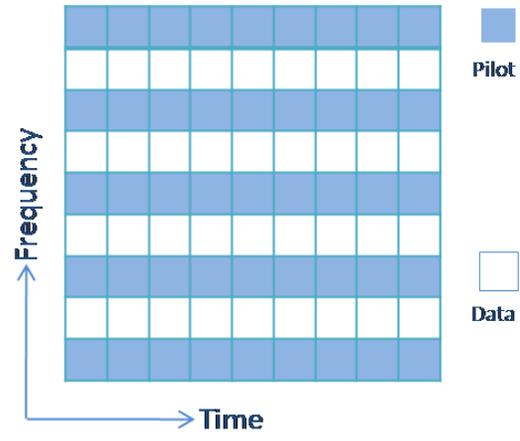
Fig 5.1 Pilots arrangement.

There are different patterns of pilots arrangement as we see in Fig.5.2. Block and comb pilot structures are among the most commonly used in the literature related to OFDM as well as in practical applications. Block pilots (see Fig.5.2(a)) assume a complete block of training data is sent periodically. The technique is appealing in slow to moderate fading. Frequency selectivity of the channel is handled at best, since all the subcarriers are used for channel estimation and thus no interpolation is needed. However, time selectivity is detrimental to block retraining, as obtained channel estimates degrade as the channel varies over time. Consequently, the retraining rate needs to be increased. For instance, the IEEE 802.11 standard for wireless LANs assumes that two complete blocks of training are sent prior to transmission of data. Data bearing blocks may also contain a few pilots for frequency synchronization purposes. Comb

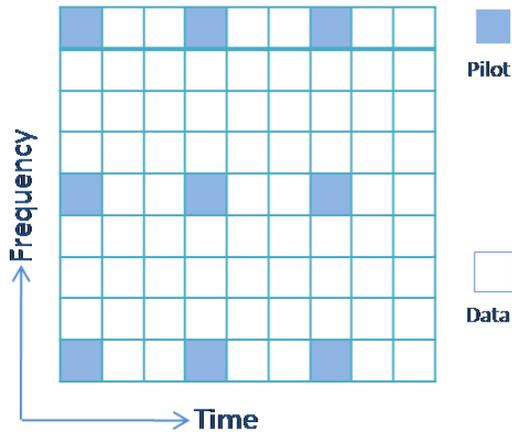
pilot structures (see Fig.5.2(b)) dedicate a specific set of subcarriers to sending training data over time. Time-selectivity is better handled with comb pilot structure than with block pilots. On the other hand, high frequency selectivity of the channel may be an issue. It may require reducing the pilot spacing and opting for more advanced interpolation procedures. Rectangular pilot structure (see Fig.5.2(c)) includes OFDM blocks with comb-pilots, which are sent periodically and not continuously. DVB-T systems use a specific pilot pattern (see Fig.5.2(d)) which combines block and rectangular structures. Blocks of pilots are meant for complete retraining, whereas comb pilots in blocks in between are used for channel tracking purposes. Hexagonal pilot designs (see Fig.5.2(e)) are optimal in the sense of sampling more efficiently the 2-D channel surface in the time-frequency plane than the rectangular pilot pattern. Partial pilots (see Fig.5.2(f)) which is a semi-blind channel estimation. The idea is to superimpose pilot to data symbols to facilitate channel estimation, without sacrificing the bandwidth efficiency while avoiding ambiguity problems. Also, the set of pilot arrangements may be adaptively selected based on the prediction of the channel estimation error at the receiver [45].



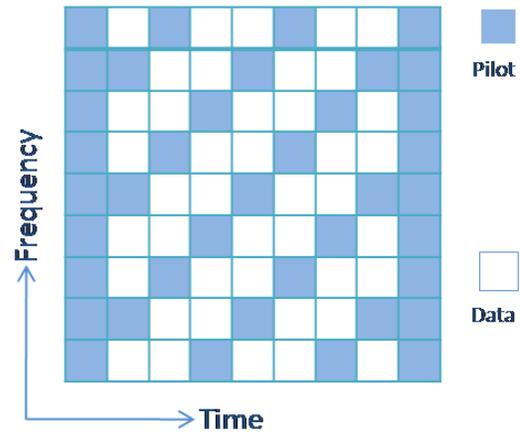
(a) Block type of pilots.



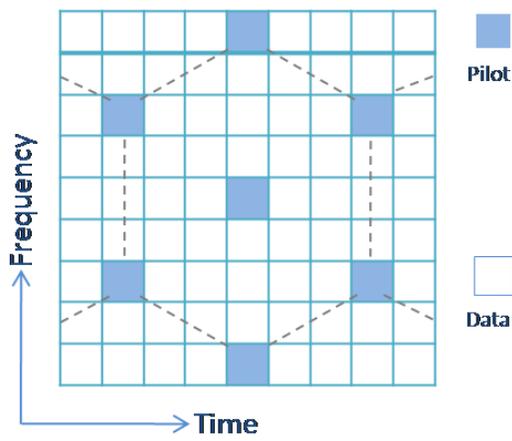
(b) comb type of pilots.



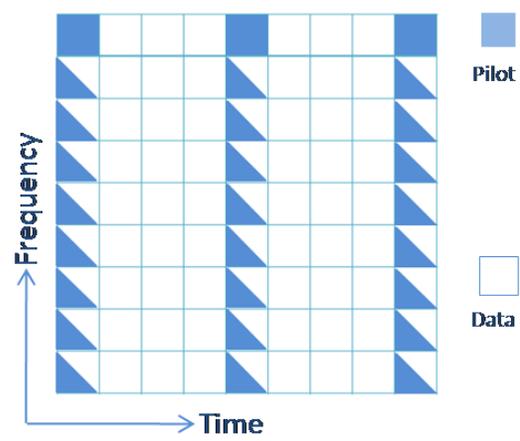
(c) Rectangular pilot pattern.



(d) Pilot pattern in DVB-T.



(e) Hexagonal pilot pattern.



(f) Partial pilots.

Fig.5.2 Pilot patterns

This chapter considers channel estimation in OFDM transmissions. In communication systems, channel estimation methods may be classified as blind, semi-blind or pilot-aided [68]. Blind algorithms do not require any training data and exploit statistical or structural properties of communication signals. Pilot-aided methods on the other hand rely on a set of known symbols interleaved with data in order to acquire the channel estimate [69]. Semi-blind methods combine a blind criterion with limited amount of pilot data, which improves both effective data rates and convergence speed. They also benefit from a larger sample support since both pilot and data are used for channel estimation. Blind, semi-blind and pilot-aided channel estimation in OFDM are successively reviewed in the following.

Here, channel estimation is performed in two steps.

- The first step is to estimate the channel frequency coefficients at the pilot symbols positions using LS estimator and Kalman estimator.
- Using the estimates of the channel frequency coefficients we then interpolate over channel frequency coefficients corresponding to the data symbols.

Our approach is to enhance the performance of the estimation process using Kalman estimator to estimate the CFR at the pilots, and comb pilots arrangement to track the time varying channel and so decrease the Doppler effects.

There are mainly two problems in the design of channel estimators for the wireless systems.

- The first problem is concerned with the choice of how the pilot information should be transmitted. Pilot symbols along with the data symbols can be transmitted in a number of ways, and different patterns yields different performances .
- The second problem is the design of an interpolation filter with both low complexity and good performance. These two problems are interconnected, since the performance of the interpolator depends on how pilot information is transmitted.

5.2.1 Blind algorithms:

The need for higher data rates motivates the search for blind channel identification and equalization methods. In OFDM, cyclic prefix occupies generally up to 20% of the transmitted data. Furthermore, if pilot symbols are used for channel estimation and synchronization purposes, those may require another 15-20% of the remaining data symbols. Therefore, blind estimators are of interest, especially in the case of continuous transmissions (e.g. DVB-T) or slowly time-varying channels (e.g. ADSL). The term blindness means that the receiver has no knowledge of the transmitted sequence and the channel impulse response. Channel identification, equalization or demodulation is then performed using only some statistical or structural properties of communication signal. Training data can then be either completely excluded or significantly reduced, and information symbols are transmitted instead. The processing in blind receivers is typically nonlinear. Common design goals for blind receivers algorithms are the following [45];

- Capability to identify any type of channel.
- Fast convergence to the desired solution.
- capability of tracking channel variations.
- Low computational complexity.

In wireless systems where bandwidth is the most precious resource, periodic training symbols can significantly reduce overall system capacity.

There are many blind algorithms for estimating OFDM channels. Some of them need to average over a number of OFDM symbols during which the channel must be static. Others work in a symbol-by-symbol manner, so can deal with fast time-varying channels. However, they assume independence of the channel for different OFDM symbols, and do not explore the time domain correlation of the channel. Other proposed blind OFDM channel estimation algorithm using cyclic correlations at the OFDM receiver. The CP is formed by copying the last T_{cp} -long part of the symbol waveform to the beginning, This algorithm does not require the cyclic prefix to be longer than the channel impulse response, but it is helpful to reduce the

error floor present in the un-shortened scenario, if this algorithm is combined with impulse response shortening. They suffer severe performance degradation in fast fading channels.

5.2.2 Semi-Blind algorithms:

Blind algorithms can also be used in cooperation with training symbols in order to achieve better performance, which are referred to as semi-blind methods. Semi-blind algorithms allow to outperform blind techniques by exploiting the knowledge of known symbols and properties of the transmitted signals. The objective of semi-blind channel estimation algorithms is to obtain better performance than blind algorithms while require fewer number of training symbols needed for training based channel estimation algorithms. In training based algorithms, the training symbols must be placed, close enough in time and frequency to accurately track a channel [70],[[71]. However, semi-blind algorithms works effectively even when the frequency-domain training symbol spacing is greater than the Nyquist spacing [72], thereby the resulting training symbols can be significantly reduced, especially when the channel varies rapidly both in time and frequency domains.

Added-pilots scheme is a semi blind algorithm which is closer to pilot-aided channel estimation than to blind techniques. The idea is to add pilot symbols directly to data symbols in time or frequency domain. In this way, no dedicated slot needs to be allocated to pilots, and the whole OFDM block may be used for information bearing symbols. The major drawback in Added Pilot Semi-Blind (APSB) channel estimation is that data interferes with pilots, and vice-versa. Thus, the influence of data should be minimized while estimating the channel, and conversely, the pilots have to be removed prior to data detection [45].

The semi-blind algorithm proposed intakes advantage of the channel and data information to get channel estimation and reduces the number of the training symbols needed to achieve this task. Specifically, the knowledge exploited includes [42]:

- Maximum delay spread of the channel.
- A prior channel statistics (mean and covariance).

- Redundancy of the input introduced due to the cyclic prefix.

Through exploiting these knowledge collectively, the channel estimation problem becomes a least-square (LS) problem. The algorithm first estimates the channel based on training symbols and then iterates between channel estimation and data detection. Thomas proposed a multi-user semi-blind channel estimator for OFDM based on the Maximum-Likelihood (ML) criteria [42].

5.2.3 Non blind algorithm training (pilot) methods:

The idea of the algorithms in blind algorithm is to exploit OFDM channel correlation, thus the statistics (mean, variance) of the channel must be known to make these algorithms work. In the non blind channel estimation methods, information of previous channel estimates or some portion of the transmitted signal are available to the receiver to be used for the channel estimation as in PSAS where a complete OFDM symbol or a portion of a symbol, which is known by the receiver, is transmitted so that the receiver can easily estimate the radio channel, by demodulating the received samples. The estimation accuracy can be improved by increasing the pilot density. However, this introduces overhead and reduces the spectral efficiency. In the limiting case, when pilot tones are assigned to all subcarriers of a particular OFDM symbol, an OFDM training symbol can be obtained (block type pilot arrangement) as in Fig.5.3. This type of pilot arrangement is usually considered for slow channel variation, but in case when channel varies between consecutive OFDM symbols, the training symbols should be inserted regularly within OFDM data symbols with respect to the time variation of the channel (Doppler spread) [42].

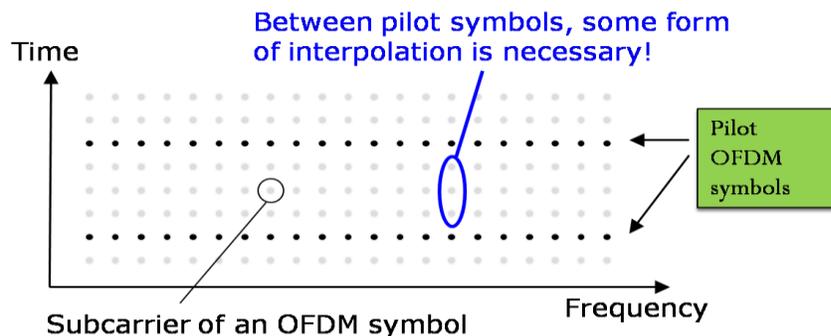


Fig.5.3 Block type Pilots arrangement.

For the PSAS channel estimation, the pilot spacing needs to be determined carefully. The spacing of pilot tones in frequency domain depends on the coherence frequency of the radio channel, which is related to the delay spread. According to the Nyquist sampling theorem, The spacing in the frequency domain D_p can be determined by frequency spacing between the subcarriers and the maximum delay spread of the channel. This can be expressed as shown in Eq.5.1:

$$D_p < \frac{1}{2} \cdot \frac{1}{2 \cdot \Delta f \tau_{max}} \quad (5.1)$$

Where Δf is the subcarrier bandwidth. When the above Equation is not satisfied, then the channel available at the pilot tones does not sample the actual channel accurately. In this case, an irreducible error floor in the estimation technique exists since this causes aliasing of the CIR tap in the time domain. When the channel is varying across OFDM symbols, in order to be able to track the variation of channel in time domain, the pilot tones need to be inserted at some ratio that is a function of coherence time (time variation of channel), which is related to Doppler spread. The maximum spacing of pilot tones across time, D_T is given by Eq.5.2

$$D_T < \frac{1}{2F_d T_s} \quad (5.2)$$

Where F_d is the maximum Doppler spread and T_s is the OFDM symbol duration. For comb-type pilot arrangements as in Fig.5.4, the pilot tones are often inserted for every OFDM symbols [67].

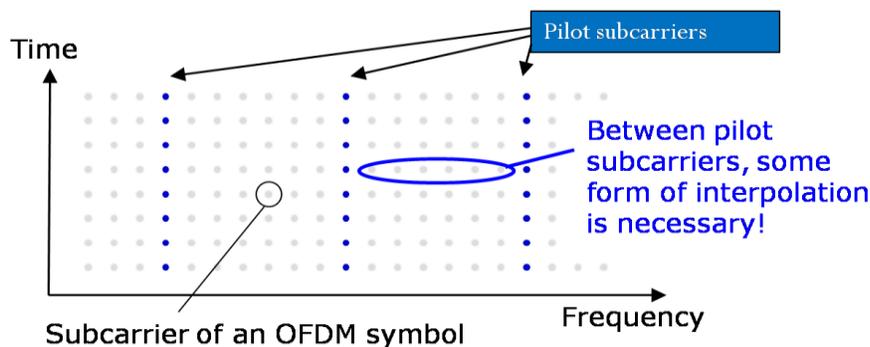


Fig.5.4 Comb Type Pilots arrangements

For block type arrangements, channel at pilot tones can be estimated by using LS or LMMSE estimation, and assumes that channel remains the same for the entire block. So in block type estimation, we first estimate the channel, and then use the same estimates within the entire block. LMMSE estimation has been shown to yield 10-12dB gain in SNR over LS estimation for the same mean square error of channel estimation [53]. In [73] low rank approximation is applied to linear MMSE by using the frequency correlations of the channel to eliminate the major drawback of MMSE, namely complexity.

Comb type pilot tone estimation, has been introduced to satisfy the need for equalizing when the channel changes even in one OFDM block. The comb-type pilot channel estimation consists of algorithms to estimate the channel at pilot frequencies and then CFR at data subchannels are obtained by interpolation between estimates at pilot locations, as will be discussed next. The estimation of channel at pilot frequencies for comb type based channel estimation can be based on LS, LMMSE or Least-Mean-Square (LMS). Obviously, each estimation method has its own advantages and disadvantages. Moreover, Kaman filter estimator will be introduced in this work and it as we will see later.

5.3 Analysis of Channel estimation based on comb type pilots pattern

5.3.1 Comb type pilots description:

For comb type pilot subcarrier arrangement, the K_P pilot signals $X_P(m)$, $m = 0, 1, 2, \dots, K_P$ are uniformly inserted into $X(k)$. That is, the total N subcarriers are divided into K_P groups, each with $L = N/K_P$ adjacent subcarriers. In each group, the first subcarrier is used to transmit pilot signal. The OFDM signal modulated on the k^{th} subcarrier can be given by;

$$X(k) = X(mL + l) \tag{5.3}$$

$$\text{Where } X(k) = \begin{cases} X_P(m) = 0, \text{ where } l = 0. \\ \text{inf. Data, where } l = 1, 2, \dots, L - 1 \end{cases} \tag{5.4}$$

$X_P(m)$ is the m^{th} pilot carrier value and the received pilot signal vector $Y_P = [Y_P(0), Y_P(1), \dots, Y_P(K_P - 1)]^T$ can be expressed as given in Eq.5.5 and Eq.5.6.

$$Y_p = X_p H_p + W_p \quad (5.5)$$

$$\text{Where } X_p = \begin{bmatrix} X_p(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & X_p(N_p - 1) \end{bmatrix} \quad (5.6)$$

H_p is the frequency response of the channel at pilot sub-carriers defined as;

$H_p = [H_p(0), H_p(1), \dots, H_p(K_p - 1)]^T$, and $W_p = [W_p(0), W_p(1), \dots, W_p(K_p - 1)]^T$, Is a vector of the Gaussian noise [51],[74].

5.3.2 LS estimator:

The estimation of pilot signals, based on least square (LS) criterion, is given in Eq.5.7 and Eq.5.8 respectively [53],[75].

$$H_{p,ls} = [H_{p,ls}(0), H_{p,ls}(1), \dots, H_{p,ls}(K_p - 1)]^T \quad (5.7)$$

$$H_{p,ls} = X_p^{-1} Y_p \quad (5.8)$$

The LS estimate of H_p is susceptible to Gaussian noise and ICI because the channel responses of data subcarriers are obtained by interpolating [53].

5.3.3 Kalman estimator

Kalman filter is an efficient recursive algorithm which estimates the state of a dynamic system from a series of noisy measurements. It has been applied in communication systems since 1970s. Kalman filter estimate the fading process by minimizing the estimation error variance $E[|H_p - \hat{H}_p|^2]$ where \hat{H}_p obtained using Kalman filter, [76]. It could also be applied to track the states of time varying channels, a p^{th} order AR model for H_p is presented as, [74].

$$H_p(l+1) = \sum_{i=0}^{p-1} F[i] H_p(l) + v(l) \quad (5.9)$$

$F[i]$ and the variance of the noise can be obtained by solving the YW equation using the ACF of the fading process. The correlation of the time-variant channel is modeled as in Eq.5.10.

$$E[h(l,i)h^*(l+1,j)] = \begin{cases} P_i J_0(2\pi F_d T_s), & i = j \\ 0, & i \neq j \end{cases} \quad (5.10)$$

where $P_i = [E h(l,i)]^2$, F_d is the maximum Doppler shift, T_s is OFDM symbol time, $J_0(\cdot)$ is the Bessel function of the first-kind and zero order and i, j are the frame and subcarrier index respectively. Using that

$$H_p(l) = Wh(l) \quad (5.11)$$

we obtain

$$E[H(l)^H H(l+1)] = J_0(2\pi F_d T_s) WPW^T \quad (5.12)$$

where P is a $N \times N$ diagonal matrix with diagonal elements P_i and W is a $P \times N$ partial DFT matrix obtained from a DFT matrix by deleting the rows that does not correspond to pilot symbols. Here P denotes the number of pilot symbols in one OFDM symbol and N is the total number of symbols [77].

So for first order AR channel model \mathbf{F} is modeled as in Eq.5.13:

$$\mathbf{F} = J_0(2\pi F_d b T_s) WPW^T \quad (5.13)$$

where b is a design parameter that determines the memory of the algorithm and is more deeply described in [77].

The state space model for Kalman filtering , state variable is defined as the channel states, which because it reflects the statistics of time varying channel, the noise is GWSSUS [52],[74].

$$H_P(l+1) = FH_P(l) + v_P(l) \quad (5.14)$$

Where $H_P(l)$ is the frequency response of the channel at the pilot symbols positions with dimension $P \times 1$ and $v_P(l)$ is the driving noise at the pilot symbols position with dimension $P \times 1$, which is with a zero-mean process that satisfies.

$$E[v(l) v^*(l+m)] = \begin{cases} V, & m = 0 \\ 0, & m \neq 0 \end{cases} \quad (5.15)$$

$v_P(l)$ which is mentioned above , is a driving noise with a zero-mean process that satisfies.

$$E[v(l) v^*(l+m)] = \begin{cases} V, & m = 0 \\ 0, & m \neq 0 \end{cases} \quad (5.16)$$

$$V = (1 - J_0(2\pi F_d b T_s))^2 W P W^T$$

The input-output relation of OFDM system is used to built the observation equation,

$$Y_P(l) = X_P(l)H_P(l) + W_P(l) \quad (5.17)$$

Using this AR-model it is possible to construct Kalman channel Estimator as given in Eq.5.18- Eq.5.21 where the gain, innovation process, channel impulse response and error covariance matrix are stated respectively;

$$K(l) = F P(l, l-1) X_P(l)^H [X_P(l) P(l, l-1) X_P(l)^H + \sigma^2 I]^{-1} \quad (5.18)$$

$$\alpha(l) = Y_P(l) - X_P(l) \hat{H}_P(l) \quad (5.19)$$

$$\dot{H}_p(l+1) = F\dot{H}_p(l) + K(l)\alpha(l) \quad (5.20)$$

$$P(l+1, l) = F[I - F^{-1}K(l)]P(l, l-1)F^H + V \quad (5.22)$$

Where $X_p(l)$ is a $P \times P$ diagonal matrix with pilot symbols on its diagonal, where P is the number pilot symbols and σ^2 is the variance of the AWGN.

5.3.4 The low-complexity Kalman estimator:

A low-complexity Kalman estimator was implemented, which is described in [77].

The complexity of Kalman estimator can be reduced by factorize $P(l, l-1)$ using Eigen value de-composition which is given by;

$$P(l, l-1) = UD(l)U^T \quad (5.22)$$

Where U is the unitary matrix whose columns is the Eigen vectors and $D(l)$ is a diagonal matrix with the Eigen values on its diagonal, then it is possible to construct the following Kalman estimator.

$$\alpha(l) = Y_p(l) - X_p(l)\dot{H}_p(l) \quad (5.23)$$

$$\dot{H}_p(l+1) = J_0(2\pi F_d K T_s)[\dot{H}_p(l) + U \cdot D(l)(D(l) + \sigma^2 I_m)^{-1} \cdot U \cdot C_p \cdot \alpha(l)] \quad (5.24)$$

To lower the complexity it is possible to update only the diagonal in D , i.e. the eigen values of $P(l, l-1)$ instead of updating the whole diagonal matrix D

After the estimation of the channel transfer function of pilot tones using Kalman Estimator an efficient interpolation technique is necessary, in order to estimate channel at data subcarriers, by using the channel information at pilot subcarriers. There are different types of interpolation schemes such as; Linear Interpolation, Spline-Cubic Interpolation, Low Pass Interpolation, Second order Interpolation and Time domain Interpolation [78],[79].

5.4 Analysis of Channel Estimation Based On Block Type Pilots pattern

In block-type pilot based channel estimation, OFDM channel estimation symbols are transmitted periodically, in which all sub-carriers are used as pilots. If the channel is constant during the block, there will be no channel estimation error since the pilots are sent at all carriers. The estimation can be performed by using either LS or MMSE. If inter symbol interference is eliminated by the guard interval, the required signal is given in Eq.5.10 in matrix notation as [53],[78];

$$Y = XEh + W \quad (5.25)$$

$$X = \text{diag}[X(0) X(1), \dots, X(N-1)] \quad (5.26)$$

$$Y = [Y(0) Y(1), \dots, Y(N-1)]^T \quad (5.27)$$

$$W = [W(0) W(1), \dots, W(N-1)]^T \quad (5.28)$$

$$H = [H(0) H(1), \dots, H(N-1)]^T = \text{DFT}(h) \quad (5.29)$$

$$E = \begin{bmatrix} W_N^{00} & \dots & W_N^{0N-1} \\ \vdots & \ddots & \vdots \\ W_N^{N-10} & \dots & W_N^{N-1N-1} \end{bmatrix} \quad (5.30)$$

Where $W_N^{nk} = \frac{1}{N} e^{-j2\pi(\frac{n}{N})k}$

If the time domain channel vector h is Gaussian and uncorrelated with the channel noise W , the frequency domain MMSE estimate is given by, [78]:

$$H_{MMSE} = FR_{hy}R_{yy}^{-1}Y \quad (5.31)$$

Where

$$R_{hY} = E\{hY\} = R_{hh} F^H X^H \quad (5.32)$$

$$R_{YY} = E\{YY\} = XFR_{hh}F^H X^H \quad (5.33)$$

are the cross covariance matrix between h and y and the auto-covariance matrix of Y . R_{hh} is the auto-covariance matrix of h , and σ^2 represents the noise variance $E\{W(k)^2\}$. The LS estimate is represented by Eq.5.33:

$$H_{ls} = X^{-1}Y \quad (5.34)$$

Which minimize $(Y - XEh)^H(Y - XEh)$.

When the channel is slow fading, the channel estimation inside the block can be updated using the decision feedback equalizer at each sub-carrier. Decision feedback equalizer for the k^{th} sub-carrier can be described as follows:

- The channel response at the k^{th} sub-carrier estimated from the previous symbol $\{H_e(k)\}$ is used to find the estimated transmitted signal $X_e(k)$

$$X_e(k) = \frac{Y(k)}{H_e(k)} \quad k = 0, 1, \dots, N - 1 \quad (5.35)$$

- $X_e(k)$ is mapped to the binary data through “signal demapper” and then obtained back through signal mapper as $X_{pure}(k)$
- The estimated channel $\{H_e(k)\}$ is updated by:

$$H_e(k) = \frac{Y(k)}{X_{pure}(k)} \quad k = 0, 1, \dots, N - 1 \quad (5.36)$$

Since the decision feedback equalizer has to assume that the decisions are correct, the fast fading channel will cause the complete loss of estimated channel parameters. Therefore, as the channel fading becomes faster, there happens to be a compromise between the estimation error due to the interpolation and the error due to loss of channel tracking. For fast fading channels, the comb-type based channel estimation performs better since it is able to track the variation of the channel in time [78].

5.5 Channel Interpolations

Once the channel frequency response (CFR) estimates have been obtained at the pilot subcarrier frequencies, they are extended to data subcarriers by interpolation. There are two types of interpolators:

- One dimensional interpolator
- Two dimensional interpolator

Later on, the theory of two-dimensional sampling was invoked, in an effort to both reduce pilot symbol rates and improve channel estimation performance. When the channel is probed simultaneously in both time and frequency domains, the overhead of pilot symbols may be reduced significantly as two-dimensional (2-D) processing captures simultaneously the correlation of the channel transfer function in both time and frequency. Two dimensional interpolators such as Wiener filtering method or Sinc interpolator are used instead of using two interpolators in time and frequency [80]. Two-dimensional time-frequency Wiener filter, which is optimal in the mean square error sense, assume knowledge of the doubly selective channel statistics, a condition which is hard to fulfill in realistic scenarios where the channel is not directly observable [45].

When pilot symbols are distributed within the OFDM block using, e.g., comb-type pilot structure, interpolation in the frequency direction is mandatory to obtain the CFR at data subcarriers. Piecewise-linear and piecewise-constant interpolation are among the simplest approach. Higher-order interpolation such as piecewise second-order polynomial interpolation

low-pass and Spline Cubic methods offer improved channel interpolation. The spacing between pilots or the amount of pilots are determined by the frequency selectivity of the channel, which relates to the maximum delay spread of the channel in time domain. With block-type of pilots, interpolation in the time domain is needed instead. Time-selectivity of the channel dictates the rate of retraining. It should be chosen smaller than the coherence time [45],[78].

- **Linear interpolation**

The channel estimation at the data subcarrier k where $L < k < (m+1)L$, using linear interpolation is given as follows:

$$H(k) = H(mL+l), \quad 0 \leq l \leq L \quad (5.37)$$

$$H(k) = H_p(m+1) - H_p(m) \left(\frac{m}{L} \right) + H_p(m) \quad (5.38)$$

- **Second order Interpolation**

In second order interpolation, channel estimate is given by:

$$H(k) = C_1 H_p(k-1) + C_0 H_p(k) + C_{-1} H_p(k+1) \quad (5.39)$$

$$\text{Where } \begin{cases} C_1 = \frac{\alpha(\alpha-1)}{2} \\ C_0 = -(\alpha-1)(\alpha+1), \alpha = \frac{l}{N} \\ C_{-1} = \frac{\alpha(\alpha+1)}{2} \end{cases}, \quad 0 \leq l \leq L \quad (5.40)$$

- **Low-Pass Interpolation**

The low-pass interpolation method is performed by inserting zeros into the original sequence and then applying a low-pass Finite-Length Impulse response (FIR) filter, which allows the original data to pass through it without any changing. This method

also interpolates such that the mean-square error between the interpolated points and their ideal values is minimized.

- **Spline Cubic Interpolation**

The spline cubic interpolation method produces smooth and continuous polynomial fitted to given data points. The fundamental idea behind Spline Cubic interpolation is based on draw smooth curves through a number of points.

- **Time Domain Interpolation**

Time Domain Interpolation method is a high resolution interpolation based on zero-padding and DFT/IDFT After obtaining the estimated $H_p(k)$, $k=0, \dots, N_p-1$ channel, we first convert it to time domain by IDFT as in Eq.5.41:

$$G_p(n) = \sum_{k=0}^{K_p-1} e^{j2\pi nk / K_p} \quad (5.41)$$

Where $n=0, \dots, K_p-1$

Then, by using the basic multi-rate signal processing properties, the signal interpolated by transforming K_p the points N into points with the following method:

$$M = \frac{K_p}{2} + 1 \quad (5.42)$$

$$G_N \begin{cases} G_p, & 0 \leq n \leq M - 2 \\ 0, & \frac{K_p}{2} \leq N - M \\ G_p(n - N + 2M - 1), & -M \leq n - N < -1 \end{cases} \quad (5.43)$$

The estimate of the channel at all frequencies is obtained by

$$H(k) = \sum_{n=0}^{N-1} G_N e^{-j\left(\frac{2\pi}{N}\right)nk} \text{ where } 0 \leq k \leq N - 1 \quad (5.44)$$

Chapter Six

Simulation Results

6.1 Introduction

An OFDM system is implemented and simulated using MATLAB to allow various parameters of the system to be varied and tested. The aim of doing the simulations is to measure the performance of OFDM system under different channel conditions, and to allow for different OFDM configurations to be tested. Simulations are carried out for different SNR. The LS and Kalman filter channel estimation methods have been used to estimate the channel at pilot frequencies, Interpolation techniques such as; linear interpolation, low-pass interpolation, and Spline Cubic interpolation are then applied to investigate the interpolation effects in estimation the CFR at the data subcarriers. The block diagram of the algorithm is illustrated in Fig.6.1.

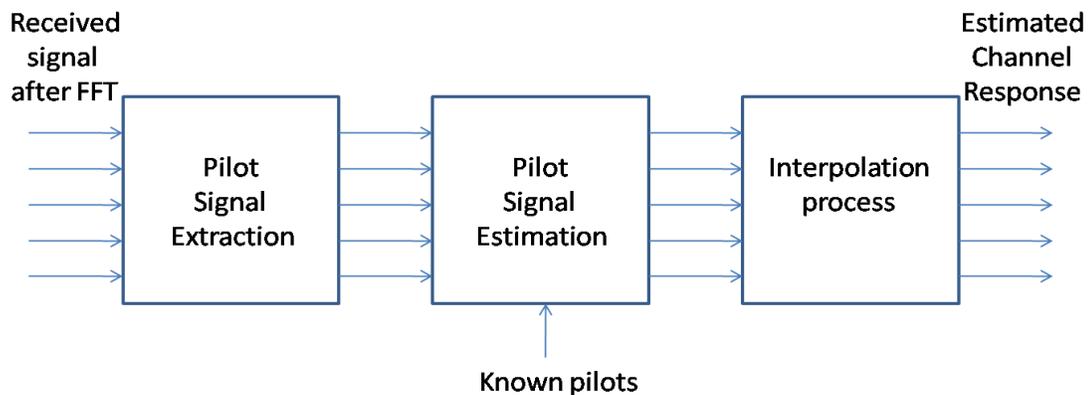


Fig 6.1 Block Diagram of Channel Estimation algorithm based on Comb-type Pilots arrangements

6.2 Kalman filter channel estimation based on comb-type pilots arrangement:

There are many parameters that affect the performance of Kalman filter channel estimation based on comb type pilots arrangement, in terms of BER such as the modulation technique, interpolation scheme, modulation method and Doppler shift .

OFDM system parameters used in the simulation are indicated in Table 6.1.

Table 6.1 Simulation Parameters-1

Parameter	Specification
Number Of Subcarrier	512
IFFT, FFT Size	512
Modulation Type	QAM,PSK
Pilot Ratio	1/8
Channel Model	GWSSUS
Guard Interval	1/32 from symbol period

Modulation techniques: Simulation results in Fig.6.2 illustrate the performance Kalman filter estimation method under different modulation techniques, it could show that 2-QAM modulation method has achieved better performance than 4, and 16-QAM. Moreover, BPSK method has given better performance than QPSK, 8-PSK and 16PSK as, obviously, BER decrease for a fixed SNR when the number of bits/symbol is reduced.

The interpolation schemes: The interpolation schemes have certain affect on the performance where linear, low-pass, and Spline-Cubic interpolation schemes have been used for channel estimation based on comb type pilots arrangement. The performance among the comb-type channel estimation with different interpolation techniques ranges from the best to the worst BER as follows: low-pass, Spline Cubic, linear interpolation, since the low-pass interpolation used in simulation does the interpolation such that the mean-square error between the interpolated points and the ideal is minimized, although Spline Cubic outperforms linear interpolation, but results in higher system complexity. Simulation results for the BER for different interpolation technique is shown in Fig.6.3.

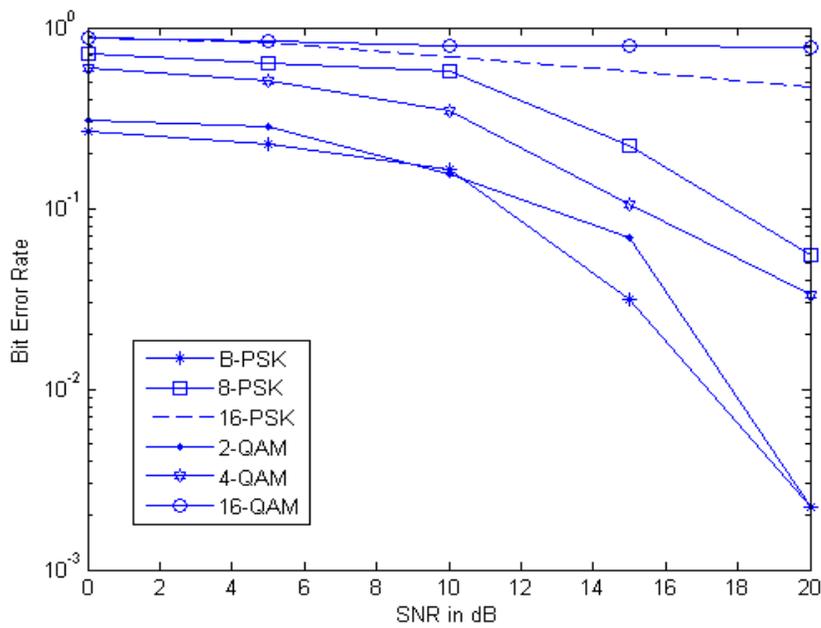


Fig.6.2 Performance for M-QAM and M-PSK modulation schemes.

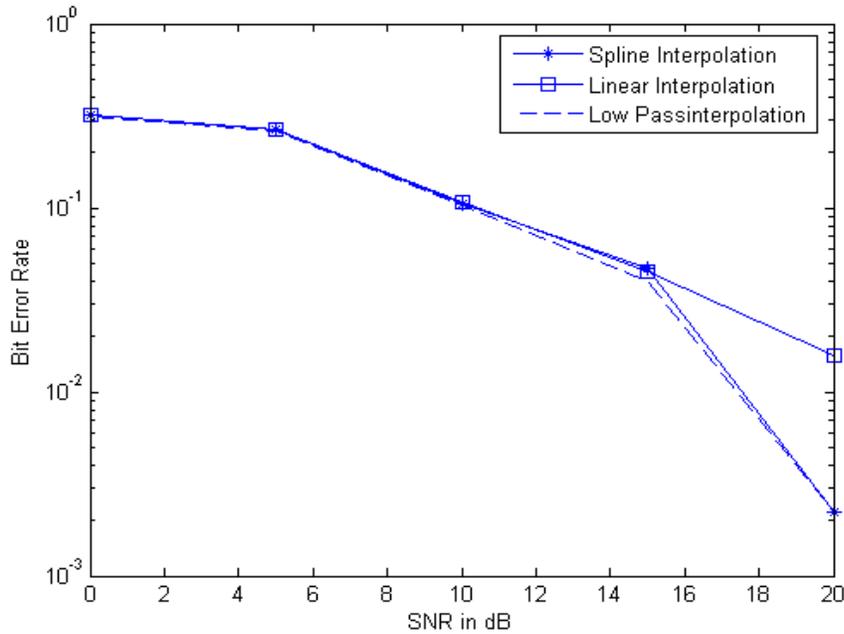


Fig.6.3 Interpolations methods performance using Kalman estimator.

Doppler effect: One of the advantages of comb type pilots arrangement in channel estimation is the ability to track the variation of the channel caused by Doppler frequency this is because every OFDM symbol has certain amount of pilots. In Fig.6.4, its clear that as the Doppler frequency increase the BER doesn't significantly increase at constant SNR, which mean that the system mobility has been enhanced, accordingly this method of pilot arrangements is interesting in OFDM system for WiMAX application. Doppler shift is the main reason to cause ICI and this will increases the noise level.

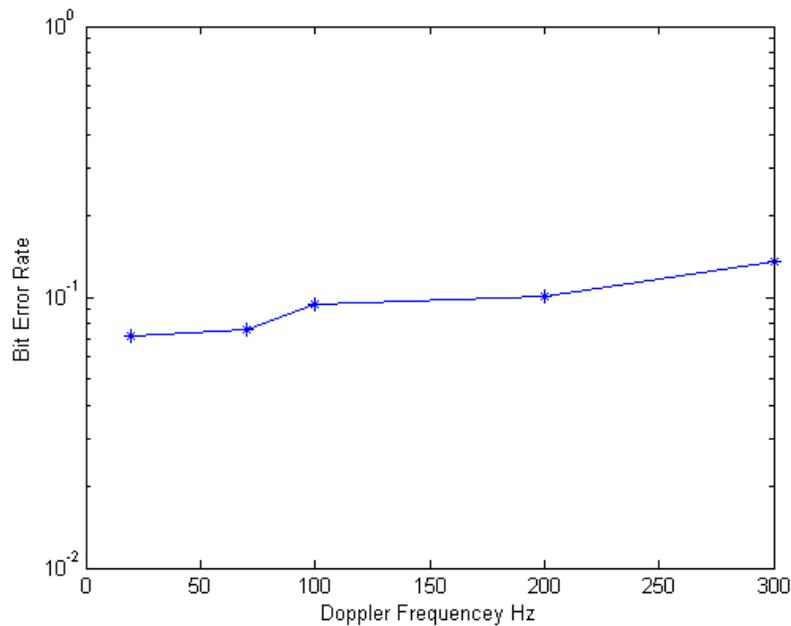


Fig.6.4 Doppler frequency effect for SNR=15dB.

Estimation methods: The system performance of using Kalman estimator to estimate the CFR at the pilots subcarriers in comb type pilots arrangement has outperformed LS estimator as illustrated in Fig.6.5. But this become clear as he SNR increases, However the complexity of the Kalman estimator must be considered in the system design. LS estimator is particularly interesting since it is one of the most simple estimation methods.

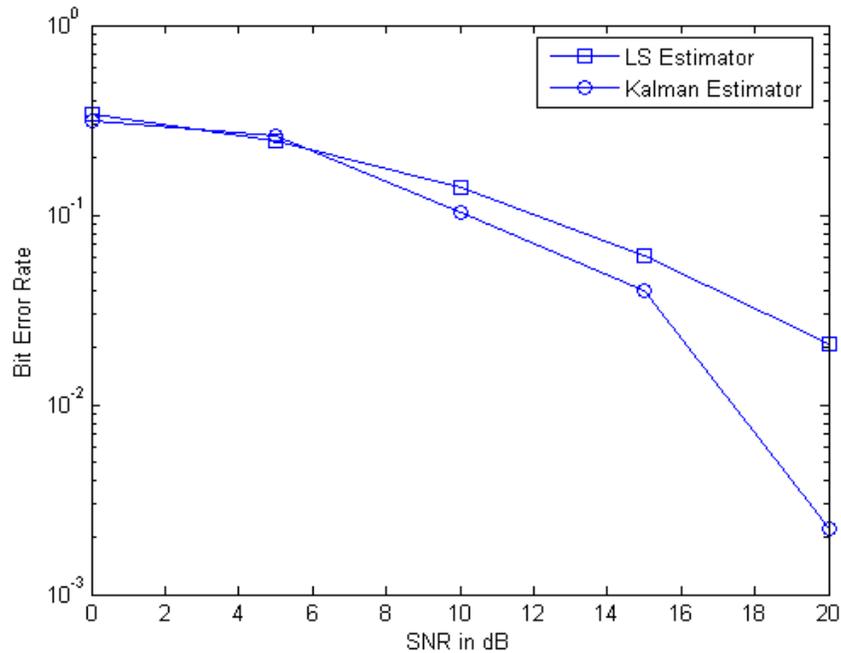


Fig.6.5 Kalman and Least square estimators performance

6.3 OFDM system over Rician fading channel

In this section we have analyzed the BER performance in OFDM system over frequency selective Rician fading channel which is implemented using MATLAB function (ricianchan.m). OFDM system parameters used in the simulation are indicated in Table 6.2.

Table 6.2 Simulation Parameters-2

Parameter	Specification
Number Of Subcarrier	512
IFFT, FFT Size	512
Modulation Type	BPSK
Pilot Ratio	1/8
Channel Model	Rician K=11 dB
Guard Interval	1/32 from symbol period

The LS channel estimation method has been used as an estimation method to estimate the CFR pilot frequencies, then interpolation techniques (linear interpolation, low-pass interpolation, Spline Cubic interpolation) are applied to investigate the interpolation effects.

Fig.6.6 illustrates the effect of the modulation technique used, it is obvious that 2-QAM modulation method has achieved better performance than 4, and 16-QAM and that BPSK method has given better performance than 8-PSK and 16-PSK this is because number of bit/symbol have been increased. Thus, a compromise between the BER performance and the bandwidth available has to be found.

Fig.6.7 illustrates the interpolation scheme effect on the performance, where linear, low-pass, and, Spline-Cubic interpolation schemes have been used for the interpolation between the estimate CFR at the pilots subcarrier. The performance among the interpolation techniques ranges from the best to the worst BER as follows: low-pass, Spline Cubic and linear interpolation.

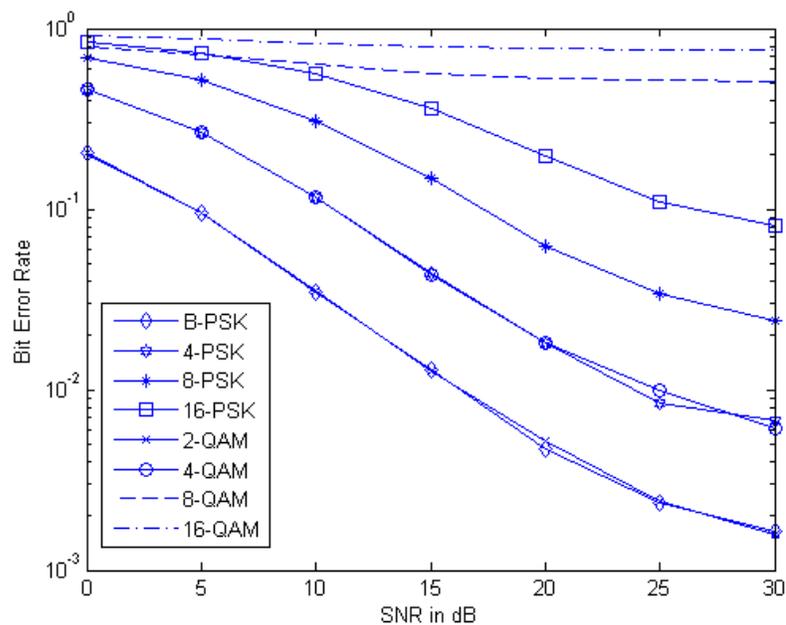


Fig.6.6 M-QAM and M-PSK modulation schemes performance

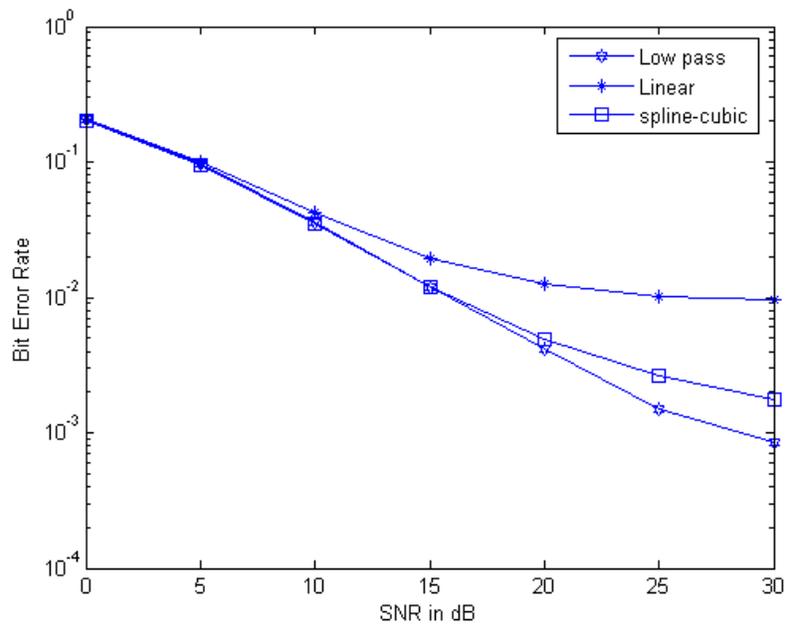


Fig.6.7 Different interpolation methods performance

The channel estimation can be improved using more pilot symbols as shown in Fig.6.8. However, this causes data rate reduction or bandwidth expansion.

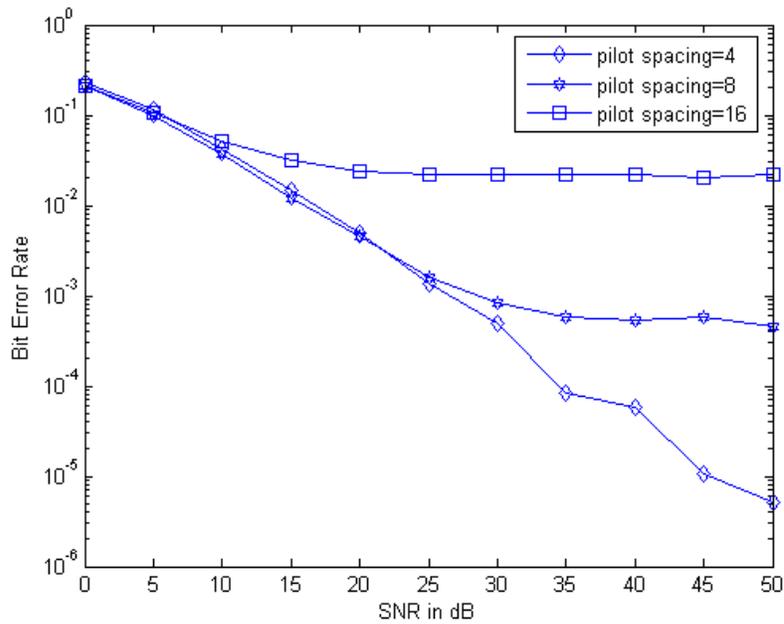


Fig.6.8 Number of pilots in OFDM symbol effects

6.4 Comparison between AR channel model and Rician fading channel

A comparison between 4th AR channel model which is used to generate a fading process and Rician fading channel model has been illustrated in Fig.6.9. It is clear that Rician channel model outperforms AR channel when using the same interpolation technique. LS estimation methods has been used .

FFT size which represents number of subchannels available for the transmission, affecting the performance of the estimation, it is obvious in Fig.6.10 and Fig.6.11 that increasing number of the available subchannels will decrease BER. In addition increasing FFT size means increase the number of samples per OFDM symbol, increase this will improve the performance of the system. This is expected since as the symbol duration is short, the ratio of distorted samples due to ISI to the total number of samples per OFDM symbol is large. However for large number of bits, the ratio of distorted samples due to ISI to the total number of samples per OFDM symbol is small value.

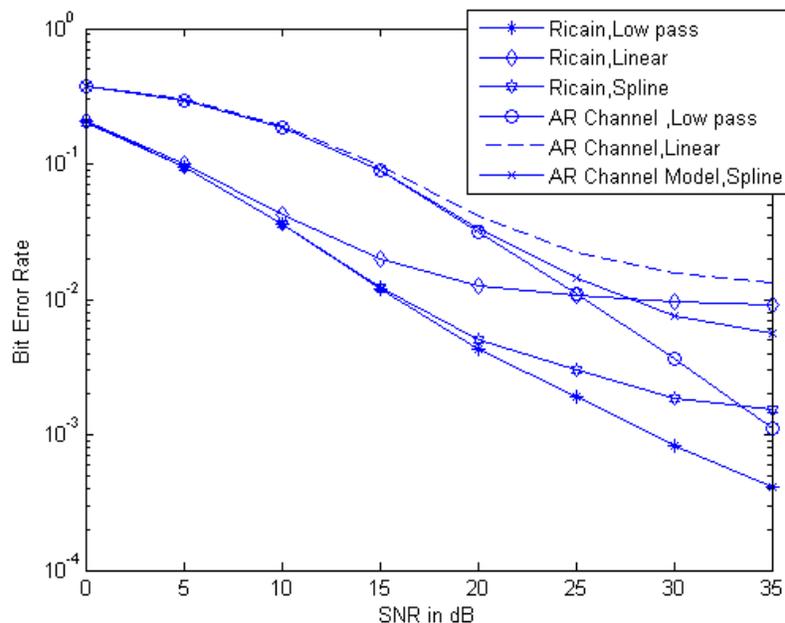


Fig.6.9. Interpolation methods performance.

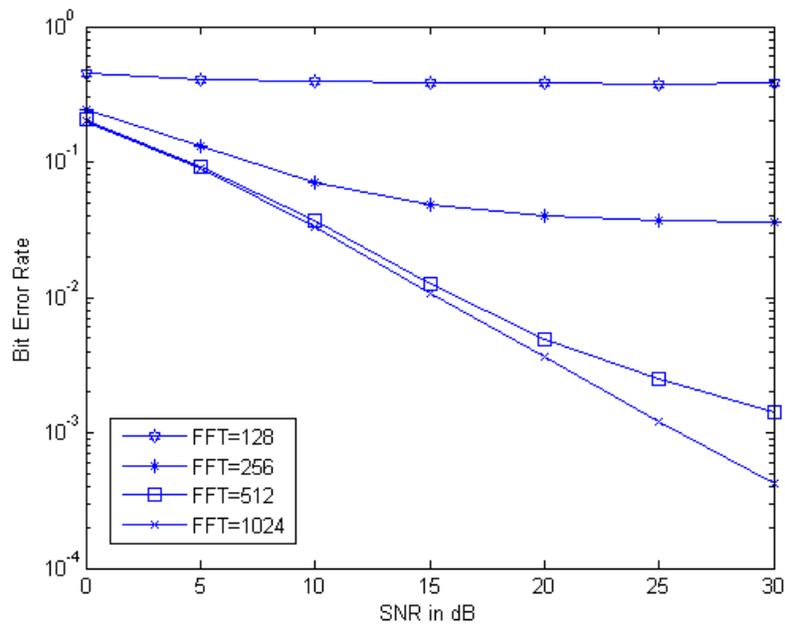


Fig.6.10 Number of subcarriers effect/Ricain fading Channel

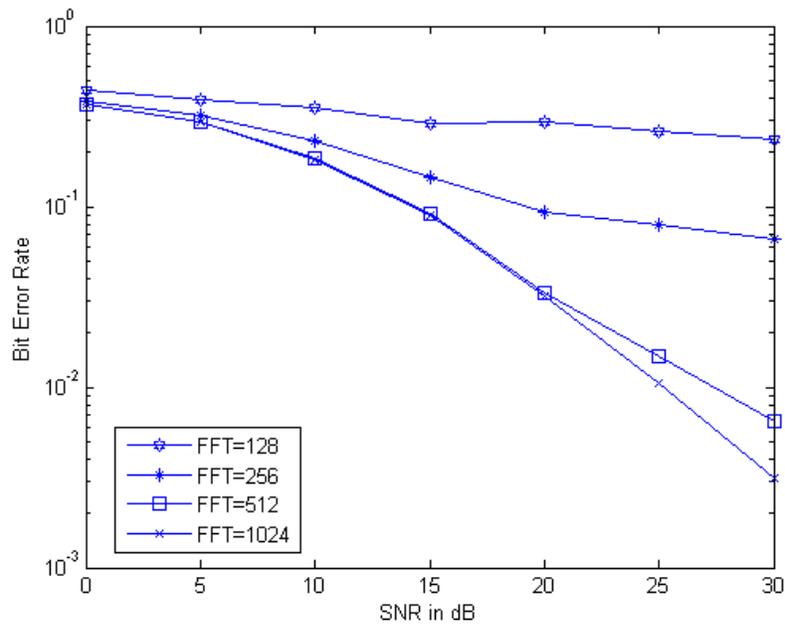


Fig.6.11 Number of subcarriers effect/AR Channel model

Channel time-variation are due to the relative motion between the base station and the mobile and/or the motion of the surrounding which in turn results in a Doppler spread. The Doppler spread is a measure of the relative frequency shift between the transmitted signal and the

received signal. The effect of Doppler frequency on both channels has been compared, it is obvious from Fig.6.12 that the estimation based in comb type pilots arrangement work well in high Doppler frequency. So this is estimation method is suitable to be used in high mobility wireless communication system as in WiMAX system.

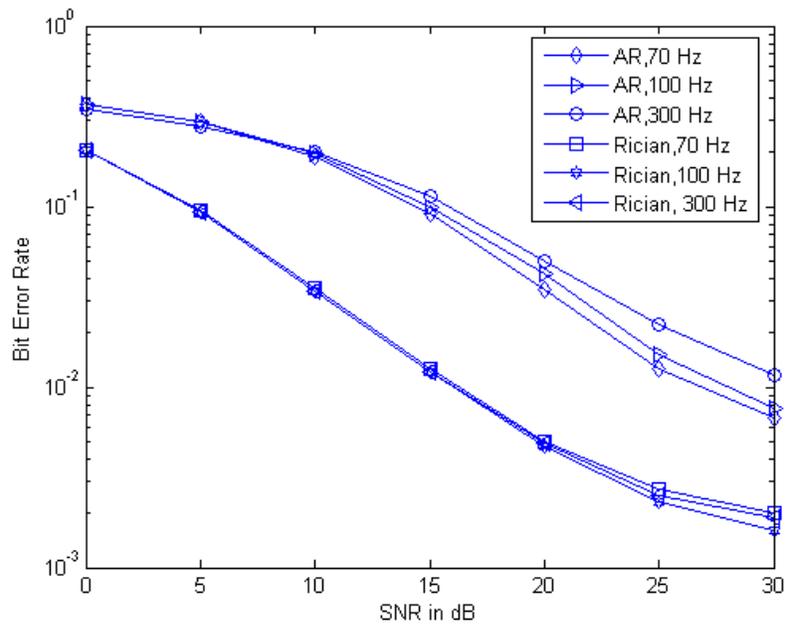


Fig.6.12 Doppler frequency effects

Chapter Seven

Conclusions

OFDM has been recently applied widely in wireless communication systems due to its high data rate transmission capability with high bandwidth efficiency and its robustness to multipath delay. In this thesis, the performance of OFDM system has been investigated in detail. One of the major advantages of OFDM systems is its robustness against multipath delay spread of the channel. Hence, its typical applications are in severe radio environments.

7.1 Benefits of OFDM

- OFDM system combats the ISI .
- High spectral efficiency because of overlapping spectra.
- Simple implementation by FFT
- Low receiver complexity as the transmitter combat the channel effect to some extends.
- Suitable for high data rate transmission
- High flexibility in terms of link adaptation
- Low complexity multiple access schemes such as orthogonal frequency division multiple access (OFDMA)

7.2 Drawbacks of OFDM are listed as follows:

- An OFDM system is highly sensitive to timing and frequency offsets.
- Demodulation of an OFDM signal with an offset in the frequency can lead to a high bit error rate.

7.3 Applications

OFDM has gained a big interest since the beginning of the 1990s as many of the implementation difficulties have been overcome. OFDM has been in use or proposed for a number of wired and wireless applications. DAB was the first commercial use of OFDM technology. OFDM has also been used for the DVB. OFDM under the acronym of Discrete Multi Tone (DMT) has been selected for ADSL.

7.4 Discussion

Channel estimation can be categorized as blind, semi-blind and Pilot-Aided channel estimation techniques. As blind methods are appealing due to their inherent bandwidth efficiency, they are not likely to be used alone in commercial applications and products. Indeed, they suffer from ambiguities and have high computational complexity. Moreover, some channels may not be identifiable. However, they may be used to refine pilot based estimates for each symbol, without requiring any modification of the transmitted signal structure. Blind methods remain an important topic of research since they are a seminal part of any semi blind algorithm. Semi-blind processing incorporates a little amount of training in order to ensure better performance, improved tracking capabilities and resolve ambiguities. It offers a more feasible implementation of blind criteria to practical systems. Semi-blind methods may find an application in static scenarios (ADSL, DVB-T) as well as in ones with moderate mobility (fixed broadband wireless access, WLAN). Table 7.1 summarizes the key features in blind, semi-blind and pilot-aided channel estimation in OFDM, and allows their comparison [45].

We reviewed the ways of performing channel estimation in pilot-aided OFDM. One needs first to acquire channel estimates at pilot symbol locations. This task is usually accomplished using LS, Kalman estimation. Then, CFR estimates at data subcarriers are obtained by interpolation between estimates at pilot locations, as well. A good choice of the pilot pattern should match the channel behavior both in time and frequency domains. In this way, the best tradeoff between channel estimator performance and transmission efficiency is found.

Table 7.1 Comparison between blind, semi-blind and pilot-aided channel estimation

	Blind	Semi-blind	Pilot-aided
Training data	None	A single or a few pilot symbols	Pilot symbols multiplexed with data.
Complexity	High.	Moderate to high.	Low to moderate.
Quality of the estimates	Low to moderate.	Moderate to high.	High.
Suitability to wireless channel	Freq. selective & time-invariant.	Freq. selective & slowly time varying.	Freq. selective & time-variant.
Benefits	High effective data rates. No modification is required to transmitted signal structure.	No ambiguity. Trade-off between effective data rate and tracking capability. Large sample support. Outperforms pilot aided estimation with the same amount of pilot information	No ambiguity. Suitable for high mobility scenarios.
Weaknesses	Ambiguities. No channel tracking capability in general. Some channels are not identifiable.	Less robust to abrupt change in channel condition.	Lower effective data rates. Time and frequency selectivity of the channel may be an issue depending on the amount of pilot data.

Pilot-Aided processing is most commonly chosen in real-world systems, especially in fourth generation mobile wireless communications where mobility and high data rates are major requirements. Interleaving sufficient amount of known symbols among the transmitted data allows to track highly time-varying channels. Moreover, the quality of the channel estimates allows often choosing high-order symbol modulations, leading to increased data rates. Complexity issues may also dictate the choice of pilot-aided methods as blind or semi-blind techniques require an increased processing power and better SNR to reach similar performance targets. This issue is critical for battery operated and hence power limited mobile terminals. Finally, pilot symbols are almost always present in practical designs. They are needed for other purposes than channel estimation solely, e.g., for time and frequency synchronization.

Channel estimation based on comb type pilot arrangement was presented by giving channel estimation method at pilot frequencies and interpolation of channel at data frequencies. Simulation results shows that comb-type pilot based channel estimation with low-pass interpolation performs the best among all other comb based channel estimation algorithms. This is expected since the comb type channel estimation allows the tracking of time varying fading channels and low pass interpolation does the interpolation in such a way that mean square error is minimized.

Decreasing the pilot spacing improves the estimation of channel frequency response but decreases bandwidth efficiency. On the other hand, increasing the pilot spacing beyond the one specified by the sampling theorem decreases the accuracy of the channel estimation but increases the bandwidth efficiency. Hence, the chosen pilot density is a tradeoff between the performance of channel estimation and bandwidth efficiency. Also since comb type pilots arrangement in channel estimation is able to track the variation of the channel caused by Doppler and the ICI has been modeled as AWGN, it is observed that ICI increases the noise level, and so to compensate this the number of pilot subcarriers required for the same performance of no ICI case increases by a significant amount. Hence, one way of compensation of ICI is to increase the number of pilots in the frequency domain.

Kalman estimation methods has been proposed to estimate the multipath fading channel in OFDM system. It is outperform LS estimation methods, but this method increase the system complexity. It a trade of between the system complexity and the performance required according to the SNR used. Channel estimation method based on comb type arrangement, with Kalman estimator is a good choice in high data rate wireless system, like in WiMAX system application since it is achieved good performance, with reasonable complexity compared with other system. The estimators perform about the same for SNR lower than 15 dB. This is an interesting property which means that the choice of channel estimator is not that important in terms of symbol errors for low SNR. When choosing a channel estimating method for low SNR the focus should instead be on how much information the estimating methods needs and also how high its complexity is.

Channel estimation based on comb-type pilot arrangements in OFDM system has been investigated over frequency selective Rician fading channel and time varying channel which is modeled as AR model. The performance of channel estimation for Rician channel model is better than for AR channel model, since Rician channel has a LOS signal, which is improve the performance of the estimation process .

Different parameters were considered to see how they are affect the BER performance. For Modulation scheme, the 2-QAM and BPSK have the least BER compared to the rest of M-QAM and M-PSK .Low pass interpolation technique was the best among others for either Rician or AR model channel.

7.5 Future Works:

7.5.1 Further studies of interest

Since the interpolation method has such a great impact on the symbol error, it would be interesting to investigate different interpolation methods more thorough.

So this work can be easily extended, to develop an algorithm, in which by taking into consideration the estimated channel attenuations, first of all estimate the Doppler frequency. Now depending on the value of estimated Doppler, different pilot arrangements will be used. For lower values of Doppler equidistance pilot arrangement is better and for higher values of Doppler, comb arrangement is better. So a new system can be developed which automatically changes the pilot arrangements by estimating the values of Doppler frequency.

Comparing this system with other systems, but we must taking in considerations the complexity, performance, applications and the requirements.

7.5.2 Related Works

OFDMA is popular and active field of research. In the uplink of an OFDMA system, each user modulates a subset of subcarriers that is impaired by a specific carrier frequency offset. Hence, the synchronization of such systems is much more demanding compared to conventional OFDM. Low synchronization errors guarantee both efficient user separation and high data rates. Reliable CFO and timing estimators with low overhead are desired. Theoretical performance bounds as well as identifiability conditions are of high importance, especially in a blind estimation framework.

A multiband version of OFDM (MB-OFDM) is gaining momentum as a candidate standard to be adopted for ultra wideband (UWB) communications . The popularity of MB-OFDM stems from its abilities to address data throughput and range requirements, while maintaining low cost, computational complexity and power consumption. Also, the transmission spectrum may be easily shaped to comply with international regulations and may be extended as well for future enhancements. Frequency hopping across multiple bands offers improved diversity and multi-user access, but gives rise to new challenges in channel estimation and carrier frequency synchronization. Very efficient signal processing techniques are therefore necessary to allow the practical deployment of such systems.

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تحليل تقدير قناة لنظام OFDM في تطبيق WiMAX

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اشراف: د علام موسى

الملخص

مع النمو السريع لقطاع الاتصالات الرقمية في الأعوام الأخيرة ، تزايد الطلب على أنظمة نقل البيانات ذات السرعات العالية. بالإضافة لهذا، من المتوقع ان تستطيع أنظمة الاتصال اللاسلكية المستقبلية ان تدعم نطاق واسع من الخدمات بما فيها الفيديو، البيانات والصوت.

يعتبر نظام OFDM كمرشح واعد لضمان سرعة نقل بيانات عالية في الوسط اللاسلكي وذلك نظراً لمقاومته لـ ISI والتي تعتبر مشكلة شائعة تحد من سرعة عمليات نقل البيانات. كما ان OFDM هو تقنية لنقل البيانات بعد عملية تقسيم البيانات ذات معدل تردد عالي الى بيانات ذات تردد قليل ليسهل عملية نقلها بشكل متوازي.

يعمل OFDM على انه متعدد الناقل (Multi-carrier) ولذلك له عدة مميزات منها الكفاءة العالية للنقل وزيادة معدل النقل حتى عند تنقل و حركة المستخدم مما يجعل قناة الاتصال متغيرة و معتمدة على الزمن . الهدف في هذه الاطروحة هو تقليل تأثير (Doppler effect) والذي يعتبر من اهم اسباب مشكلة تداخل قنوات الاتصال في نظام OFDM وهو ما يعرف ب (ICI). وللحد من هذه المشكله يتم تقدير قناة الاتصال عن طريق استخدام الرمز الموجه ، من خلال بث رموز موجهة مع البيانات وزيادة عدد هذه الرموز يؤدي الي تقليل تأثير ال (Doppler effect). في هذه الرسالة سنحلل أشكال مختلفة من الرموز الموجهة من خلال دراسة نسبة خطأ الاستقبال وسنقترح الرموز الموجهة بطريقة (comb-pilots arrangement).

يتم تقدير قنوات الاتصال التي تحمل الرموز الموجه باستخدام خوارزمية ال LS وال Kalman filter) حيث يتفوق (Kalman filter) على خوارزمية ال LS عندما كانت SNR اكثر من 10 ديسيبل وهذا يعطينا حرية الاختيار بين الطريقتين حسب ال SNR المعمول بها ومفاضله بين نسبة التعقيد في الخوارزميه و الاداء لهذه الخوارزميه .

بعد تقدير قنوات الاتصال التي تحمل الرموز الموجه يتم تقدير قنوات الاتصال التي تحمل البيانات باستخدام طرق مختلفه لل (Interpolation) مثل (Linear , Spline cubic and Low pass)

تم مقارنة نسبة خطأ الاستقبال باستخدام هذه الانواع من (interpolations) واستخدام انواع مختلفه من التضمين مثل (M-PSK and M-QAM).

