CHANNEL ESTIMATION IN OFDM SYSTEMS

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KING FAHD UNIVERSITY
OF PETROLEUM AND MINERALS

Dhahran, Saudi Arabia

August 2002

KING FAHD UNIVERSITY OF PETROLEUM AND MINERALS

DHAHRAN 31261, SAUDI ARABIA

DEANSHIP OF GRADUATE STUDIES

This thesis, written by **Kamran Arshad** under the direction of his thesis advisor and approved by his thesis committee, has been presented to and accepted by the Dean of Graduate Studies, in partial fulfillment of the requirements for the degree of **MASTER OF SCIENCE In Electrical Engineering.**

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Dedicated to

My Beloved Parents

and

Fiancee

ACKNOWLEDGEMENTS

In the name of Allah, the Most Gracious and the Most Merciful

All praise and glory goes to Almighty Allah (Subhanahu Wa Ta'ala) who gave me the courage and patience to carry out this work. Peace and blessings of Allah be upon His last Prophet Muhammad (Sallulaho-Alaihe-Wassalam) and all his Sahaba (Razi-Allaho-Anhum) who devoted their lives towards the prosperity and spread of Islam.

First and foremost gratitude is due to the esteemed university, the **King Fahd University of Petroleum and Minerals** for my admittance, and to its learned faculty members for imparting quality learning and knowledge with their valuable support and able guidance that has led my way through this point of undertaking my research work.

My deep appreciation and heartfelt gratitude goes to my thesis advisor **Dr. Asrar U.H. Sheikh** for his constant endeavour, guidance and the numerous moments of attention he devoted throughout the course of this research work. His valuable suggestions made this work interesting and knowledgeable for me. Working with him in a friendly and motivating environment was really a joyful and learning experience. I would also acknowledge Dr. Asrar for his cooperation in providing the TRL lab facility.

I extend my deepest gratitude to my thesis committee members Dr. Azzedine Zerguine and Dr. Mohammad Deriche for their constructive and positive criticism, extraordinary attention and thought-provoking contribution in my research. It was surely an honor and an exceptional learning to work with him.

Acknowledgement is due to my senior fellows Saad Azher and Mohammad Moin Uddin for helping me on issues relating to LATEX and MATLAB. I also appreciate the help provided by my fellow Sajid Khan in programming on MATLAB. Especially, i would like to thank my friend Aamir Siddiqui for his constant moral support throughout my thesis work.

Sincere friendship is the spice of life. I owe thanks to my house mates, colleagues and my friends for their help, motivation and pivotal support. A few of them are Ajmal Khan, Moin Uddin, Saad Azhar, Sajid Khan, Waleed Saif, Abdelwahab Bentercia, Aamir Siddiqui, Arshad Murshid, Owais Ullah, Junaid Laiq, Shiraz, Naji, Syed Akhter Ghazi and many others; all of whom I will not be able to name here. They made my work and stay at KFUPM very pleasant and joyful. My heartfelt thanks to my days old friends Hasib, Abdullah, Farhad and Waqas. They truly are my great friends, I wish we could be together again.

Family support plays a vital role in the success of an individual. I would like to thank my parents, siblings, my fiancee, my aunty (sister of my father) and other family members including all my uncles ,aunts and my loving cousins; from the core of my heart. Their prayers and encouragement always help me take the right steps in life.

May Allah help us in following Islam according to Quran and Sunna! (Aameen)

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Nomenclature

Abbreviations

LS	Least Squares
4G	Fourth Generation
BW	Bandwidth of Transmitted Signal
ISI	Inter Symbol Interference
ICI	Inter Carrier Interference
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
AWGN	Additive White Gaussian Noise
SNR	Signal to Noise Ratio
LMS	Least Mean Squares
MSE	Mean Square Error
FFT	Fast Fourier Transform
BER	Bit Error Rate
MCM	Multi Carrier Modulation

DAB	Digital Audio Broadcasting
DVB	Digital Video Broadcasting
IFFT	Inverse Fast Fourier Transform
OFDM	Orthogonal Frequency Division Multiplexing
CDMA	Code Division Multiple Access
FDMA	Frequency Division Multiple Access
TDMA	Time Division Multiple Access
AMPS	Advanced Mobile Phone Service
PSAM	Pilot Symbol Assisted Modulation
EM	Expectation Maximization
LMMSE	Linear Minimum Mean Square Error

English Symbols

N	Number of Sub Carriers
f_n	Carrier Frequency of n^{th} subcarrier
f_m	Doppler Frequency
T_s	OFDM Symbol Duration
$J(n)$	Cost Function at time n
Δf	Sub Carrier Bandwidth
Δp	Minimum Pilot Spacing

$u(n)$	Vector Representing the Input Sequence at time n
$d(n)$	Desired Response at time n
$h(n)$	Vector of Channel Coefficients at time n
$\hat{h}(n)$	Estimated Vector of Channel Coefficients at time n
$e(n)$	Error Signal at time n
R_{pp}	Auto Covariance Matrix of Pilot Estimates
R_{hp}	Cross Covariance Matrix Pilot and Channel Estimates
p	Channel Estimates at Pilot Frequencies

Greek Symbols

μ	The learning rate
τ	Delay Spread of Channel
τ_{rms}	RMS value of Power Delay Profile of Channel Taps
β	Constant depends on Signal Constellation
σ_n^2	Variance of the Noise

Operators

$E[\ ]$	Expectation Operator
$J_0(\ )$	Zeroth Order Bessel Function of the First Kind

$(\ )^H$	Hermitian Transpose
$Re[\ ]$	Real Part of [ ]
$Im[\ ]$	Imaginary Part of [ ]

THESIS ABSTRACT

Name: Kamran Arshad

Title: Channel Estimation in OFDM Systems

Degree: MASTER OF SCIENCE

Major Field: Electrical Engineering

Date of Degree: June 2003

With the rapid growth of digital communication in recent years, the need for high speed data transmission is increased. Moreover, future wireless systems are expected to support a wide range of services which includes video, data and voice. OFDM is a promising candidate for achieving high data rates in mobile environment, due to its resistance to ISI, which is a common problem found in high speed data communication. In OFDM, modulation may be differential or coherent. When using differential modulation there is no need for a channel estimate but its performance is inferior than the coherent system. Coherent modulation requires the channel estimation which gives better performance but with relatively more complex receiver structure. Pilot Symbol Assisted Modulation is used to achieve reliable channel estimates by transmitting pilots along with data symbols. In this thesis, we will analyze different pilot patterns in terms of BER and propose a new scheme for transmitting pilot symbols in wireless OFDM systems. We will also propose an adaptive scheme *of channel estimation in wireless OFDM systems, which tracks the multipath fading channel by using the LMS algorithm. OFDM systems are highly sensitive to frequency offsets, and one main reason for drifts in carrier frequencies is channel phase. We will also propose an adaptive scheme for channel phase compensation.*

Keywords: *OFDM, Pilot tones, Channel Estimation, Frequency Offsets, Multipath Fading Channel.*

King Fahd University of Petroleum and Minerals, Dhahran.
June 2003

ملخص الرسالة

الاســــــــــــــم: كمران ارشد

عنـــوان الـرسالــة: تقدير القناة في انظمة ال OFDM

الـتـخصــــــــص: الهندسة الكهربائية

تـاريـخ الـتـخـــرج: يونيو 2003م

مع النمو السريع لقطاع الأتصالات الرقمية في الأعوام الأخيرة , تزايد الطلب على انظمة نقل البيانات ذات السرعات العالية . بالأضافة لهذا , من المتوقع ان تستطيع انظمة الاتصال اللا سلكية المستقبلية ان تدعم نطاق واسع من الخدمات بما فيها الفيديو , البيانات و الصوت .يعتبر نظام OFDM كمرشح واعد لضمان سرعة نقل بيانات عالية في الوسط اللاسلكي و ذلك نظرا" لمقاومته ل ISI و التي تعتبر مشكلة شائعة تحد من سرعة عمليات نقل البيانات . في OFDM , التضمين يتم بطريقة تفاضلية او تلاحمية . عند استخدام التضمين التفاضلي لا توجد ضرورة لتقدير القناة اللاسلكية , و لكن هذه الطريقة اداؤها اقل من طريقة التضمين التلاحمي . النضمين التلاحمي يتطلب تقدير القناة اللاسلكية الذي يعطي اداء" افضل ولكن بالنقابل يتطلب مستقبل اكثر تعقيدا". التضمين المساعد ذو الرمز الموجه يستخدم للحصول على تقدير قناة موثوق , من خلال بث رموز موجهة مع البيانات. في هذه الرسالة سنحلل اشكال مختلفة من الرموز الموجهة من خلال دراسة نسبة خطأ الا ستقبال و سنقترح خطة جديدة لنقل الرموز الموجهة في انظمة OFDM. و سوف نقترح ايضا" خطة تكييفية جديدة لتقدير القناة اللاسلكية في انظمة OFDM, و التي تلاحق التلاشي في الممرات المتعددة من خلال استخدام خوارزمية LMS. انظمة OFDM حساسة جدا" للتغيرات الصغيرة في الترددات , و احد الأسباب لهذه التغيرات في الترددات هو زاوية القناة . و سنقوم باقتراح خطة تكييفية لتعويض التغيير الناجم في زاوية القناة.

درجة الماجستر في العلوم
جامعة الملك فهد للبترول و المعادن
الظهران- المملكة العربية السعودية
ربيع الثاني 1424ه

Chapter 1

Introduction

Radio transmission has allowed people to communicate without any physical connection for more than hundred years. When Marconi managed to demonstrate a technique for wireless telegraphy, more than a century ago, it was a major breakthrough and the start of a completely new industry. May be one could not call it a mobile wireless system, but there was no wire! Today, the progress in the semiconductor technology has made it possible, not to forgot affordable, for millions of people to communicate on the move all around the world.

The Mobile Communication Systems are often categorized as different generations depending on the services offered. The fist generation comprises the analog frequency division multiple access (FDMA) systems such as the NMT and AMPS (Advanced Mobile Phone Services) [2]. 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), PDC and 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 [3]-[4]-[5] with high data rates. To achieve such an objective, the next generation personal communication networks will need to be support a wide range of services which will include high quality voice, data, facsimile, still pictures and streaming video. 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 inter symbol interference (ISI). Orthogonal Frequency Division Multiplexing (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. By combining OFDM with Turbo Coding and antenna diversity, the link budget and dispersive-fading limitations of the cellular mobile radio environment can be overcomed and the effects of co-channel interference can be reduced [6].

1.1 Digital Communication Systems

A digital communication system is often divided into several functional units as shown in Figure 1.1.

The task of the source encoder is to represent the digital or analog information by bits in an efficient way. The bits are then fed into the channel encoder, which adds bits in a structured way to enable detection and correction of transmission errors. The bits from the encoder are grouped and transformed to certain symbols, or waveforms by the modulator, and waveforms are mixed with a carrier to get a signal suitable to be transmitted through the channel. At the receiver the reverse function takes place. The received signals are demodulated and soft or hard values of the corresponding bits are passed to the decoder. The decoder analyzes the structure of received bit pattern and tries to detect or correct errors. Finally, the corrected bits are fed to the source decoder that is used to reconstruct the analog speech signal or digital data input.

This thesis deals with the three blocks to the right in Figure 1.1: the modulator, the channel and the demodulator. The main question is how to design certain parts of the modulator and demodulator to achieve efficient and robust transmission through a mobile wireless channel. The wireless channel has some properties that make the design especially challenging: it introduces time varying echoes and phase shifts as well as a time varying attenuation of the amplitude(fade). This thesis focuses on the following parts in the modulator-demodulator chain.

Orthogonal Frequency Division Multiplexing (OFDM)has proven to be a modulation technique well suited for high data rates on time dispersive channels [7]. There are some specific requirements when designing wireless OFDM systems, for example, how to choose the bandwidth of the sub-channels used for transmission and how to achieve reliable synchronization. The latter is especially important in packet-based systems since synchronization 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 with an interpolation filter.

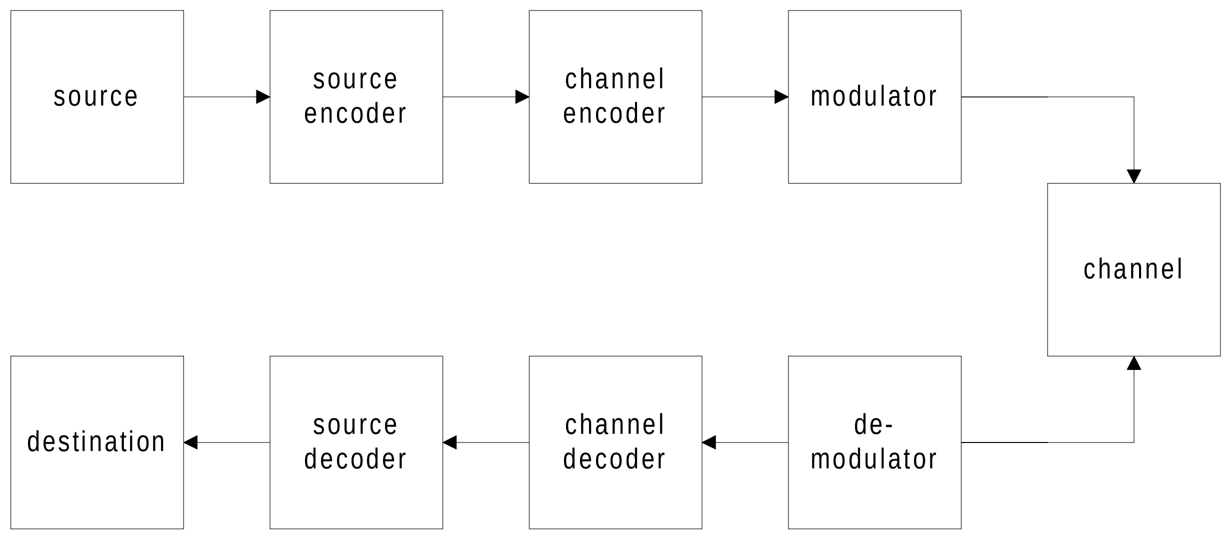

Figure 1.1: Functional Blocks in a Communication System

One of the challenging problems in the design of digital communication systems is how to choose the modulation scheme in order to get robust and efficient systems. This choice is greatly influenced by the environment in which system is supposed to work. Here in this thesis, Quadrature Amplitude Modulation (QAM) scheme is used. The influence of the channel can be described by its impulse response and often, there is also additive white Gaussian noise (AWGN) representing different disturbances in the system. For mobile or wireless applications, the channel is often described as a set of independent multipath components. The time varying impulse response can be described by [8],

$$x(t) = \sum_{i=1}^{M} a_i(t)\delta(\tau - \tau_i(t)) \tag{1.1}$$

where $a_i(t)$ denotes the complex valued tap gain for path number i, $\tau_i(t)$ is the delay of tap i, and δ is the Dirac delta function. Among the most important parameters when choosing the modulation scheme are the delay and the expected received power for different delays. Large delays for stronger paths mean that the interference between the different received signal parts can severe, especially when the symbol rate is high so that the delay exceeds several symbols. In that case one have to introduce an equalizer to mitigate the effects of intersymbol interference (ISI). Another alternative is to use many parallel channels so that the symbol time on each of the channels is long. This means that only a small part of the symbol is affected by ISI and this the idea behind orthogonal frequency division multiplexing, OFDM.

1.2 Wireless Systems

Wireless Systems are operating in an environment which has some specific properties compared to fixed wireline systems and these call for special design considerations. In a wired network, there are no fast movement of terminals or reflection points and the channel parameters are changing very slowly. In addition, time dispersion is less severe in a wired system, though it might still be a hard problem due to high data rates. In a mobile system the terminals are moving around, the received signal strength as well as the phase of the received signal, are changing rapidly. Further, the signal transmitted over the radio channel is reflected by buildings and other means of transportation on the ground, leading to different paths to the receiver, as shown in Figure 3.2.

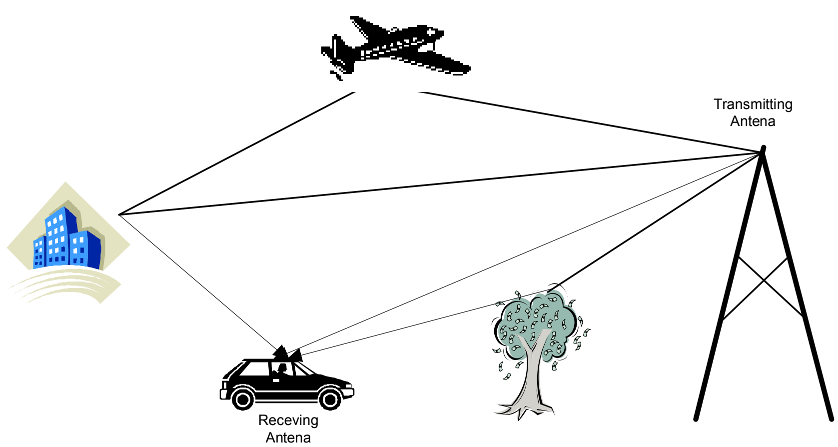

Figure 1.2: Multipath Reception

If the length of the paths differ, the received signal will contain several delayed versions of the transmitted signal according to the channel impulse response, defined in Equation (1.1). As mentioned earlier, the delays make it necessary to use complex receiver structures. In a mobile wireless system, the terminals are of course intended to be portable. This means that power consumption is important since batteries sometimes will power the equipment. Therefore, low complexity and low power consumption are properties that are even more desirable in wireless systems than in a wired system.

1.2.1 Packet-Based versus Circuit-Switched Systems

A wireless system can either be packet-based or circuit switched. In a packet based system the information bits are grouped and transmitted in packets, and transmission occur only when there is a need for communication. These systems are suitable for bursty traffic conditions, such as data communication. In circuit switched systems, a physical or virtual connection is established and occupied as long as communication proceeds. Circuit switched systems are well suited for real time traffic when delay is a limiting factor. In packet based systems, the receiver has to achieve synchronization in a very short time. It is hard to track channel variations between the packets, and therefore fast acquisition algorithms are required. In circuit switched systems, the receiver needs to enter in acquisition mode more seldom due to transmission over steady channels, therefore requirements on fast acquisition can be loosened in these systems. In circuit switched systems, we also require continuous channel tracking.

Today there is a trend towards more and more packet based systems due to increased data traffic. For example, both the third generation mobile systems based on W-CDMA and the HiperLAN/2 system based on OFDM use packet-based communication for data traffic.

1.2.2 Coherent versus Non-Coherent Systems

In general, coherent systems result in better detection performance compared to differential systems, but these require channel estimation in order to form time and phase references for the decisions. Differential schemes on the other hand require no channel estimation, but there is a performance loss compared to coherent detection [8].

In coherent schemes, the channel estimates are often achieved by multiplexing known, so called, pilot symbols into the data sequence and this technique is called Pilot Symbol Assisted Modulation (PSAM) [9]. PSAM was introduced by Moher and Lodge [10] and analyzed by Cavers [11] for single carrier systems. The receiver observes the influence of the channel on the pilot symbols and uses interpolation to get an estimate of the channel impact on data symbols. The receiver then removes that impact in order to make decisions. The pilot symbols transmit no data and therefore there is a small overhead causing a bandwidth expansion and an energy loss. Both these looses depends on the pilot-to-data symbol ratio.

1.3 Third Generation Wireless Networks

The expansion of the use of digital networks has led to the need for the design of new higher capacity communications networks. The demand for cellular-type systems in Europe is predicted to be between 15 and 20 million users by the year 2000 [12], and is already over 30 million (1995) in the U.S. [2]. Wireless services have been growing at a rate greater than 50% per year [2], with the current second generation European digital systems (GSM) being expected to be filled to capacity by the early 2000s [13]. The telecommunications industry is also changing, with a demand for a greater range of services such as video conferencing, Internet services, and data networks, and multimedia. This demand for higher capacity networks has led to the development of third generation telecommunications systems.

One of the proposed third generation telecommunication systems is the Universal Mobile Telecommunications System (UMTS), with the aim of providing more flexibility, higher capacity, and a more tightly integrated service. Other systems around the world are being developed, however many of these technologies are expected to be combined into the UMTS.

The World Wide Web (WWW) has become an important communications media, as its use has increased dramatically over the last few years. This has resulted in an increased demand for computer networking services. In order to satisfy this, telecommunications systems are now being used for computer networking, Internet access and voice communications. A WWW survey revealed that more than 60% of users access the Internet from residential locations, where the bandwidth is often limited to 28.8 kbps [14] . This restricts the use of the Internet, preventing the use of real time audio and video capabilities. Higher speed services are available, such as integrated-services digital network (ISDN). These provide data rates up to five times as fast, but at a much increased access cost. This has led to the demand of a more integrated service, providing faster data rates, and a more universal interface for a variety of services. The emphasis has shifted away from providing a fixed voice service to providing a general data connection that allows for a wide variety of applications, such as voice, Internet access, computer networking, etc.

The increased reliance on computer networking and the Internet has resulted in an increased demand for connectivity to be provided any where, any time, leading to an increase in the demand for wireless systems. This demand has driven the need to develop new higher capacity, high reliability wireless telecommunications systems. The development and deployment of third generation telecommunication systems aim to overcome some of the shortcomings of current wireless systems by providing a high capacity, integrated wireless network. There are currently several third generation wireless standards, including IMTS-CDMA, IMTS-TDD, IMTS-TF etc.

1.3.1 Evaluation of Telecommunication Systems

Many mobile radio standard have been developed for wireless systems thoughout the world, with more standard likely to emerge.

Most first generations systems were introduced in the mid 1980s, and can be characterized by the use of analog transmission techniques, and the use of simple multiple access techniques such as Frequency Division Multiple Access (FDMA). First generation telecommunications systems such as Advanced Mobile Phone Service (AMPS) [15], only provided voice communications. They also suffered from a low user capacity, and security problems due to the simple radio interface used.

Second generation systems were introduced in the early 1990s, and all use digital technology. This provided an increase in the user capacity of around three times [2]. This was achieved by compressing the voice waveforms before transmission [16].

Third generation systems are an extension on the complexity of second generation systems and are already introduced. The system capacity is expected to be increased to over ten times original first generation systems. This is going to be achieved by using complex multiple access techniques such as Code Division Multiple Access (CDMA), or an extension of TDMA, and by improving flexibility of services available.

Figure 1.3 shows the evolution of current services and networks to the aim of combining them into a unified third generation network. Many currently separate systems and services such as radio paging, cordless telephony, satellite phones, private radio systems for companies etc, will be combined so that all these services will be provided by third generation telecommunications systems.

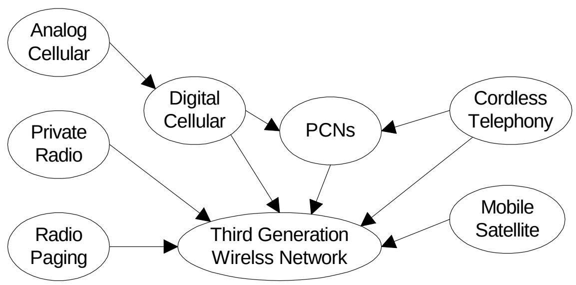

Figure 1.3: Evolution of current networks to the next generation of wireless networks

1.4 Fourth Generation Wireless Systems

Although carriers are reluctant to discuss 4G, vendors are always mapping future of 4G systems. Its still a decade away (at least), but 4G is already a big topic of discussion behind closed doors. Main advantages of 4G are its spectrum optimization, network capacity and faster data rates, however, carriers are still reluctant to discuss 4G, either because they refuse to take a public position on it when 3G roll-outs still are unfulfilled, or because they are in denial. But carriers soon will find that 4G is not going away. 3G systems are not enough for many services like data transfer between wireless phones or multimedia. Equipment vendors are coming together to speed the adoption of OFDM, which will be part of the 4G set of standards.

Orthogonal Frequency Division Multiplexing OFDM [7] is a multicarrier transmission technique, many carriers, each one being modulated by a low rate data stream share the transmission bandwidth. OFDM is similar to FDMA in that the multiple user access is achieved by subdividing the available bandwidth into multiple channels, that are then allocated to users. However, OFDM uses the spectrum much more efficiently by spacing the channels much closer together. This is achieved by making all the carriers orthogonal to one another, preventing interference between the closely spaced carriers.

In FDMA each user is typically allocated a single channel of certain bandwidth, which is used to transmit all the user information. The allocated bandwidth is made wider than the minimum amount required to prevent channels from interfering with one another. This extra bandwidth is to allow for signals from neighboring channels to be filtered out, and to allow for any drift in the center frequency of the transmitter or receiver. In a typical system, up to 50% of the total spectrum is wasted due to the extra spacing between channels.

TDMA partly overcomes this problem by using wider bandwidth channels, which are used by several users. Multiple users access the same channel by transmitting in their data in time slots. Thus, many low data rate users can be combined together to transmit in a single channel which has a sufficient bandwidth so that the spectrum can be used efficiently.

There are however, two main problems with TDMA. There is an overhead associated with the change over between users due to time slotting on the channel. This limits the number of users that can be sent efficiently in each channel. In addition, the symbol rate of each channel is high (as the channel handles the information from multiple users) resulting in problems with multipath delay spread.

OFDM overcomes most of the problems with both FDMA and TDMA. OFDM splits the available bandwidth into many narrow band channels (typically 100-8000 Hz). The carriers for each channel are made orthogonal to one another, allowing them to be spaced very close together, with no overhead as in the FDMA example. Because of this there is no great need for users to be time multiplexed as in TDMA, thus there is no overhead associated with switchizng between users.

The orthogonality of the carriers means that each carrier has an integer number of cycles over a symbol period. Due to this, the spectrum of each carrier has a null at the location of each of the other carriers in the system. This results in no interference between the carriers, allowing them to be as close as theoretically possible. This overcomes the problem of overhead carrier spacing required in FDMA. Each carrier in an OFDM signal has a very narrow bandwidth (i.e.1kHz), thus the resulting symbol rate is low. This results in the signal having a high tolerance to multipath delay spread, as the delay spread must be very long to cause significant inter-symbol interference (e.g. ≥ 500 μsec). We will discuss these aspects of OFDM system, in much detail, in Chapter 2.

1.5 Literature Survey

The first OFDM scheme was proposed by [7] in 1966 for dispersive fading channels, which has also undergone a dramatic evolution due to the efforts of [17]. Recently OFDM was selected as the high performance local area network transmission technique. A method to reduce the ISI is to increase the number of subcarriers by reducing the bandwidth of each subchannel while keeping the total bandwidth constant [18]. The ISI can instead be eliminated by adding a guard interval at the cost of power loss and bandwidth expansion [19]. These OFDM systems have been employed in military applications since the 1960's, for example by Bello [20], Zimmerman [21], Powers and Zimmerman [22], and others. The employment of discrete Fourier transform (DFT) to replace the banks of sinusoidal generators and the demodulators was suggested by Weinstein and Ebert [17] in 1971, which significantly reduces the implementational complexity of OFDM modems. Hirosaki [23], suggested an equalization algorithm in order to suppress both intersymbol and intersubcarrier interference caused by the channel impulse response or timing and frequency errors. Simplified model implementations were studied by Peled [24] in 1980. Cimini [25] and Kalet [26] published analytical and early seminal experimental results on the performance of OFDM modems in mobile communication channels.

Most recent advances in OFDM transmission were presented in the impressive state of art collection of works edited by Fazel and Fettweis [27]. OFDM transmission over mobile communications channels can alleviate the problem of multipath propagation [28]. Recent research efforts have focused on solving a set of inherent difficulties regarding OFDM, namely peak-to-mean power ratio, time and frequency synchronization, and on mitigating the effects of the frequency selective fading channels.

Channel estimation and equalization is an essential problem in OFDM system design. Basic task of equalizer is to compensate the influences of the channel [8]. This compensation requires, however, than an estimate of the channel response is available. Often the channel frequency response or impulse response is derived from training sequence or pilot symbols, but it is also possible to use nonpilot aided approaches like blind equalizer algorithms [29]. Channel estimation is one of the fundamental issue of OFDM system design, without it non coherent detection has to be used, which incurs performance loss of almost 3-4dB compared to coherent detection [30]. If coherent OFDM system is adopted, channel estimation becomes a requirement and usually pilot tones are used for channel estimation [31].

A popular class of coherent demodulation for a wide class of digital modulation schemes has been proposed by Moher and Lodge [10], and is known as *Pilot Symbol Assisted Modulation, PSAM*. The main idea of PSAM channel estimation is to multiplex known data streams with unknown data. Conventionally the receiver firstly obtain tentative channel estimates at the positions of the pilot symbols by means of remodulation and than compute final channel estimates by means of interpolation.

Aghamohammadi [32] et al. and Cavers [11] were among the first analyzing and optimizing PSAM given different interpolation filters. The main disadvantage of this scheme is the slight increase of the bandwidth. One class of such pilot symbol assisted estimation algorithms adopt an interpolation technique with fixed parameters (two dimensional [33]-[34] and one dimensional [35]) to estimate the frequency domain channel impulse response by using channel estimates obtained at the lattices assigned to the pilot tones. Linear, Spline and Gaussian filters have all been studied [35].

Channel estimation using superimposed pilot sequences [36] is also a completely new area, idea for using superimposed pilot sequences has been proposed by various authors for different applications [37]. In [38], superimposed pilot sequences are used for time and frequency synchronization. In [39], superimposed pilot sequences are introduced for the purpose of channel estimation, and main idea here is to linearly add a known pilot sequence to the transmitted data sequence and perform joint channel estimation and detection in the receiver. But the main problem in [39] is that the complexity of receiver is quite high and therefore low complexity approximations are of interest.

In [40], expectation maximization (EM) algorithm was proposed, and in [41] EM algorithm was applied on OFDM systems for efficient detection of transmitted data as well as for estimating the channel impulse response. Here, maximum likelihood estimate of channel was obtained by using channel statistics via the *EM algorithm*.

In [42], performance of low complexity estimators based on DFT has been analyzed. In [43], block and comb type pilot arrangements have been analyzed.

There are some other techniques, proposed for channel estimation and calculation of channel transfer function in OFDM systems. For example, the use of correlation based estimators working in the time domain [25] and channel estimation using singular value decomposition [44]. Its basically based on pilot symbols but in order to reduce its complexity, statistical properties of the channel are used in a different way. Basically the structure of OFDM allows a channel estimator to use both time and frequency correlations, but particularly it is too complex. In [44], they analyzed a class of block oriented channel estimators for OFDM, where only the frequency correlations of the channel is used in estimation. Whatever, their level of performance, they suggested that they may be improved with the addition of second filter using the time correlation [45]. In [46], they proposed a channel estimation algorithm based polynomial approximations of the channel parameters both in time and frequency domains. This method exploits both the time and frequency correlations of the channel parameters.

Use of the pilot symbols for channel estimation is basically an overhead of the system, and it is desirable to keep the number of pilot symbols to a minimum. In [47], Julia proposed a very good approach for OFDM symbol synchronization in which synchronization (correction of frequency offsets) is achieved simply by using pilot carriers already inserted for channel estimation, so no extra burden is added in the system for the correction of frequency offsets. Similarly in [48], it has been shown that the number of pilot symbols for a desired bit error rate and Doppler frequency is highly dependant on the pilot patterns used, so by choosing a suitable pilot pattern we can reduce the number of pilot symbols, but still retaining the same performance. Most common pilot patterns used in literature are block and comb pilot arrangements [43], [49]. Comb patterns perform much better than block patterns in fast varying environments [43].

1.6 Motivation

The focus of future fourth-generation (4G) mobile systems is on supporting high data rate services and ensuring seamless provisioning of services across a multitude of wireless systems and networks, for indoor to outdoor, from one interface to another, and from private to public network infrastructure [50]-[51].

Higher data rates allow the deployment of multi-media applications which involve voice, data, pictures, and video over the wireless networks. At this moment, the data rate envisioned for 4G networks is 1Gb/s for indoor and 100Mb/s for outdoor environments [52]. High data rate means the signal waveform is truly wideband, and the channel is frequency-selective from the waveform perspective, that is, a large number of resolvable multipaths are present in the environment. Orthogonal frequency division multiplexing (OFDM), which is a modulation technique for multicarrier communication systems, is a promising candidate for 4G systems since it is less susceptible to intersymbol interference introduced in the multipath environment [6].

It is not possible to make reliable data decisions unless a good channel estimate is available. Thus, an accurate and efficient channel estimation procedure is necessary to coherently demodulate the received data. As we mentioned earlier that although differential detection could be used to detect the transmitted signal in the absence of channel estimates, it would result in about 3-4dB loss [8] in signal to noise ratio compared to coherent detection. Moreover, as opposed to former standards using OFDM modulation, the new standards rely on QAM modulation and thus require channel estimation. Hence, the complexity of channel estimation is of crucial importance, especially for time varying channels, where it has to be performed periodically or even continuously. Several channel estimation techniques related with OFDM systems have been proposed in literature [53]. Number of pilot symbols for a desired error rate and Doppler frequency is highly dependent on how we transmit pilots [48] in OFDM systems. Rearrangement of pilot symbols, in some cases, can handle 10 times higher Doppler frequencies alternatively reduce the needed pilot symbols the same amount, still retaining the same bit error rate [48].

1.7 Objectives and Outline of Thesis

In this research work, channel estimation in OFDM systems is investigated. The main objective of this thesis is to investigate the performance of channel estimation in OFDM systems, study different patterns of pilot symbols which already have been proposed in literature, and then suggest a new scheme of transmitting pilots. We compare proposed pattern with the existing patterns, and discuss the usefulness of proposed scheme.

The main objectives of this thesis are: (1) Investigate the effectiveness of Orthogonal Frequency Division Multiplexing (OFDM) as a modulation technique for wireless radio applications. Main factors effecting the performance of a OFDM system are multipath delay spread and channel noise. The performance of OFDM is assessed using computer simulations performed using Matlab. It was found that OFDM performs extremely well, providing a very high tolerance to multipath delay spread and channel noise. (2) In pilot assisted channel estimation [48], we study different pilot arrangements, and investigate how to select a suitable pilot pattern for wireless OFDM transmission. We compare our proposed pattern of pilots with existing pilot patterns and discus the effectiveness of proposed scheme. (3) In wireless OFDM systems, effect of the channel is two fold, phase of the channel induces shifts in the frequencies of sub carriers, due to which we have high error rates in OFDM transmission. We propose a new method for the compensation of channel phase, and verify it by simulations, that method works well for slowly varying channels, and gives improvement in error rate.

This thesis is organized as follows: In Chapter 2, the basics of OFDM are presented. It is explained how an OFDM signal is formed using the inverse fast fourier transform, how the cyclic extension helps to mitigate the effects of multipath. Chapter 3 focuses on various aspects of multipath fading channel, what are the different attributes of multipath fading channel, and which model we use in our simulations. In Chapter 4, an overview of different approaches of channel estimation in OFDM systems is presented. We also discuss different channel estimation and interpolation techniques in Chapter 4. Chapter 5 demonstrates the simulation result of OFDM system employing a single antenna under AWGN and a 2-ray static multipath channel. Basic design rules are also given here how to choose the OFDM parameters, giving a required bandwidth, multipath delay spread and maximum Doppler spread. Chapter 6 describes the proposed pilot arrangement and compares it with the existing arrangements via simulations and results are discussed. We also describe a new scheme for adaptive channel estimation and for the compensation of frequency offsets in OFDM systems. Chapter 7 concludes the thesis and summarizes the results of the work. Areas for future work are also suggested.

Chapter 2

Introduction To OFDM

Orthogonal frequency division multiplexing (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 bandpass 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 others. 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 [54].

In the more conventional approach the traffic data is applied directly to the modulator with a carrier frequency at the center of the transmission band $f_0, ..., f_{N-1}$, ,i.e., at $(f_{N-1} + f_0)/2$, and the modulated signal occupies the entire bandwidth W. When the data is applied sequentially the effect of a deep fade in a mobile channel is to cause burst errors. Figure 2.1 shows the serial transmission of symbols $S_0, S_1, ..., S_{N-1}$, while the solid shaded block indicates the position of the error burst which affects only $k < N$ symbols.

By contrast, during the N-symbol period of the conventional serial system, each OFDM modulator carriers only one symbol, and the error burst causes sever signal degradation of the duration of k-serial symbols. This degradation is shown cross-hatched. However, if the error burst is only a small fraction of the symbol period than each of the OFDM symbols may only be slightly affected by the fade and they can still be correctly demodulated. Thus while the serial system exhibits an error burst, no errors or few errors may occur using the OFDM approach.

A further advantage of OFDM is that because the symbol period has been increased, the channel delay spread is significantly a shorter fraction of a symbol period than in the serial system, potentially rendering the system less sensitive to ISI than the conventional serial system.

A disadvantage of the OFDM approach, shown in Figure 2.2, is the increased complexity over the conventional system caused by employing N modulators and filters at the transmitter and N demodulators and filters at the receiver. However, this complexity can be removed by the use of the FFT and IFFT at the receiver and transmitter, respectively [17].

2.1 Brief History of OFDM

The concept of using parallel data transmission by means of frequency division multiplexing (FDM) was published in mid 60's by Chang [7], [27]. Some early developers can be traced back in the 50's a U.S. patent was filled and issued in January, 1970. The idea was to use parallel data streams and FDM with overlapping subchannels to avoid the use of high speed equalization, and to combat impulsive noise, and multipath distortion as well as to fully use the available bandwidth. The initial applications were in the military communications. In the telecommunications field, the term of discrete Multitone, multi-channel modulation and multi-carrier modulation (MCM) are widely used and sometimes they are interchangeable with OFDM. In OFDM, each carrier is orthogonal to all other carriers. However, this condition is not always maintained in MCM. OFDM is an optimum version of multi carrier transmission schemes [6].

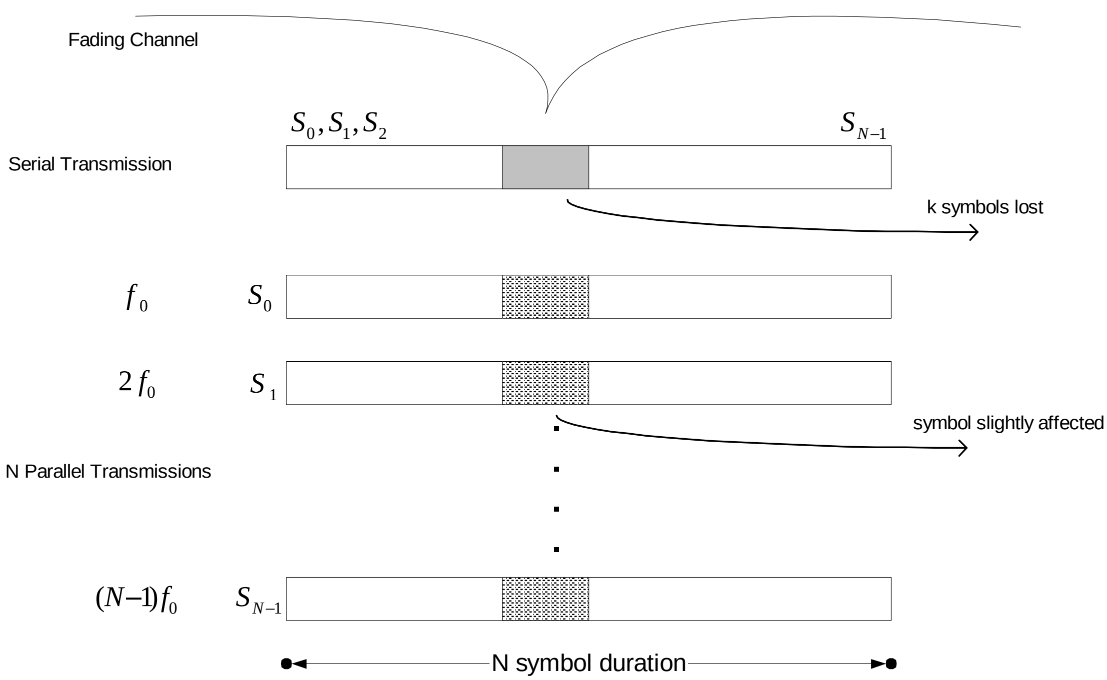

Figure 2.1: Effect of a fade on serial and parallel systems

For a large number of subchannels, the arrays of sinusoidal generators and coherent demodulator require in parallel system become unreasonably expensive and complex. The receiver needs precise phasing of the demodulator carriers and sampling times in order to keep crosstalk between subchannels acceptable.

Weinstein and Ebert [17] applied the discrete Fourier transform to parallel data as part of modulation and demodulation process. But the problem with FFT is that here we can use only limited frequencies, which are the integral multiples of $\frac{1}{T}$, where T is the symbol time period. In the 1980's, OFDM was studied for high-speed modems, digital mobile communications [24] and high density recording. One of the systems, used a pilot tone [55] for stabilizing carrier and clock frequency control and trellis coding was implemented. In 1990's, OFDM was exploited for wideband data communications over mobile radio FM channels, high bit-rate digital subscriber line, asymmetric digital subscriber line, very high speed digital subscriber lines, digital audio broadcasting (DAB) [56] and HDTV terrestrial broadcasting.

2.2 Generation of OFDM Symbols

A baseband OFDM symbol can be generated in the digital domain before modulating on 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). These data symbols are then converted to parallel streams before modulating subcarriers. Subcarriers are sampled with sampling rate $\frac{N}{T_s}$, where N is the number of subcarriers and T_s is the OFDM symbol duration. The frequency separation between two adjacent subcarriers is $\frac{2\pi}{N}$. Finally, samples on each subcarrier are summed together to form an OFDM sample. An OFDM symbol generated by an N-subcarrier OFDM system consists of N samples and the m-th sample of an OFDM symbol is given by [57].

$$x_m = \sum_{n=1}^{N-1} X_n e^{j\frac{2\pi mn}{N}} \qquad 0 \leq m \leq N-1 \tag{2.1}$$

where X_n is the transmitted data symbol on the nth carrier. Equation (2.1) is equivalent to the N-point inverse discrete Fourier transform (IDFT) operation on the data sequence with the omission of a scaling factor. It is well known [58] that IDFT can be implemented efficiently using inverse fast Fourier transform (IFFT). 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 sub channels available for transmission, but all of the channels needs to be active. The sub-channel bandwidth is given by

$$f_{sc} = \frac{1}{T_s} = \frac{f_{samp}}{N} \tag{2.2}$$

where f_{samp} is the sample rate and T_s is the symbol time.

Finally, a baseband OFDM symbol is modulated by a carrier to become a bandpass signal and transmitted to the receiver. In the frequency domain, this corresponds to translating all the subcarriers from baseband to the carrier frequency simultaneously. Figure 2.2 shows a 3-subcarrier OFDM transmitter and the process of generating one OFDM symbol.

2.3 Intersymbol and Intercarrier Interference

In a multipath environment, a transmitted symbol takes different times to reach the receiver through different propagation paths. From the receivers point of view, the channel introduces time dispersion in which the duration of the received symbol is stretched. Extending the symbol duration causes the current received symbol to overlap previous received symbols and results in intersymbol interference (ISI) [8]. 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 [28]. 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 intercarrier interference (ICI).

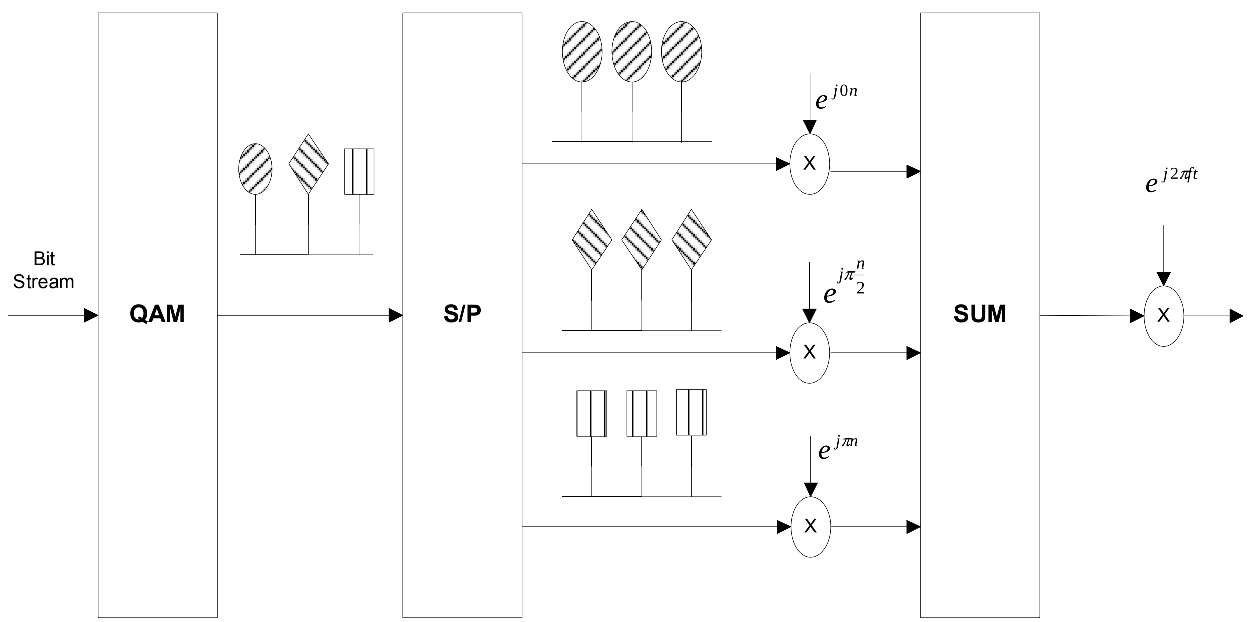

Figure 2.2: A 3 Sub-carrier OFDM Transmitter

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. 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. Figure 2.3 shows the spectra of four subcarriers in the frequency domain for the orthogonality case.

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. Figure 2.4 shows the spectra of four subcarriers in the frequency domain when orthogonality is lost.

ICI occurs when the multipath channel varies over one OFDM symbol time [59]. 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. This situation can be viewed from the time domain perspective, in which the integer number of cycles for each subcarrier within the FFT interval of the current symbol is no longer maintained due to the phase transition introduced by the previous symbol. Finally, any offset between the subcarrier frequencies of the transmitter and receiver also introduces ICI to an OFDM symbol.

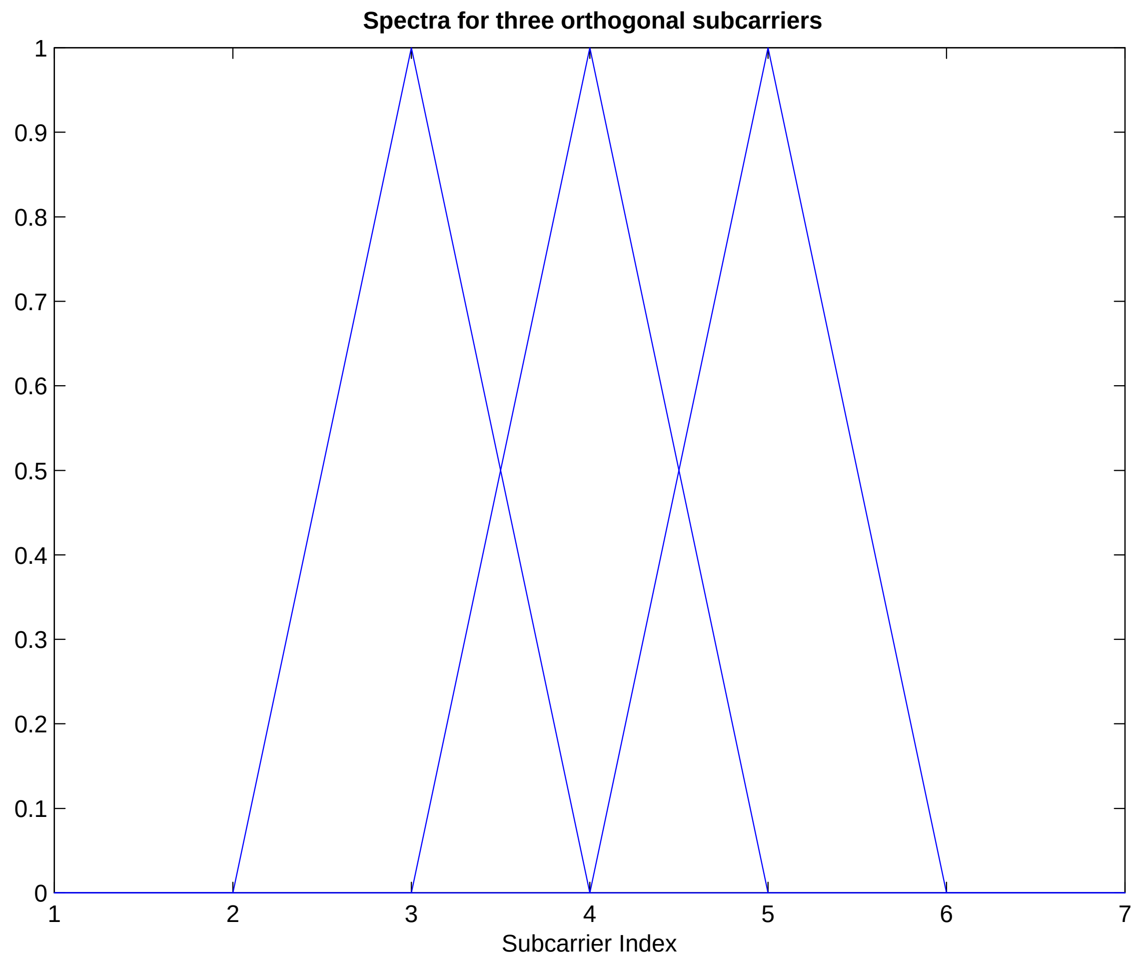

Figure 2.3: Spectra of three orthogonal subcarriers

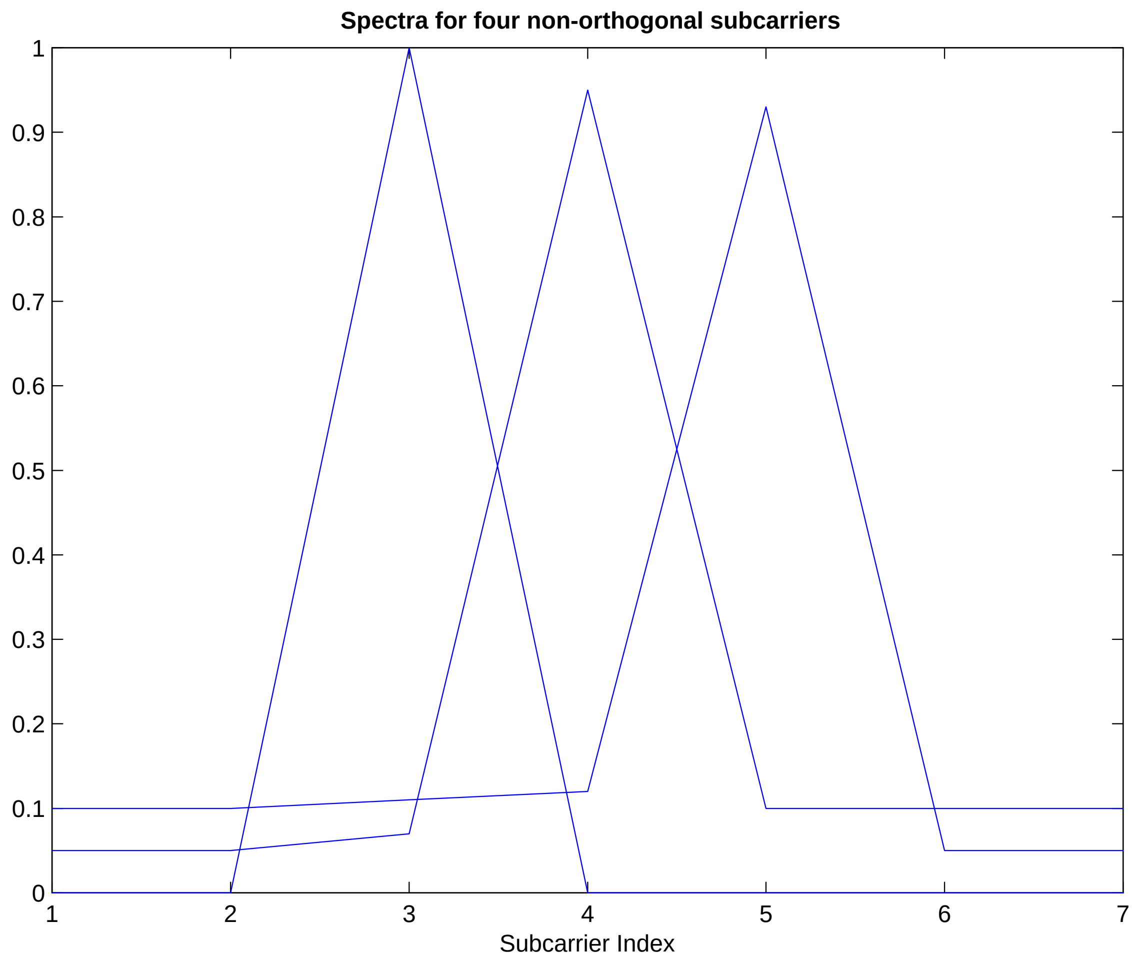

Figure 2.4: Spectra of three non-orthogonal subcarriers

2.4 Guard Time Insertion

OFDM is resilient to ISI because its symbol duration is long compared with the data symbols in the serial data stream. For an OFDM transmitter with N subcarriers, if the duration of a data symbol is $T^{'}$, the duration of the OFDM symbol at the output of the transmitter is

$$T_s = T^{'}N \tag{2.3}$$

Thus if the delay spread of a multipath channel is greater than $T^{'}$ but less then T_s, the data symbol in the serial data stream will experience frequency-selective fading while the data symbol on each subcarrier will experience only flat-fading. Moreover, to further reduce the ISI, a guard time is inserted at the beginning of each OFDM symbol before transmission as shown in Figure 2.5, and removed at the receiver before the FFT operation.

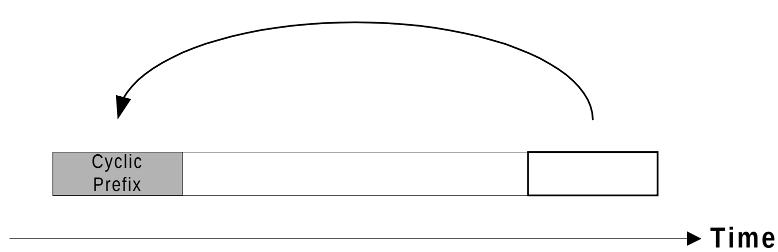

Figure 2.5: The cyclic prefix is a copy of the last part of OFDM symbol

If the guard time is chosen such that its duration is longer than the delay spread, the ISI can be completely eliminated. Figure 2.6 illustrates the concept of guard time insertion in an OFDM system. Figure 2.7 demonstrates the idea of eliminating ISI from OFDM symbols. In Figure 2.7(a), an OFDM symbol received is interfered from the previous OFDM symbol. On the other hand, Figure 2.7(b) 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. Guard time insertion can be introduced in many ways [60], but the most effective way of inserting guard period [57] is to extract a portion of an OFDM symbol at the end and append it to the beginning of the OFDM symbol. Samples after guard time can be expressed as

$$x_k^g = x_{(k+N-GI)_N}, \quad 0 \leq k \leq (N + GI - 1) \tag{2.4}$$

where k is the sample index of an OFDM symbol, N is the number of subcarriers, GI is the guard time duration, and $(k)_N$ is the residue modulo N.

2.5 Mathematical Model of OFDM System

The basic idea of OFDM is to divide the available spectrum into several subchannels (subcarriers) by making all subchannels narrowband, they experience almost flat fading, which makes equalization very simple or may not require equalization. To obtain a high spectral efficiency the frequency response of the subchannels are

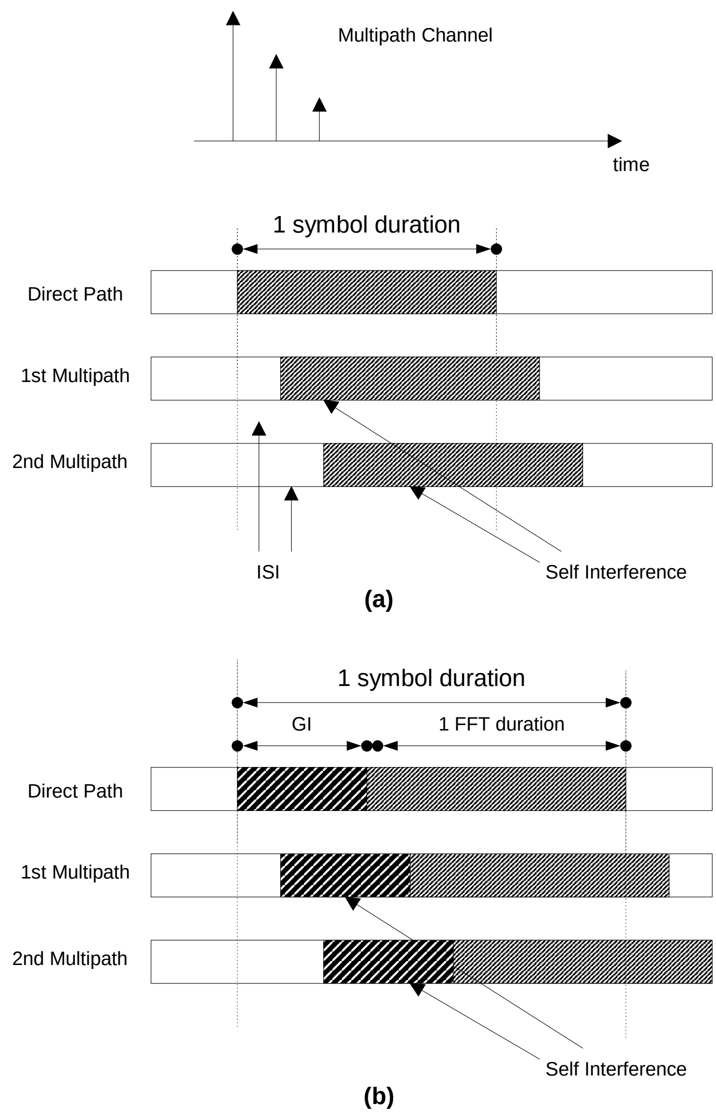

Figure 2.6: Received OFDM Symbol Components after passing through a multipath channel (a) without guard interval (b) with guard interval

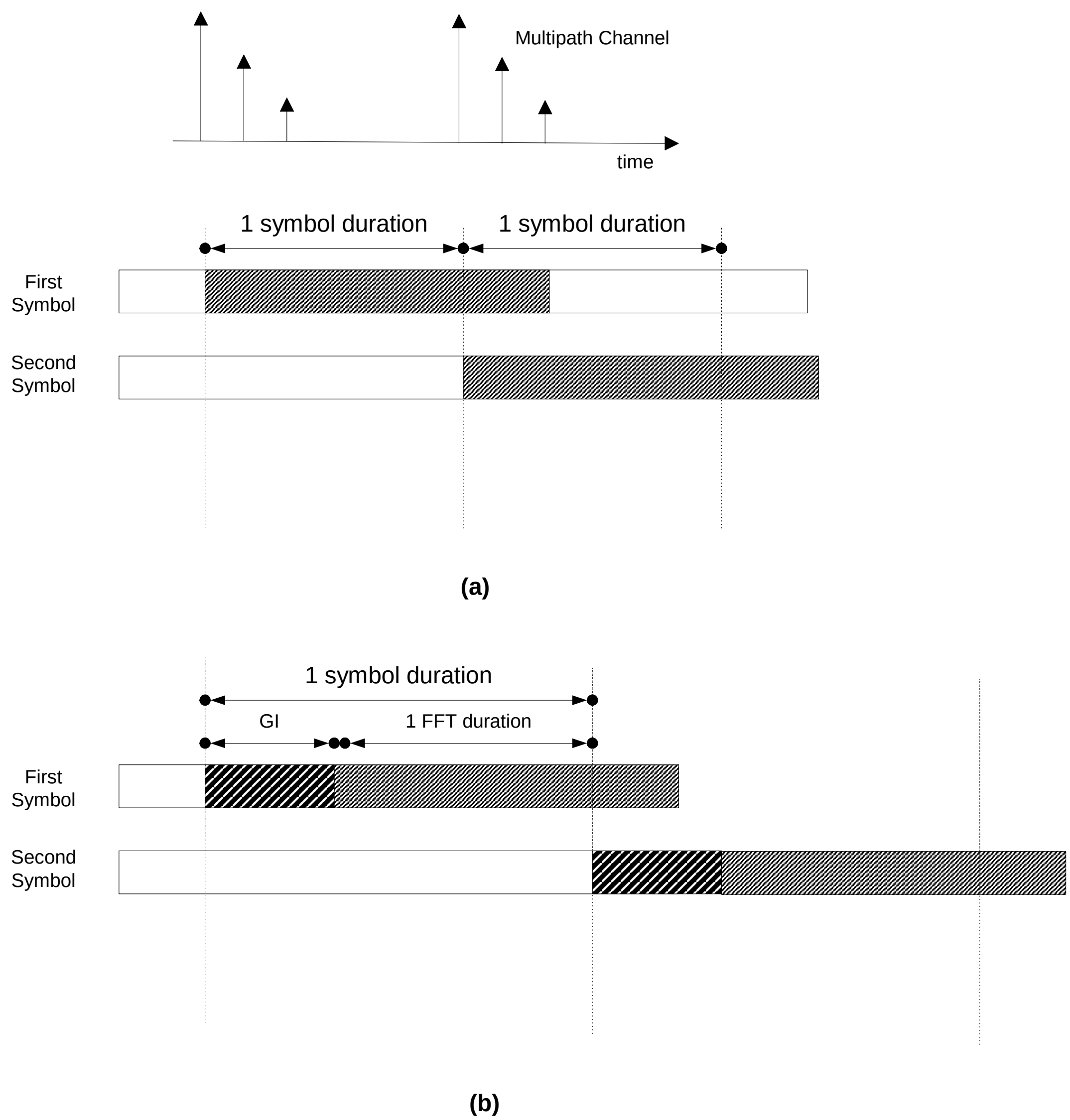

Figure 2.7: Received OFDM Symbols after passing through a multipath channel (a) without guard interval (b) with guard interval

overlapping and orthogonal, hence the name OFDM. This orthogonality can be completely maintained, even though the signal passes through a time dispersive channel, by introducing a cyclic prefix [54]. There are several versions of OFDM, see e.g., [61]-[62]-[17] but we focus on systems using such a cyclic prefix [24]. A cyclic prefix is a copy of the last part of the OFDM symbol which is prepended to the transmitted symbol, see Figure 2.5. This makes the transmitted signal periodic, which plays a decisive roll in avoiding intersymbol and intercarrier interference [61]. This is explained later in this section, although the cyclic prefix introduces a loss in signal to noise ratio (SNR), it is usually a small price to pay to mitigate interference.

A schematic diagram of a baseband OFDM system is shown in Figure 2.8. For this system we employ the following assumptions:

1. A cyclic prefix is used.

2. The impulse response of the channel is shorter than the cyclic prefix.

3. Transmitter and Receiver are perfectly time synchronized.

4. The fading is slow enough for the channel to be considered constant during one OFDM symbol interval.

In order to go thorough analysis of the system, we make some assumptions and develop a system model. Therefore, it is common practice to use simplified models resulting in a tractable analysis. We classify these OFDM system models into two different classes, continuous time and discrete time.

2.5.1 Continuous Time Model

The earlier OFDM systems did not employ digital modulation and demodulation. Hence, the continuous time OFDM model presented below can be considered as the ideal OFDM system, which in practice is digitally synthesized. Since this is the first model described, we move through it in a step by step fashion. We start with the waveforms used in the transmitter and proceed all the way to the receiver.

- **Transmitter**

  Assuming an OFDM system with N subcarriers, a bandwidth of W Hz and symbol length of T seconds, of which T_{cp} seconds is the length of the cyclic prefix, the transmitter uses the following waveforms:

$$\phi_k(t) = \begin{cases} \frac{1}{\sqrt{T-T_{cp}}} e^{j2\pi \frac{w}{N} k(t-T_{cp})} & \text{if t } \epsilon \text{ [0, T]} \\ 0 & \text{otherwise} \end{cases} \tag{2.5}$$

  where $T = \frac{N}{W} + T_{cp}$. Note that $\phi_k(t) = \phi_k(t + \frac{N}{W})$, where T is within the cyclic prefix $[0, T_{cp}]$.

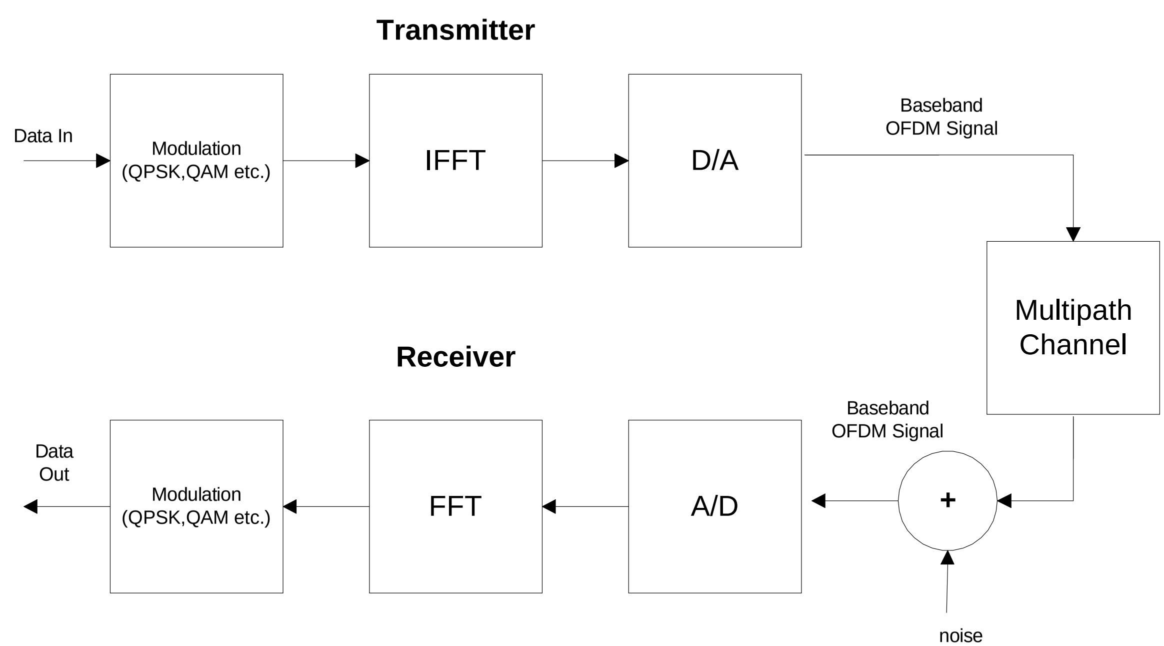

Figure 2.8: Basic FFT OFDM Transmitter and Receiver

Since $\phi_k(t)$ is the rectangular pulse modulated on the carrier frequency $\frac{kW}{N}$, the common interpretation of OFDM is that it uses N subcarriers, each carrying a low bit-rate. Now the transmitted baseband signal for OFDM symbol number l is

$$s_\ell(t) = \sum_{k=0}^{N-1} x_{k,\ell}\phi_k(t - \ell T) \tag{2.6}$$

where $x_{0,\ell}, x_{1,\ell}, ..., x_{N-1,\ell}$ are complex symbols form a set of signal constellation points. When an infinite sequence of OFDM symbols is transmitted, the output from the transmitter is a juxtaposition of individual OFDM symbols:

$$s(t) = \sum_{\ell=-\infty}^{\infty} s_\ell(t) = \sum_{\ell=-\infty}^{+\infty}\sum_{k=0}^{N-1} x_{k,\ell}\phi_k(t - \ell T) \tag{2.7}$$

- **Physical Channel** We assume that the support of the (possibly time variant)impulse response $g(\tau; t)$ of the physical channel is restricted to the interval $\tau \in [0, T_{cp}]$ i.e. to the length of the cyclic prefix. The received signal becomes

$$r(t) = g(\tau, t) * s(t) + n(t) = \int_0^{T_{cp}} g(\tau; t)s(t - \tau)d\tau + \tilde{n}(t) \tag{2.8}$$

where $\tilde{n}(t)$ is additive white, complex Gaussian channel noise.

- **Receiver** The OFDM receiver consists of a filter bank, matched to the last part $[T_{cp}, T]$ of the transmitter waveforms $\phi_k(t)$, i.e.,

$$\psi_k(t) = \begin{cases} \phi^*(T - t) & \text{if t } \epsilon \text{ } [0, \mathrm{T} - \mathrm{T_{cp}}] \\ 0 & \text{otherwise} \end{cases} \tag{2.9}$$

Effectively, this means, that the cyclic prefix is removed in the receiver. Since the cyclic prefix contains all ISI from the previous symbol, the sampled output from the receiver filter bank contains no ISI. Hence we can ignore the time index ℓ when calculating the sampled output at the kth matched filter. By using Equation (2.7), (2.8) and (2.9), we get

$$\begin{aligned} y_k \;=\; & \int_{T_{cp}}^{T} \Big(\int_{0}^{T_{cp}} g(\tau;t) \, \big[\sum_{k'=0}^{N-1} x_{k'} \phi_{k'}(t-\tau) \big] \, d\tau \Big) \, \phi_k^*(t) \, dt \; + \\ & \int_{T_{cp}}^{T} \tilde{n}(T-t) \, \phi_k^*(t) \, dt \end{aligned} \tag{2.10}$$

We consider the channel to be fixed over the OFDM symbol interval and denote it by $g(\tau)$, which gives

$$\begin{aligned} y_k \;=\; & \sum_{k'=0}^{N-1} x_{k'} \int_{T_{cp}}^{T} \Big(\int_{0}^{T_{cp}} g(\tau) \, \phi_{k'}(t-\tau) \, d\tau \Big) \, \phi_k^*(t) \, dt \; + \\ & \int_{T_{cp}}^{T} \tilde{n}(T-t) \, \phi_k^*(t) \, dt \end{aligned} \tag{2.11}$$

The integration intervals are $T_{cp} < t < T$ and $0 < \tau < T_{cp}$ which implies that $0 < t - \tau < T$ and the inner integral can be written as

$$\int_{0}^{T_{cp}} g(\tau) \, \phi_{k'}(t-\tau) \, d\tau = \int_{0}^{T_{cp}} g(\tau) \frac{e^{j2\pi k'(t-\tau-T_{cp})\frac{W}{N}}}{\sqrt{T-T_{cp}}} \, d\tau$$

$$= \frac{e^{j2\pi k'(t-T_{cp})\frac{W}{N}}}{\sqrt{T-T_{cp}}} \int_{0}^{T_{cp}} g(\tau) e^{-j2\pi k'\tau\frac{W}{N}} \, d\tau, \quad T_{cp} < t < T \tag{2.12}$$

The latter part of this expression is the sampled frequency response of the channel at frequency $f = k^{'}\frac{W}{N}$, i.e., at the $k^{'}$th subcarrier frequency:

$$h_{k^{'}} = G(k^{'}\frac{W}{N}) = \int_{0}^{T_{cp}} g(\tau)e^{-j2\pi k^{'}\tau\frac{W}{N}}\ d\tau, \tag{2.13}$$

where $G(f)$ is the Fourier transform of $g(\tau)$. Using this notation the output from the receiver filter bank can be simplified to

$$\begin{aligned} y_k \ &= \ \sum_{k^{'}=0}^{N-1} x_{k^{'}} \int_{T_{cp}}^{T} \frac{e^{j2\pi k^{'}(t-T_{cp})\frac{W}{N}}}{\sqrt{T-T_{cp}}} h_{k^{'}}\ \phi_k^*(t)\ dt\ +\ \int_{T_{cp}}^{T} \tilde{n}(T-t)\ \phi_k^*(t)\ dt \\ &= \sum_{k^{'}=0}^{N-1} x_{k^{'}} h_{k^{'}} \int_{T_{cp}}^{T} \phi_{k^{'}}(t)\ \phi_k^*(t)\ dt + n_k, \end{aligned} \tag{2.14}$$

where

$$n_k = \int_{T_{cp}}^{T} \tilde{n}(T-t)\ \phi_k^*(t)\ dt$$

Since the transmitter filters $\phi_k(t)$ are orthogonal,

$$\begin{aligned} \int_{T_{cp}}^{T} \phi_{k^{'}}(t)\ \phi_k^*(t)\ dt \ &= \ \int_{T_{cp}}^{T} \frac{e^{j2\pi k^{'}(t-T_{cp})\frac{W}{N}}}{\sqrt{T-T_{cp}}}\ \frac{e^{-j2\pi k(t-T_{cp})\frac{W}{N}}}{\sqrt{T-T_{cp}}}\ dt \\ &= \ \delta[k-k^{'}], \end{aligned} \tag{2.15}$$

where $\delta[k]$ is the Kronecker delta function [58], we can simplify Equation (2.14) and obtain

$$y_k = h_k x_k + n_k, \tag{2.16}$$

where n_k is the additive white Gaussian noise (AWGN).

The benefit of cyclic prefix is twofold: it avoids both ISI (since it acts as a guard space) and ICI (since it maintains the orthogonality of the subcarriers). By re-introducing the time index ℓ, we may now view the OFDM system as a set of parallel Gaussian channels, according to Figure 2.9.

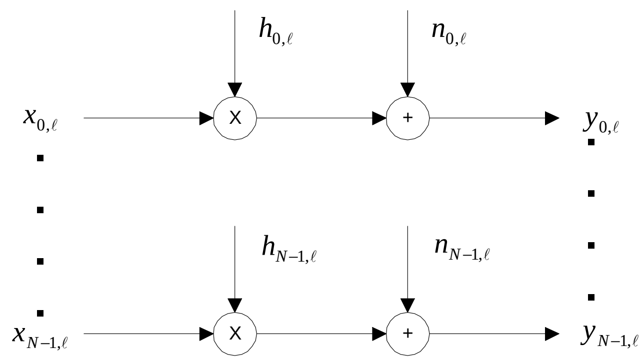

Figure 2.9: The continuous time OFDM system interpreted as parallel Gaussian channels.

2.5.2 Discrete-time model

In discrete time model, as compared to continues time model, the modulation and demodulation are replaced by an inverse DFT (IDFT) and a DFT, respectively, and the channel is a discrete time convolution. The cyclic prefix operates in the same fashion and the calculations can be performed in essentially the same way. The main difference is that all integrals are replaced by sums.

From the receiver's point of view, the use of a cyclic prefix longer than the channel will transform the linear convolution in the channel to a cyclic convolution.

Denoting cyclic convolution by $'\otimes'$, we can write the whole OFDM system as:

$$y_l = DFT(IDFT(x_l) \otimes g_l + \tilde{n}_l)$$

$$= DFT(IDFT(x_l) \otimes g_l) + n_l, \tag{2.17}$$

where y_l contains N received data points, s_l the N transmitted constellation points, g_l the impulse response of the channel (padded with zero to obtain a length of N), and $\tilde{n}_l$ the channel noise. Since the channel noise is assumed white and Gaussian, the term $n_l = DFT(\tilde{n}_l)$ represents the uncorrleated Gaussian noise. Further, we use that the DFT of two cyclically convolved signals is equivalent to the product of their individual DFTs. Denoting element-by-element multiplication by (.), the above expression can be written

$$y_l = x_l.DFT(g_l) + n_l = x_l.h_l + n_l, \tag{2.18}$$

where $h_l = DFT(g_l)$ is the frequency response of the channel. Thus we have obtained the same type of parallel Gaussian channels as for the continuous-time model. The only difference is the channel attenuations h_l are given by the N point DFT of the discrete time channel, instead of the sampled frequency response as in Equation (2.15).

2.6 Equalization and Channel Estimation

2.6.1 Equalization

Although the guard time which has longer duration than the delay spread of a multipath channel can eliminate ISI because of the previous symbol, but it is still have some ISI because of the frequency selectivity of the channel. In order to compensate this distortion, a one-tap channel equalizer is needed. At the output of FFT on the receiver side, the sample at each subcarrier is multiplied by the coefficient of the corresponding channel equalizer. The coefficient of an equalizer can be calculated based on the zero-forcing (ZF) criterion or the minimum mean-square error (MMSE) criterion [63]. The ZF criterion forces ISI to be zero at the sampling instant of each subcarrier. The coefficient of a one-tap ZF equalizer is calculated as follows:

$$C_n = \frac{1}{H_n} \tag{2.19}$$

where H_n is the channel frequency response within the bandwidth of the n-th subcarrier. The disadvantage of ZF criterion is that it enhances noise at the n-th subcarrier if H_n is small, which corresponds to spectral nulls [64].

2.6.2 Channel Estimation

Equation (2.19) showed that one needs to perform channel estimation in order to obtain weights for equalizers on individual subcarriers. Training symbols known as Pilot Symbols, are often used to perform channel estimation [48]. In OFDM, since equalization is performed in the frequency domain, it is the channel frequency response that must be estimated. In the multipath environment, the demodulated symbol X_n on the n-th sub-carrier at the output of FFT without ISI and ICI can be represented in Equation (2.20).

$$Y_n = [\sum_{l=0}^{GI-1} H^l(0)e^{-j\frac{2\pi nl}{N}}]X_n + N_n, \tag{2.20}$$

where GI is the number of multipath components, N_n is the FFT of the additive white Gaussian noise (AWGN) on the n-th subcarrier and $H_l(0)$ is the channel frequency response of the l-th multipath component at the zero-th frequency. To estimate the channel frequency response, pilot symbols are inserted to the subcarriers in the frequency domain, i.e., they are inserted before IFFT operation at the transmitter side. Let H_n be the channel frequency response experienced by X_n, i.e.,:

$$Y_n = \sum_{l=0}^{GI-1} H^l(0)e^{-j\frac{2\pi nl}{N}} \tag{2.21}$$

The channel frequency response experienced by the pilot symbol P_n on the n-th subcarrier can be estimated as:

$$\hat{H}_n = H_n + \frac{N_n}{P_n} \tag{2.22}$$

Since pilot symbols usually occupy a small amount of bandwidth for spectral efficiency, interpolation across frequency is required to estimate the channel frequency response where pilot symbols are not located. The channel frequency response at the m-th subcarrier $\hat{H_m}$ can be interpolated linearly as [54]:

$$\hat{H_m} = [1 - \frac{m}{N}]\hat{H}_{p1} + \frac{m}{N}\hat{H}_{p2}, \quad p_1 \leq m \leq p_2, \tag{2.23}$$

where $\hat{H}_{p1}$ and $\hat{H}_{p2}$ are the channel frequency response estimated by the pilot symbols on the p_1-th and p_2-th subcarriers. Furthermore, if the multipath channel is time varying in nature, then interpolation across time may also require tracking the channel.

Chapter 3

Multipath Fading Channel

Radio channel is the link between the transmitter and the receiver that carries information bearing signal in the form of electromagnetic waves. The radio channel is commonly characterized by scatterers (local to the receiver) and reflectors (local to the transmitter). Small scale fading, or simply fading, is used to describe the rapid fluctuations of the amplitude of a radio signal over a short period of time or travel distance, so that large scale path loss effects may be ignored.

3.1 Propagation Characteristics of Mobile Radio Channels

In an ideal radio channel, the received signal would consist of only a single direct path signal, which would be a perfect reconstruction of the transmitted signal. However, in a real channel the signal is modified during transmission. The received signal consists of a combination of attenuated, reflected, refracted, and diffracted replicas of the transmitted signal. On top of all this, the channel adds noise to the signal and can cause a shift in the carrier frequency if either of the transmitter or receiver is moving (Doppler effect). Understanding of these effects on the signal is important because the performance of a radio system is dependent on the radio channel characteristics.

3.1.1 Attenuation

Attenuation is the drop in the signal power when transmitting from one point to another. It can be caused by the transmission path length, obstructions in the signal path, and multipath effects. Figure 3.1 shows some of the radio propagation effects that cause attenuation. Any objects which obstruct the line of sight of the signal from the transmitter to the receiver, can cause attenuation.

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 the 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

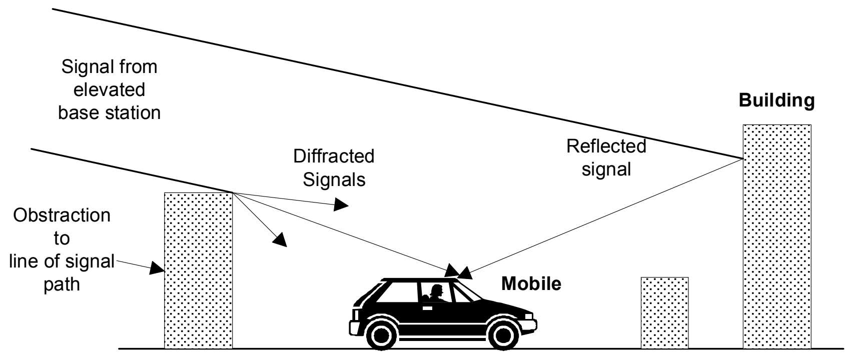

Figure 3.1: Radio Propagation Effects

diffraction is dependent on the radio frequency used, with high frequencies scatter more than low frequency signals. Thus high frequency signals, especially, Ultra High Frequencies (UHF) and microwave signals require line of sight for adequate signal strength, because these scatter too much. To over come the problem of shadowing, transmitters are usually elevated as high as possible to minimise the number of obstructions. Typical amounts of variation in attenuation due to shadowing are shown in Table 3.1

Shadowed areas tend to be large, resulting in the rate of change of the signal

Description	**Typical Attenuation due to shadowing**
Heavily built-up urban center	20dB variation from street to street
Sub-urban area (fewer large buildings)	10dB greater signal power than built-up urban center
Open rural area	20dB greater signal power than sub-urban areas
Terrain irregularities and tree foliage	3-12dB signal power variation

Table 3.1: Typical Attenuation in a radio channel (values from [1])

power being slow. For this reason, it is named slow-fading, or log-normal shadowing.

3.1.2 Multipath Effects

Rayleigh Fading

In a radio link, the RF signal from the transmitter may be reflected from objects such as hills, buildings, or vehicles. This gives rise to multiple transmission paths at the receiver. Figure 3.2 shows some of the possible ways in which multipath signals can occur.

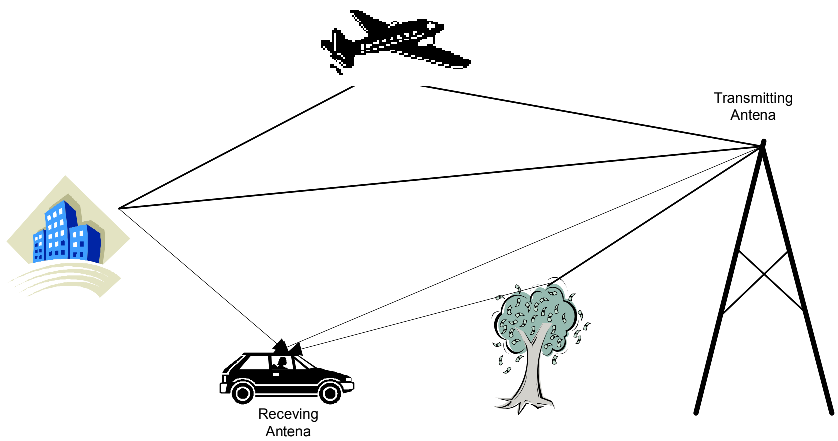

Figure 3.2: Multipath Reception

The relative phase of multiple reflected signals can cause constructive or destructive interference at the receiver. This is experienced over very short distances (typically at half wavelength distances), which is given the term fast fading. These variations can vary from 10-30dB over a short distance. The Rayleigh distribution is commonly used to describe the statistical time varying nature of the received signal power. It describes the probability of the signal level being received due to fading. Table 3.2 shows the percentage probability of the signal level being less than the value given in column 1 of Table 3.2, for the Rayleigh distribution.

Signal Level(dB)	**Percentage Probability**
10	99
0	50
-10	5
-20	0.5
-30	0.05

Table 3.2: Cummulative distribution for Rayleigh distribution (values from [2])

Frequency Selective Fading

In any radio transmission, the channel spectral response is not flat. It has dips or fades in the response due to reflections causing cancellation of certain frequencies at the receiver. Reflections off near-by objects (e.g. ground, buildings, trees, etc) can lead to multipath signals of similar signal power as the direct signal. This can result in deep nulls in the received signal power due to destructive interference.

For narrow bandwidth transmissions if the null in the frequency response occurs at the transmission frequency then the entire signal can be lost. This can be partly overcome in two ways. By transmitting a wide bandwidth signal or spread spectrum as in the case of CDMA, any dips in the spectrum only result in a small loss of signal power, rather than a complete loss. Another method is to split the transmission up into many carriers carrying low rate data, as is done in a COFDM/OFDM transmission. In CDMA, the original signal is spread over a wide bandwidth thus, any nulls in the spectrum are unlikely to occur at all of the carrier frequencies. This will result in only some of the carriers being lost, rather then the entire signal. The information in the lost carriers can be recovered provided enough forward error corrections is sent.

Delay Spread

The received radio signal from a transmitter consists of typically a direct signal plus signals reflected off object such as buildings, mountains, and other structures. The reflected signals arrive at a later time than the direct signal because of the extra path length, giving rise to a slightly different arrival time of the transmitted pulse. The signal energy confined to a narrow pulse is spreading over a longer time. Delay spread is a measure of how the signal power is spread over the time between the arrival of the first and last multipath signal seen by the receiver.

In a digital system, the delay spread can lead to inter-symbol interference. This is due to the delayed multipath signal overlapping symbols that follows. This can cause significant errors in high bit rate systems, especially when using time division multiplexing (TDMA). As the transmitted bit rate is increased the amount of inter-symbol interference also increases. The effect starts to become very significant when the delay spread is greater then 50% of the bit time.

Table 3.3 shows the typical delay spread that can occur in various environments.

The maximum delay spread in an outdoor environment is approximately 20μsec, thus significant intersymbol interference can occur at bit rates as low as 25kbps. In some mountainous areas, like salt lake city USA, delay spread of 100μsec has been measured.

Inter-symbol interference can be minimized in several ways. One method is to reduce the symbol rate by reducing the data rate for each channel (i.e. split the bandwidth into more channels using frequency division multiplexing). Another is to use a coding scheme which is tolerant to inter-symbol interference such as CDMA.

3.1.3 Doppler Shifts

When a signal source and/or a receiver are moving relative to one another, the frequency of the received signal will not be the same as the source. When they are moving away, the frequency of the received signal is lower then the source, and when they are approaching each other the frequency increases. This is called the Doppler effect. An example of this is the change of pitch in a cars horn as it approaches then passes by. This effect becomes important when developing mobile radio systems.

The amount the frequency changes due to the Doppler effect depends on the relative motion between the source and receiver and on the speed of propagation of

Environment	**Delay Spread**	**Maximum Path Length Difference**
Indoor	40$n$sec - 200$n$sec	12m - 60m
Outdoor	1μsec - 20μsec	300m - 6km

Table 3.3: Cummulative distribution for Rayleigh distribution (values from [2])

the wave. The Doppler shift in frequency can be written [65]:

$$\Delta f_d \approx \pm f_o \frac{v}{c} \cos\alpha \tag{3.1}$$

where Δf is the change in frequency of the source seen at the receiver, f_o is the frequency of the source, v is the speed difference between the source and transmitter, and c is the speed of light, and α is the angle between the line joining the transmitter and receiver at the direction of travel of mobile.

For example: Let $f_0 = 1GHz$ and $v = 60km/hr$, then the doppler shift will be

$$\Delta f_d = 10^9 . \frac{16.67}{3X10^8} = 55.5\ Hz$$

This shift of 55Hz in the carrier will generally not effect the transmission. However, Doppler shift can cause significant problems if the transmission technique is sensitive to carrier frequency offsets (for example in OFDM) or the relative speed is higher (for example in low earth orbiting satellites).

3.2 Modeling of Mobile Radio Channels

In last few decades, modeling and characterization of fading channels had gained considerable interest. Over many years, a large number of experiments have been carried out to investigate fading channels. Earlier work in this area includes the contributions of Bello [66], Clarke [67] and Jakes [68].

Assuming a low-pass equivalent model for the channel, the received signal $r(t)$

over a fading multipath channel can be represented by [8]

$$r(t) = \int_{-\infty}^{+\infty} h(\tau, t)s(t - \tau)d\tau \tag{3.2}$$

where $s(t)$ is the transmitted signal and $h(\tau, t)$ is the channel impulse response at delay τ and time instant t. In discrete form,

$$r(n) = \sum_{i=-\infty}^{\infty} h(iT_s, n)s(n - iT_s) \tag{3.3}$$

where T_s is the symbol duration and n represents the sampling index. Defining a compact notation for the time varying channel coefficients in the form,

$$h_i(n) = h(iT_s, n)$$

Equation (3.3) can be written as

$$r(n) = \sum_{i=-\infty}^{\infty} h_i(n)s(n - iT_s) \tag{3.4}$$

The form of received signal in Equation (3.4) suggests that the impulse response of fading multipath channel can be modeled as a tapped delay line filter, a finite impulse response filter, with tap spacing T_s and time varying coefficients $h_i(n)$. The time varying coefficients are characterized as random processes because of the constantly changing physical characteristics of the channel. The tap weights, $h_i(n)$, in Figure 3.3 can be expressed as

$$h_i(n) = \sqrt{\rho_i}G_i(n) \tag{3.5}$$

where ρ_i is the strength of the ith path and $G_i(n)$ is the complex stochastic process specified by its mean square value and power spectrum density.

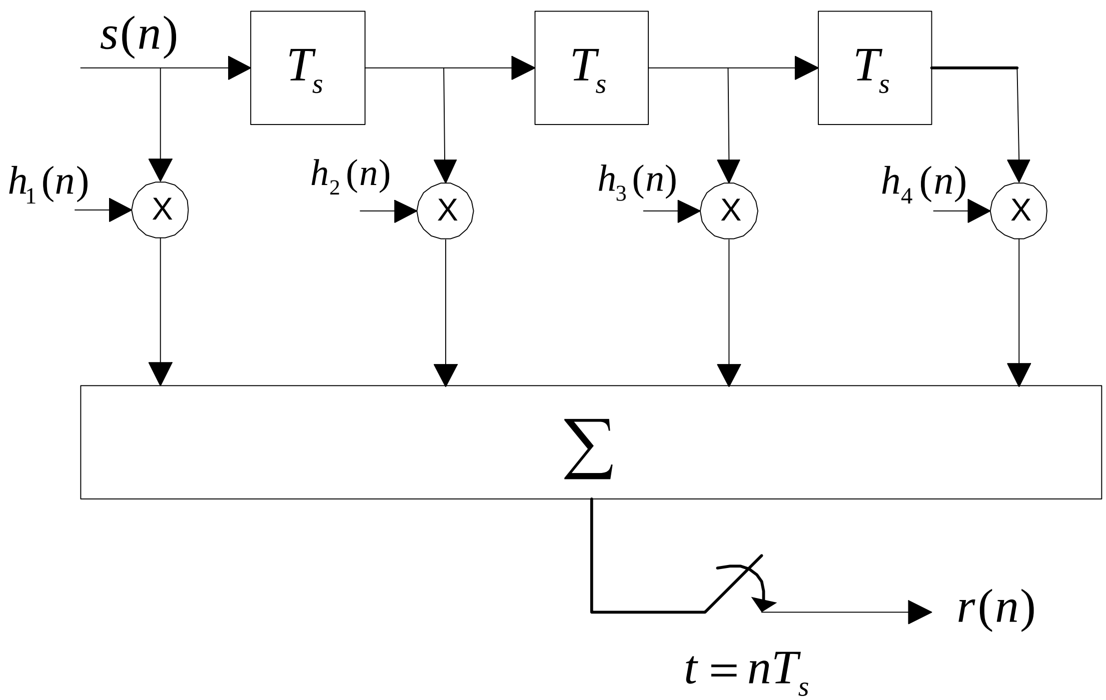

Figure 3.3: Tapped delay line model of fading channel with four taps

3.2.1 Fading Statistics

Fading describes the rapid fluctuations of the amplitude of radio signal when passes through radio channels, over a short period of time. The complex stochastic process $G_i(n)$ in Equation (3.5) represents the fading and can be completely characterized by specifying the pdf of its amplitude, phase and the autocoorelation function.

The simplest process that can exhibit time-selective and frequency-selective fading is *wide-sense stationary uncorrelated scattering (WSSUS)* process introduced by Bello [66]. The number of uncorrelated paths is sufficiently large so that quadrature components of the fading process are Gaussian distributed according to the central limit theorem [2]. In the absence of direct path, the Gaussian process has zero mean and the pdf of the envelope is Rayleigh [68] given by

$$f_G(g) = \frac{g}{\sigma^2} e^{-g^2/2\sigma^2}, \qquad g \geq 0 \tag{3.6}$$

where $\sigma^2 = E[GG^*]$ is the variance of Gaussian process. The phase pdf has uniform distribution [68],

$$f_\theta(\theta) = \frac{1}{2\pi}, \qquad 0 \leq \theta \leq 2\pi \tag{3.7}$$

A typical and often assumed shape for the power spectral density, also known as Doppler Spectrum, of the fading process for mobile radio channels is given by Jakes Spectrum [68],

$$S(f) = \begin{cases} \frac{\sigma^2}{\pi f_m \sqrt{1-(f/f_m)^2}} & |\mathrm{f}| \leq \mathrm{f_m} \\ 0 & \text{elsewhere} \end{cases} \tag{3.8}$$

where f_m is the maximum doppler frequency shift found from Equation (3.1). The autocorrelation function of the fading process is given by

$$\Re(\tau) = \sigma^2 J_0(2\pi f_m \tau) \tag{3.9}$$

where J_0 is the first order Bessel's function.

3.2.2 Propagation Delay Profile

To specify the tapped delay line model of fading channel completely, the relative strength of the multi-paths, ρ_i is also required as mentioned in Equation (3.5). The relative strength of paths is usually represented by the propagation delay profile of the channel which is dependent on environment and is usually determined by experiments and measurements. In this thesis, we use exponential propagation delay profile.

3.3 Realization of Rayleigh fading in simulations

In this section, we describe the simulator that will be used to simulate the channel. Basically, we need colored Gaussian noise to realize fading channel statistics which may be Rayleigh, Rician or any other. These colored Gaussian processes can be generated either by filtering white Gaussian noise [67] or by deterministic methods [68] or by Monte Carlo approach [69].

Filtering method requires large number of filter taps [70] to reshape the spectrum and it is based on Clarkes model. A deterministic method to simulate mobile fading channels is based on Rice's sum of sinusoids [68]. In this case, a colored Gaussian noise is approximated by a finite sum of sinusoids with proper weights and frequencies.

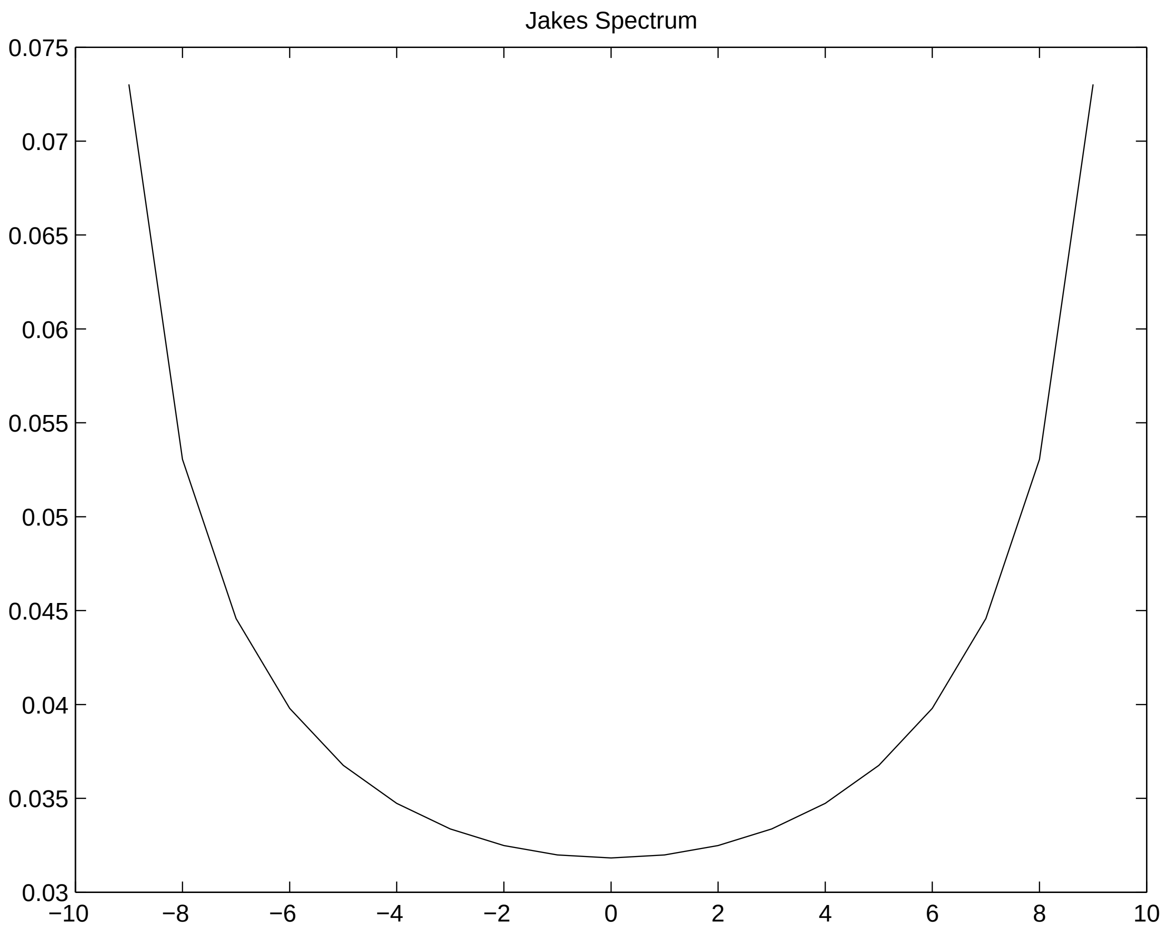

Figure 3.4: Doppler Spectrum

Jakes also presented a realization for the simulation of fading channel model which generates real and imaginary parts of the channel taps coefficients as weighted sum of sinusoids. Jakes simulator has been widely used and extensively studied over the past three decades [71], [72]. Recently, Pop and Beaulieu [73] have highlighted few shortcomings in the Jakes model. They propose to include a random phase in the low frequency oscillators of the Jakes model. In this thesis, we have implemented this modified Jakes model as fading channel simulator.

The real and imaginary parts of the channel taps are generated as [68]

$$g_I(t) = 2\sum_{n=1}^{N_0} \cos\beta_n \cos(\omega_n t + \phi_n) + \sqrt{2}\cos\alpha\cos(\omega_m t) \tag{3.10}$$

$$g_Q(t) = 2\sum_{n=1}^{N_0} \sin\beta_n \cos(\omega_n t + \phi_n) + \sqrt{2}\sin\alpha\cos(\omega_m t) \tag{3.11}$$

where

$$\beta_n = \frac{n\pi}{N_0 + 1} \qquad \omega_n = \omega_m \cos(\frac{2\pi n}{N}) \qquad N_0 = \frac{1}{2}(\frac{N}{2} - 1)$$

where $t = kT_s$ and $\phi_1, ..., \phi_{N_0}$ are uniformly distributed random variables over $[0, 2\pi]$. We have implemented the technique proposed by Jakes [68] and modify it according to [73]. In this technique, the nth oscillator is given an additional phase shift $\gamma_{nl} + \beta_n$ while retaining gains. For lth path the in-phase component of the fading can be written as:

$$g_\ell(t) = 2\sum_{n=1}^{N_0} \cos\beta_n \cos(\omega_n t + \phi_n + \theta_{nl}) + \sqrt{2}\cos\alpha\cos(\omega_m t) \tag{3.12}$$

where,

$$\theta_{nl} = \gamma_{nl} + \beta_n \qquad \beta_n = \frac{n\pi}{N_0 + 1} \qquad \gamma_{nl} = \frac{2\pi(\ell - 1)n}{N_0 + 1}$$

The quadrature component can be modified in the same manner. The normalized complex channel tap for lth path is

$$G_\ell(t) = g_\ell^I(t) + jg_\ell^Q(t) \tag{3.13}$$

which will be normalized such that $E[G_\ell G_\ell^*] = 1$. The Rayleigh distribution of channel taps is shown in Figure 3.5. At lower values of doppler frequencies, taps of channel varies slowly as shown in Figure 3.6, while at higher doppler frequencies taps of the channel changing fastly, implying the fact that mobile terminal is now moving with a higher speed and channel changes accordingly, as shown in Figure 3.7.

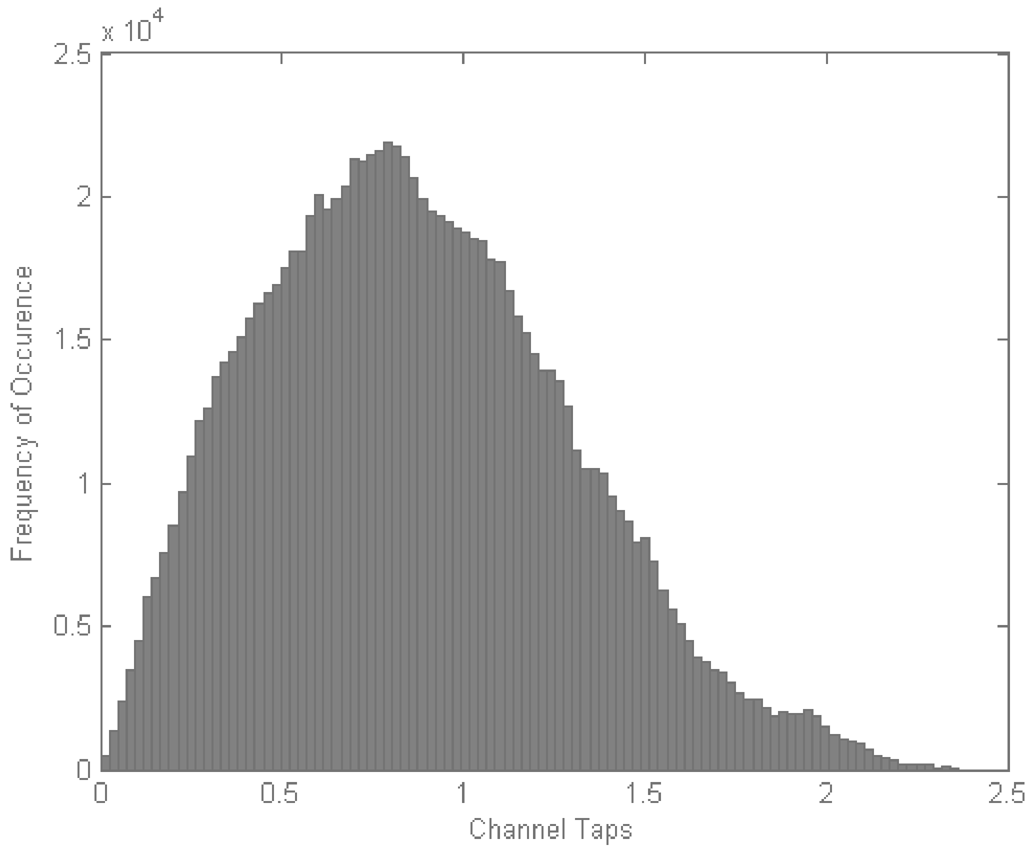

Figure 3.5: Channel Taps Distribution

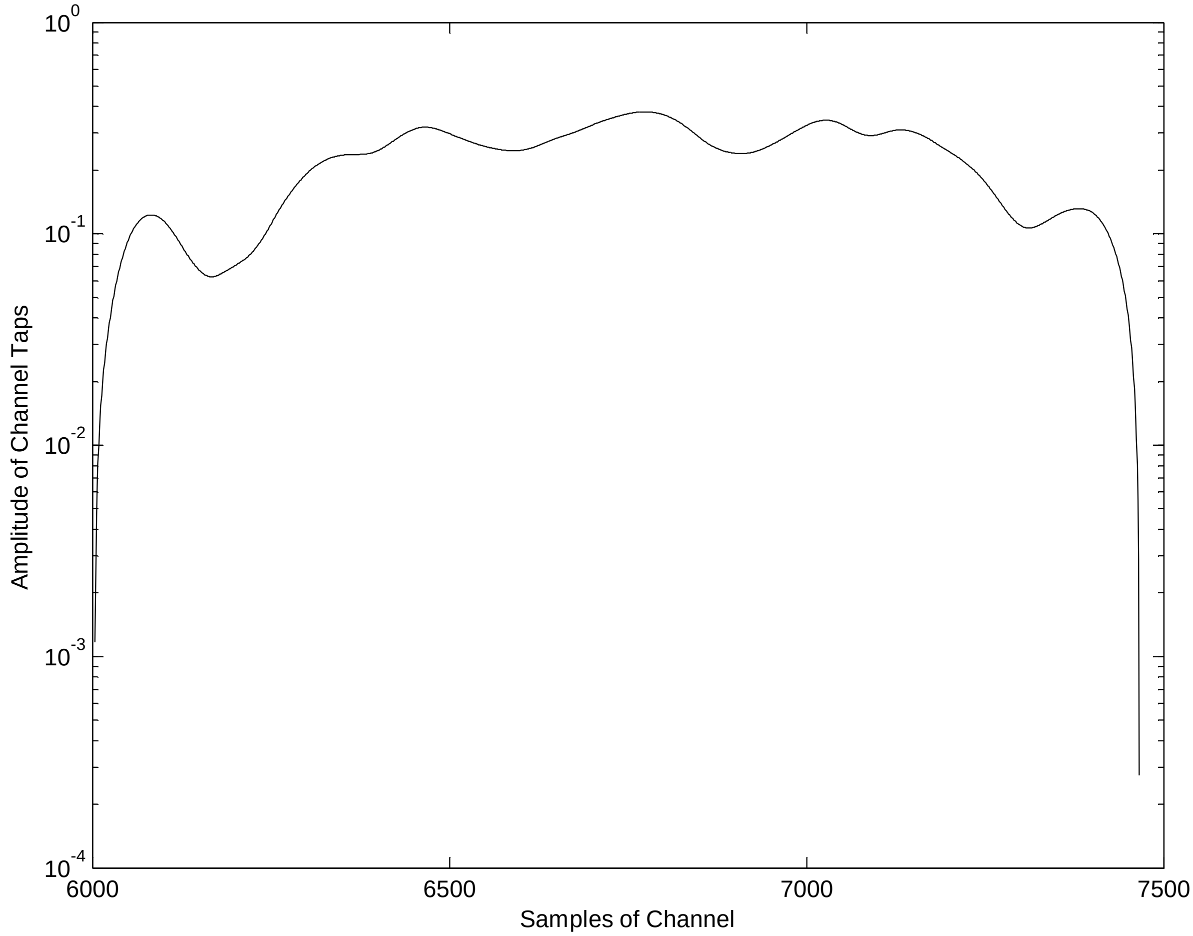

Figure 3.6: Fading Envelope at $f_d = 10$ Hz

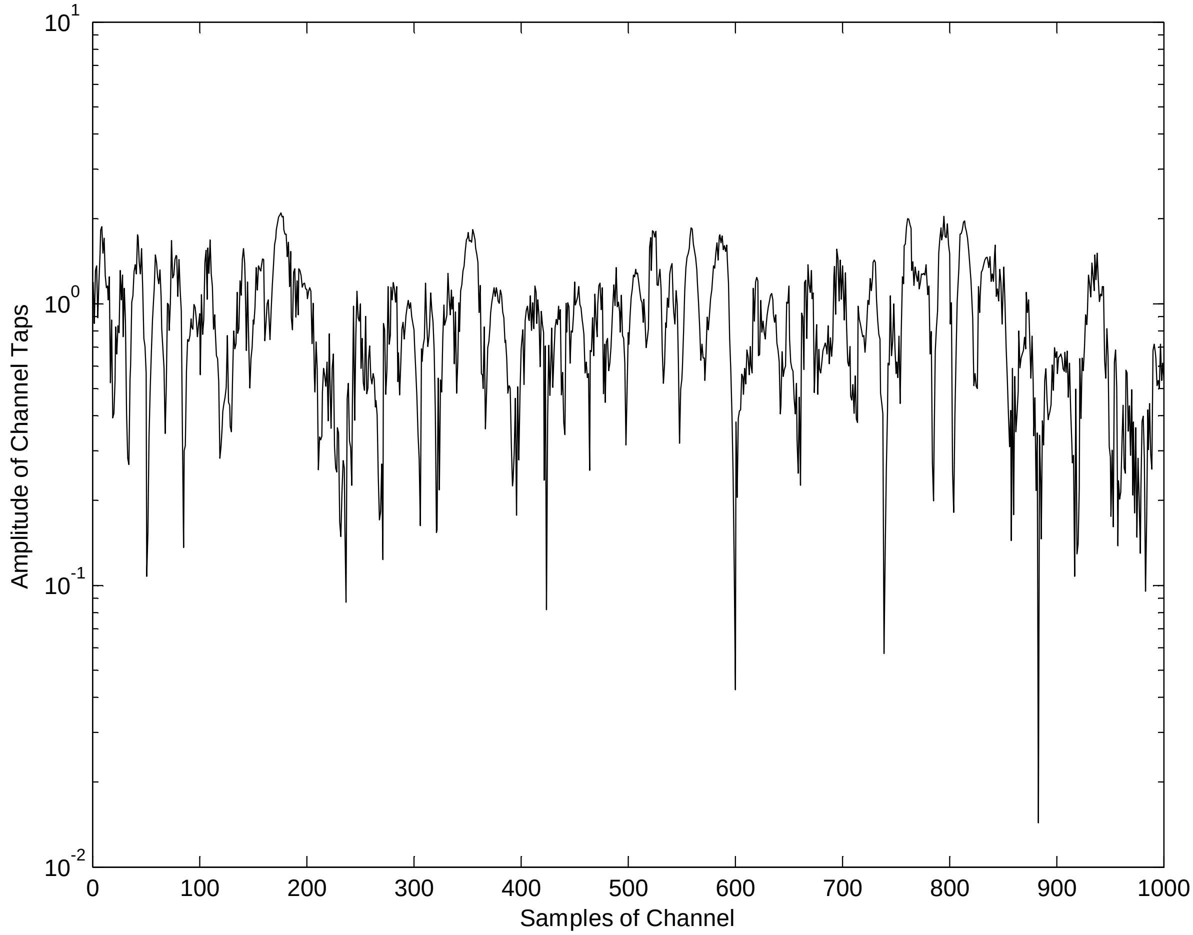

Figure 3.7: Fading Envelope at $f_d = 240$ Hz

Chapter 4

Channel Estimation in OFDM Systems

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 [8]. This is a common technique in wireless communication system, which, since no channel estimates is needed, reduces the complexity of the receiver. Differential modulation is used in European DAB standard [56]. The drawbacks are about a 3 dB noise enhancement [8], and inability to use efficient multiamplitude constellations. An interesting alternative of DPSK is differential amplitude phase shift keying [74], 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 bit error rate is high, such as in DVB [75].

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 systems only .

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 [48]. 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.

4.1 Pilot Symbol Assisted Modulation

Channel estimation usually needs some kind of pilot information as a point of reference. Channel estimates are often achieved by multiplexing known symbols, so called, pilot symbols into the data sequence, and this technique is called Pilot Symbol Assisted Modulation (PSAM) [9]. This method relies upon the insertion of known phasors into the stream of useful information symbols for the purpose of channel sounding. These pilot symbols allow the receiver to extract channel attenuations and phase rotation estimates for each received symbol, facilitating the compensation of fading envelope and phase. Closed form formula for the BER of PSAM were provided by Cavers [11] for binary phase shift keying (BPSK) and quadrature phase shift keying (QPSK), while for 16-QAM he derived a tight upper bound of the BER.

A fading channel requires constant tracking, so pilot information has to be transmitted more or less continuously. Decision directed channel estimation [76] can also be used, but even in these type of schemes pilot information has to be transmitted regularly to mitigate error propagation. Pilot symbols are transmitted at certain locations of the OFDM frequency time lattice, instead of data, and in [48], it was addressed how you choose those locations. An example of this is shown in Figure 4.1, which shows both scattered and continual pilot symbols. In general fading channel can be viewed as a 2-D signal (time and frequency), which is sampled at pilot positions and channel attenuations between pilots are estimated by interpolations. However, as in single carrier case [45], the pilot patterns should be designed so that the channel is oversampled at the receiver.

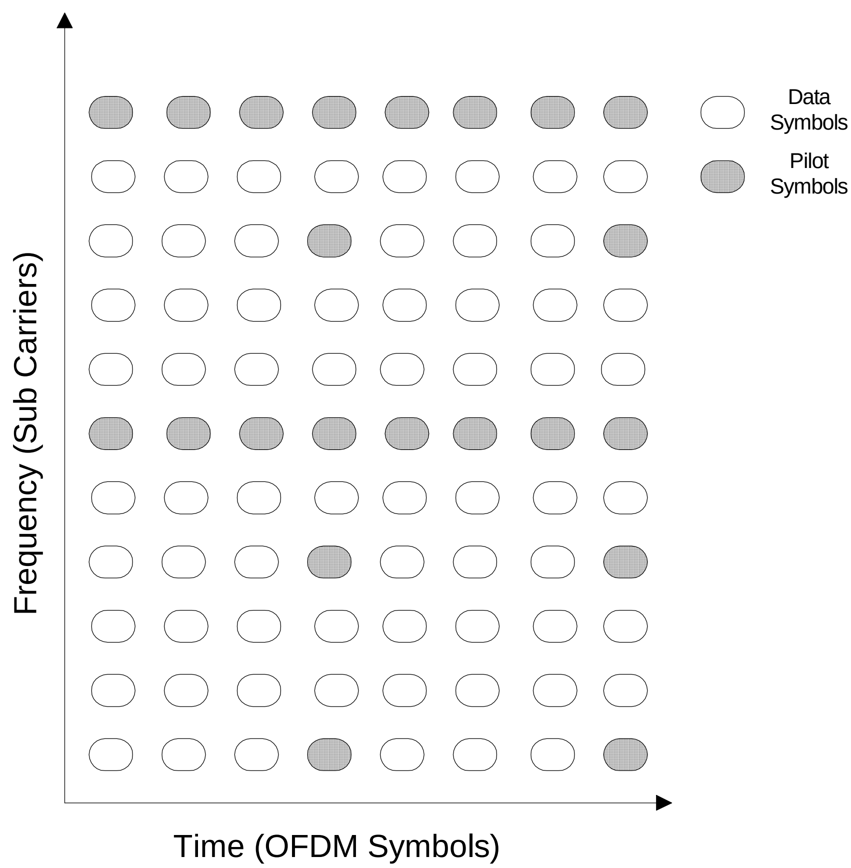

Figure 4.1: An Example of Pilot Information Transmission both as Scattered and Continual on certain subcarriers

4.2 Pilot Arrangements Used in Literature

In our knowledge, there is little published work on how to transmit pilot symbols in wireless OFDM systems, except [48], in which a comparison was shown, between some common arrangements of pilots. Use of pilot symbols for channel estimation introduces overhead and it is desirable to keep the number of pilot symbols as minimum as possible. The problem is to decide where and how often to insert pilot symbols. The spacing between pilot symbols is small enough to make channel estimates reliable and large enough not to increase overhead too much. The number of pilot tones necessary to sample the transfer function can be determined on the basis of sampling theorem as follows [54]: The frequency domain channel's transfer function *H(f)* is the fourier transform of the impulse response *h(t)*. Each of the impulses in the impulse response will result a complex exponential function $e^{\frac{-j2\pi\tau}{T_s}}$ in the frequency domain, depending on its time delay τ, where T_s is the symbol time. In order to sample this contribution to $H(f)$ according to the sampling theorem, the maximum pilot spacing Δp in the OFDM symbol is:

$$\Delta p \leq \frac{N}{2\tau/T_s}\Delta f. \tag{4.1}$$

where Δf is the subcarrier bandwidth.

Using a dense pilot patterns means that the channel is oversampled, implying that low-rank estimation methods can work well [44]. This type of low complexity estimation projects the observations into a subspace of smaller dimension and perform the estimation in that subspace. By oversampling the channel, that is placing the pilot symbols close to each other, the observations essentially lie in a subspace and low rank estimation is very effective [77].

The channel estimation can be performed by either inserting pilot tones into all of the subcarriers of OFDM symbols with a specific period or inserting pilot tones into each OFDM symbol [43]. The first one, block type pilot channel estimation, has been developed under the assumption of slow fading channel. This type of pilot arrangements works well when the channel transfer function is not changing very rapidly. The later one, comb type pilot arrangement, can be used easily for tracking fast channels. In comb arrangements, every OFDM symbol have some pilot tones, therefore these type of patterns works well in highly varying environments, as will be demonstrated in simulations in subsequent Chapters. Block and comb arrangements are shown in Figure 4.2 and 4.3 respectively.

4.3 Pilot Signal Estimation

Channel can be estimated at pilot frequencies by two ways:

1. (LS) Estimation
2. (LMMSE) Estimation

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

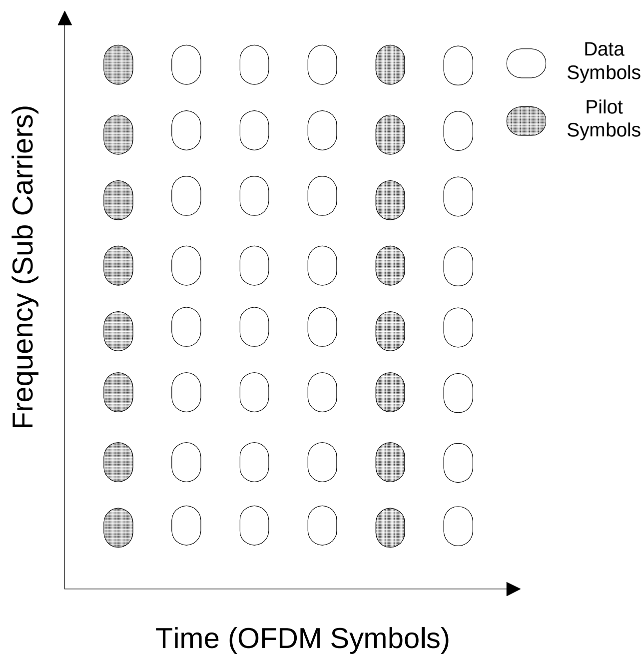

Figure 4.2: Block Pilot Patterns

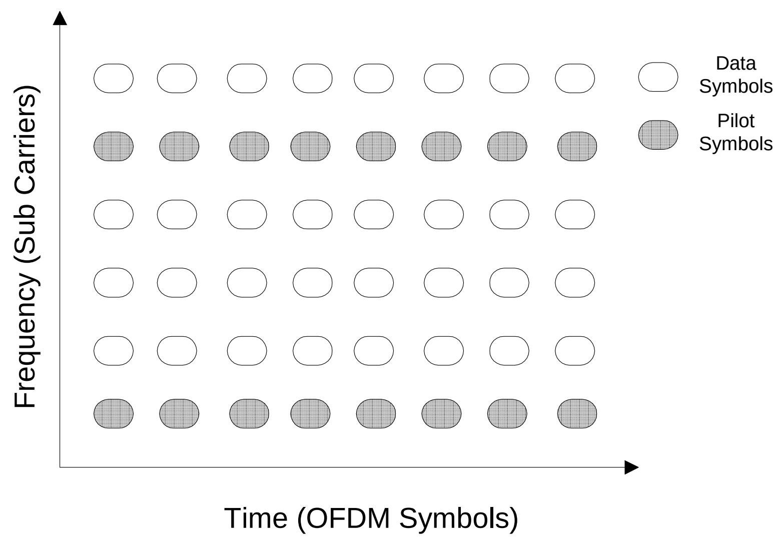

Figure 4.3: Comb Pilot Patterns

block. So in block type estimation, we first estimate the channel, and than use the same estimates within the entire block. LMMSE estimation has been shown to yield 10-12dB gain in signal to noise ratio (SNR) over LS estimation for the same mean square error of channel estimation [77]. In [44], a 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 to interpolate the channel, 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) [43]. MMSE has been shown to perform much better than LS. In [49], the complexity of MMSE is reduced by deriving an optimal low rank estimator with singular value decomposition (in actual its basically eigen value decomposition).

4.3.1 Least Square Estimation

The idea behind least squares is to fit a model to measurements in such a way that weighted errors between the measurements and the model are minimized [77]. The LS estimate of the attenuations $\mathbf{h}$, given the received data $\mathbf{Y}$ and the transmitted symbols $\mathbf{X}$ is [6]:

$$\hat{\mathbf{h}}_{\ell s} = \mathbf{X}^{-1}\mathbf{Y} = [\frac{y_0}{x_0} \quad \frac{y_1}{x_1} \quad ... \quad \frac{y_{N-1}}{x_{N-1}}]^T \tag{4.2}$$

For comb type pilot subcarrier arrangement, the N_p pilot signals $\mathbf{X_p}(m)$, $m = 0, 1, ..., N_p - 1$ are uniformly inserted into $\mathbf{X}(k)$. That is, the total N subcarriers are divided into N_p groups, each with $GI = N/N_p$ adjacent subcarriers. In each group, the first subcarrier is used to transmit pilot signal. The OFDM signal modulated on the kth subcarrier can be expressed as

$$\begin{aligned} X(k) &= X(mGI + \ell) \\ &= \begin{cases} x_p(m), & \ell = 0 \\ infinite\ data, & \ell = 1, ..., N-1 \end{cases} \end{aligned} \tag{4.3}$$

The pilot signal $\mathbf{X_p}(m)$ can be either complex values c to reduce the computational complexity, or random generated data that can also be used for synchronization.

Let

$$\begin{aligned}\mathbf{H}_p &= [H_p(0) \;\; H_p(1) \;\; ... \;\; H_p(N_p-1)]^T & (4.4)\\ &= [H(0) \;\; H(GI-1) \;\; ... \;\; H((N_p-1).GI-1)]^T & (4.5)\end{aligned}$$

be the channel response of pilot subcarriers, and

$$\mathbf{Y}_p = [Y_p(0) \;\; Y_p(1) \;\; ... \;\; Y_p(N_p-1)]^T \tag{4.6}$$

be the vector of received pilot signals. The received pilot signal vector $\mathbf{Y}_p$ can be expressed as

$$\mathbf{Y}_p = \mathbf{X}_p.\mathbf{H}_p + \mathbf{I}_p + \mathbf{W}_p \tag{4.7}$$

where

$$\mathbf{X}^p = \begin{bmatrix} X_p(0) & 0 \\ & \\ 0 & X_p(N_p-1) \end{bmatrix} \tag{4.8}$$

$\mathbf{I}_p$ is a vector of ICI and $\mathbf{W}_p$ is the vector of Gaussian noise in pilot subcarriers.

In conventional comb type pilot estimation, the estimate of pilot signals based on least square (LS) criterion, is given by [53],

$$\mathbf{H}_{p,\ell s} = [H_{p,\ell s}(0) \;\; H_{p,\ell s}(1) \;\; \dots \;\; H_{p,\ell s}(N_p - 1)]^T \tag{4.9}$$

$$= \mathbf{X}_p^{-1}\mathbf{Y}_p \tag{4.10}$$

$$= [\frac{Y_p(0)}{X_p(0)} \;\; \frac{Y_p(1)}{X_p(1)} \;\; \dots \;\; \frac{Y_p(N_p-1)}{X_p(N_p-1)}]^T \tag{4.11}$$

The LS estimate of $\mathbf{H}_p$ is susceptible to Gaussian noise and inter-carrier interference (ICI). Because the channel responses of data subcarriers are obtained by interpolating the channel responses of pilot subcarriers, the performance of OFDM system based on comb type pilot arrangement is highly dependent on the rigorousness of estimate of pilot signals. Thus a estimate better than LS estimate is required.

4.3.2 Linear Minimum Mean Square Error Estimation

The linear minimum mean square error (LMMSE) estimate has been shown to be better than the LS estimate for channel estimation in OFDM systems based on block type pilot arrangement [53]. Regarding the mean square error estimation shown in [53], the LMMSE estimate has about 10-15dB gain in SNR over LS estimate for the same MSE values. The major drawback of the LMMSE estimate is its high complexity, which grows exponentially with observation samples [49]. In [44], a low rank approximation is applied to a linear minimum mean squared error estimator (LMMSE estimator) that uses the frequency correlations of the channel.

Assume that all the available LS estimates are arranged in a vector $\hat{\mathbf{p}}$ and the channel values that have to be estimated from $\hat{\mathbf{p}}$ are in a vector $\mathbf{h}$. The channel estimation problem is now to find the channel estimates $\hat{\mathbf{h}}$ as a linear combination of pilot LS estimates $\hat{\mathbf{p}}$. According to [77], the minimum mean square error estimate for this problem is given by

$$\hat{\mathbf{h}}_{\ell\mathbf{mmse}} = \mathbf{R}_{h\hat{p}}(\mathbf{R}_{\hat{p}\hat{p}})^{-1}\hat{\mathbf{p}} \tag{4.12}$$

$R_{h\hat{p}}$ is the cross-covariance matrix between $\mathbf{h}$ and the noisy pilot estimates $\hat{\mathbf{p}}$, given by

$$\mathbf{R}_{h\hat{p}} = E\{\mathbf{h}\hat{\mathbf{p}}^{\mathbf{H}}\} \tag{4.13}$$

$\mathbf{R}_{\hat{p}\hat{p}}$ is the auto-covariance matrix of the pilot estimates, and is given by [28]:

$$\mathbf{R}_{\hat{p}\hat{p}} = E\{\hat{\mathbf{p}}\hat{\mathbf{p}}^{\mathbf{H}}\} \tag{4.14}$$

$$= \mathbf{R}_{pp} + \sigma_n^2(\mathbf{p}\mathbf{p}^{\mathbf{H}})^{-1} \tag{4.15}$$

where σ_n^2 is the variance of additive channel noise. The superscript $(.)^H$ denotes Hermitian transpose. Now for the case of block-type pilot channel estimation, Equation (4.12) can be modified as:

$$\hat{\mathbf{h}}_{\ell mmse} = \mathbf{R}_{hh}(\mathbf{R}_{hh} + \sigma_n^2(\mathbf{p}\mathbf{p}^{\mathbf{H}})^{-1})^{-1}\hat{\mathbf{p}} \tag{4.16}$$

In the following, we assume, without loss of generality, that the variances of the channel attenuations in $\mathbf{h}$ are normalized to unity, i.e. $E\{|\mathbf{h_k}|^2\} = 1$.

The LMMSE estimator defined in Equation (4.16) is of considerable complexity, since a matrix inversion is needed every time the training data in $\mathbf{p}$ changes.

The complexity of this estimator can be reduced by averaging over the transmitted data [8], i.e., we replace the term $(\mathbf{pp^H})^{-1}$ in Equation (4.16) with its expectation $E\{(\mathbf{pp^H})^{-1}\}$. Assuming the same signal constellation on all tones and equal probability on all constellation points, we have $E\{(\mathbf{pp^H})^{-1}\} = E\{|\frac{1}{\mathbf{p_k}}|^2\}\mathbf{I}$, where $\mathbf{I}$ is the identity matrix. Defining the average signal-to-noise ratio as

$$SNR = E\{|p_k|^2\}/\sigma_n^2$$

we obtain a simplified estimator [44],

$$\hat{\mathbf{h}}_{\ell mmse} = \mathbf{R}_{hh}(\mathbf{R}_{hh} + \frac{\beta}{SNR}\mathbf{I})^{-1}\hat{\mathbf{p}} \tag{4.17}$$

where

$$\beta = E\{|\mathbf{p_k}|^2\}E\{|1/\mathbf{p_k}|^2\} \tag{4.18}$$

is a constant depending on the signal constellation. In the case of 16-QAM transmission, $\beta = \frac{17}{9}$. Because $\mathbf{p}$ is no longer a factor in the matrix calculation, the inversion of $\mathbf{R}_{hh} + \frac{\beta}{SNR}\mathbf{I}$ does not need to be calculated each time the transmitted data in $\mathbf{p}$ changes. Furthermore, if $\mathbf{R}_h h$ and SNR are known beforehand or are set to fixed nominal values, the matrix $\mathbf{R}_h h(\mathbf{R}_{hh} + \frac{\beta}{SNR}\mathbf{I})^{-1}$ needs to be calculated at once. Under these conditions, the estimation requires N multiplications per tone. Estimator can be further simplified by using low rank approximations as discussed in [44].

4.4 Channel Interpolation

After the estimation of the channel transfer function of pilot tones, the channel transpose of data tones can be interpolated according to adjacent pilot tones. The linear interpolation has been studied in [78], and is shown to be better than piecewise-constant interpolation. Here in this thesis, we consider the following interpolation schemes:

1. Linear Interpolation
2. Spline Interpolation
3. Cubic Interpolation
4. Low Pass Interpolation

In [43], cubic and spline interpolations has been shown to perform better than the linear interpolation, which is also consistent with our simulation results as discussed in the next Chapter. In this thesis, we propose a new pilot insertion scheme, and then compare proposed pattern with the above mentioned schemes.

4.4.1 Linear Interpolation

In the linear interpolation algorithm, two successive pilot subcarriers are used to determine the channel response for data subcarriers that are located in between the pilots [78]. For data subcarrier k, $mGI \leq k \leq (m+1)GI$, the estimated channel response using linear interpolation method is given by:

$$\hat{H}(k) = \hat{H}(mGI + \ell) \tag{4.19}$$

$$= (1 - \frac{\ell}{GI})\hat{H}(m) + \frac{\ell}{GI}\hat{H}_p(m+1) \tag{4.20}$$

The linear channel interpolation can be implemented by using digital filtering such as Farrow-structure [79]. Furthermore, by carefully inspecting Equation (4.19), we find that if GI is chosen as a power of 2, the multiplications operations involved in Equation (4.12) can be replaced by shift operations, and therefore no multiplication operation is needed in the linear channel interpolation.

4.4.2 Spline and Cubic Interpolation

Spline and Cubic interpolations are done by using "interp1" function of matlab. Spline and Cubic interpolations produce a smooth and continuous polynomial fitted to given data points. Spline interpolations works better than linear interpolation for comb pilot arrangement [43].

4.4.3 Low Pass Interpolation

The low pass interpolation is performed by inserting zeros into the original sequence and then applying a lowpass FIR filter that allows the original data to pass through unchanged and interpolates between such that the mean-square error between the interpolated points and the ideal values is minimized.

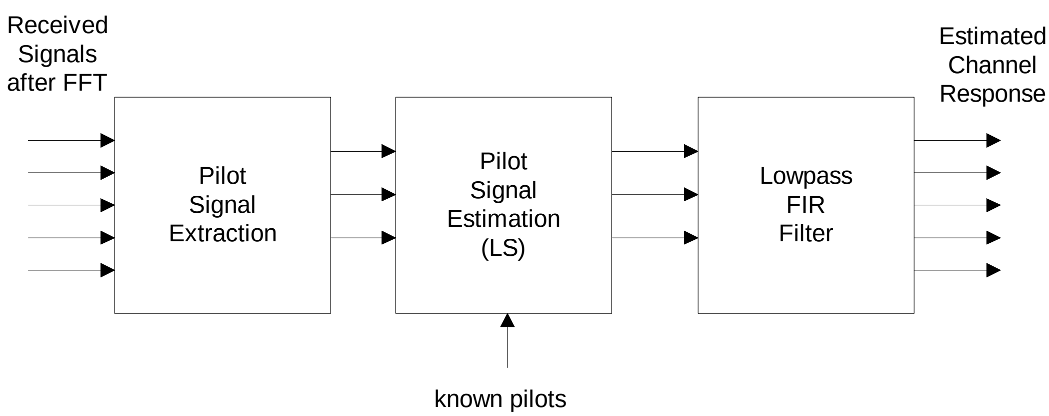

Figure 4.4: Block Diagram of Channel Estimation algorithm based on Comb-type pilots using Lowpass FIR filter

Chapter 5

Simulation Result and Discussions of Basic OFDM System

An OFDM system is modeled 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. Two main criteria are used to assess the performance of the OFDM system, which are its tolerance to multipath delay spread and channel noise.

5.1 OFDM Model Used in Simulations

An OFDM system model used is shown in Figure 5.1.

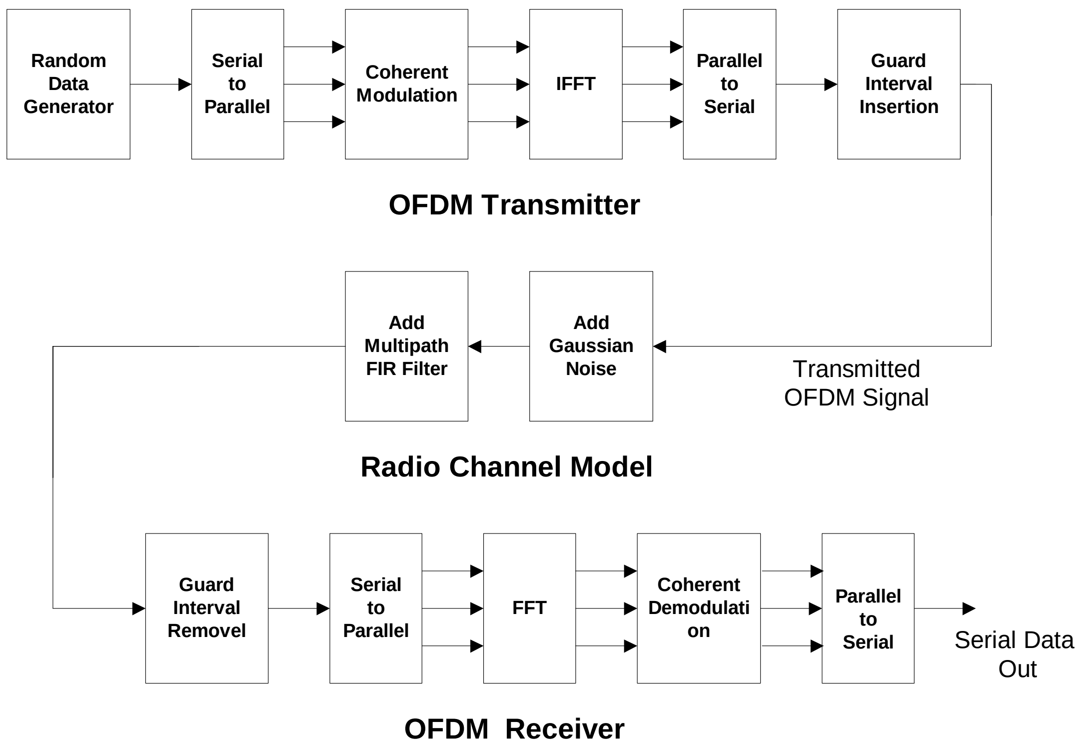

Figure 5.1: OFDM Model Used For Simulations

The input signal which we used is the random data generated by $randn()$ function of the matlab, and limit the data to its maximum value e.g. 16 (for 16QAM). However, we can extract data from some other sources e.g. from an audio file.

5.1.1 Serial to Parallel Conversion

The input serial data stream is formatted into the word size required for transmission, e.g. 2bit/word for QPSK, and shifted into a parallel format. The data is then transmitted in parallel by assigning each data word to one carrier in the transmission.

5.1.2 Modulation of Data

The data to be transmitted on each carrier is modulated into a QAM format. The data on each symbol is mapped and a QAM signal is generated. Here we used 16 QAM modulation.

5.1.3 Inverse Fourier Transform

After the required spectrum is worked out, an inverse fourier transform is used to find the corresponding time waveform. The guard period is then added to the start of each symbol.

5.1.4 Channel Model Used

A Channel model is then applied to the transmitted signal. The model allows for the signal to noise ratio and multipath to be controlled. The signal to noise ratio is set by adding a known amount of white noise to the transmitted signal. 16QAM modulation scheme is used for a 64-subcarrier OFDM system with a two ray multipath channel. The power of the second path is 6dB lower than the first one.

5.1.5 Receiver

The receiver basically does the reverse operation to the transmitter. The guard period is removed. The FFT of each symbol is then taken to find the original transmitted spectrum. Each transmission carrier is then evaluated and converted back to the data word by demodulating the received symbol. The data words are then combined back to the same word size as the original data.

5.2 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 [28] :

- As a rule of thumb, the guard time GI should be at least twice the delay spread, i.e.

$$GI \geq 2\tau$$

- To minimize the signal-to-noise ratio (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.

$$T_s \geq 5GI$$

- The frequency spacing between two adjacent subcarriers $\triangle f$ is

$$\triangle f = \frac{1}{T_s}$$

- 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

$$N = \frac{B_{info}}{R_{sub}}$$

- The OFDM signal bandwidth is defined as

$$BW = N\triangle f$$

Two observations are made from the above calculations:

1. 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.

2. Increasing the number of subcarriers increases the number of samples per OFDM symbol. However, it does not necessary 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 an 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.

5.3 Gaussian Noise Tolerance of OFDM

It was found that the SNR performance of OFDM in AWGN channel is similar to a standard single carrier digital transmission [60]. This is to be expected as the transmitted signal is similar to a standard frequency division multiplexing (FDM) system. Figure 5.2 shows the results from the simulations.

Performance of OFDM in AWGN channel with different values of M is shown

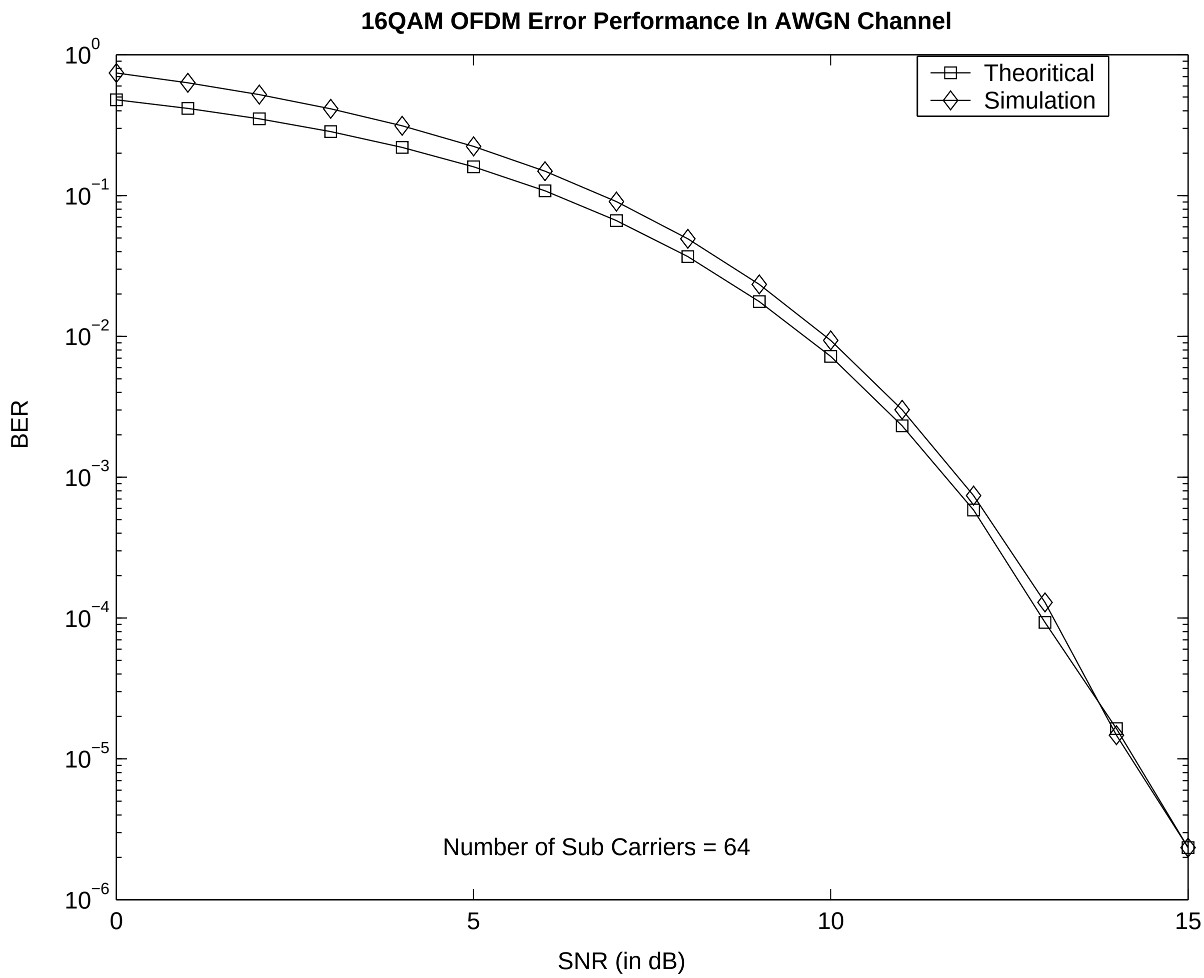

Figure 5.2: BER versus Channel SNR for OFDM in AWGN Channel

in Figure 5.3. The results shows that using QAM transmission can tolerate a SNR of greater than 5dB. The bit error rate gets rapidly worse as the SNR drops below 5dB. However, using 4QAM allows the BER to be improved in a noisy channel, at the expense of transmission data capacity. Using 4QAM the OFDM transmission can tolerate a SNR of 3-4dB. In a low noise link the capacity can be increased by using 16PSK.

5.4 Multipath Delay Spread Immunity

This section provides the performance evaluation of OFDM systems in the static multipath environment. 16QAM modulation scheme is used for a 64-subcarrier OFDM system with a two ray multipath channel, as mentioned earlier. The power of the second path is 6dB lower than the first one. No noise is present at the receiver in order to have a clear idea of the influence of ISI and ICI on the system performance with respect to these two parameters.

Figure 5.4 shows the 16QAM signal constellation diagram for input transmitted symbols. Figure 5.5 shows the 16QAM signal constellation diagram for delay spread less than guard time and no channel estimation is implemented at the receiver. It shows that the received signal points have a circular pattern around the transmitted signal points. Self-interference moves some of the signal points over the decision boundaries and results in significant degradation in bit error rate (BER)

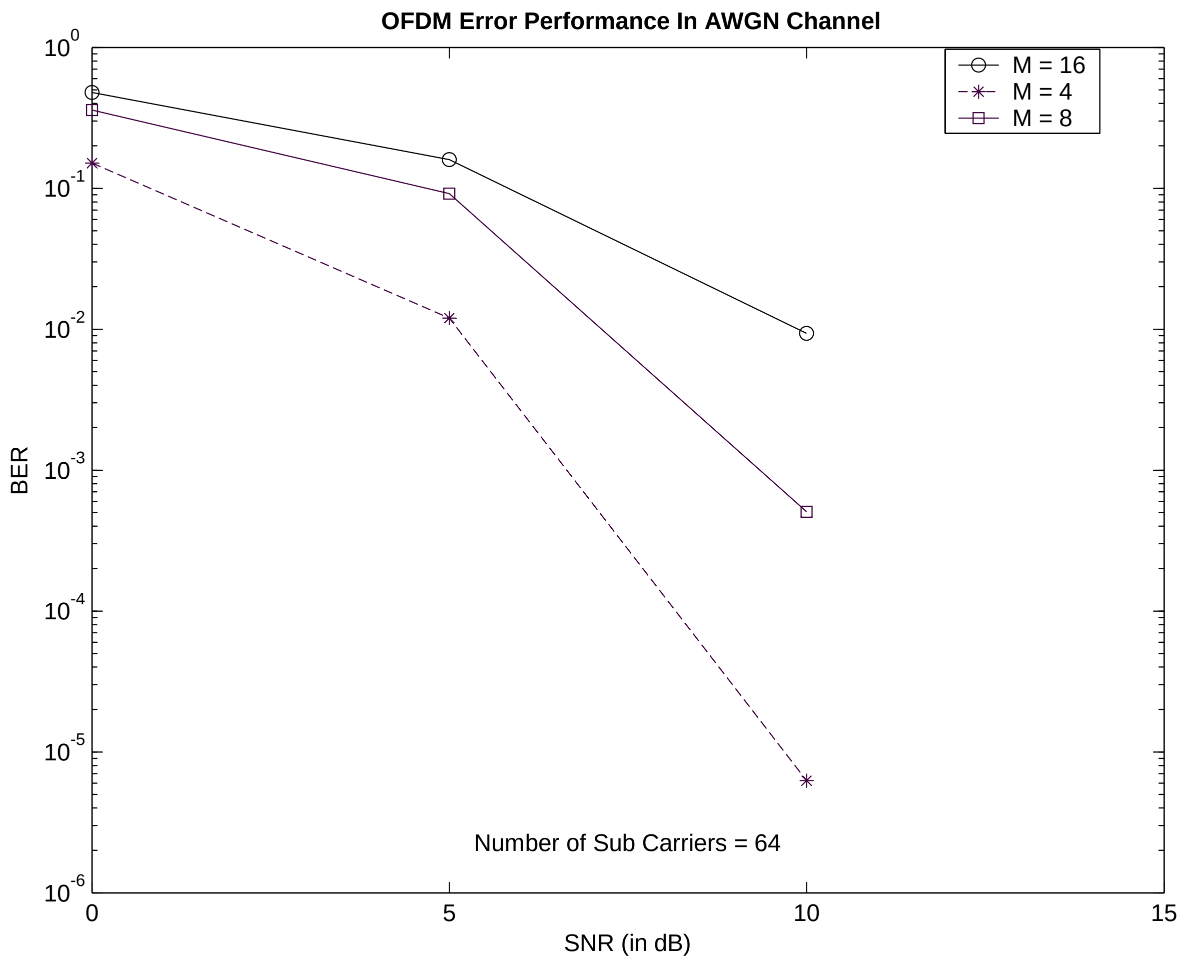

Figure 5.3: BER verse SNR for OFDM using 4QAM, 8QAM and 16QAM

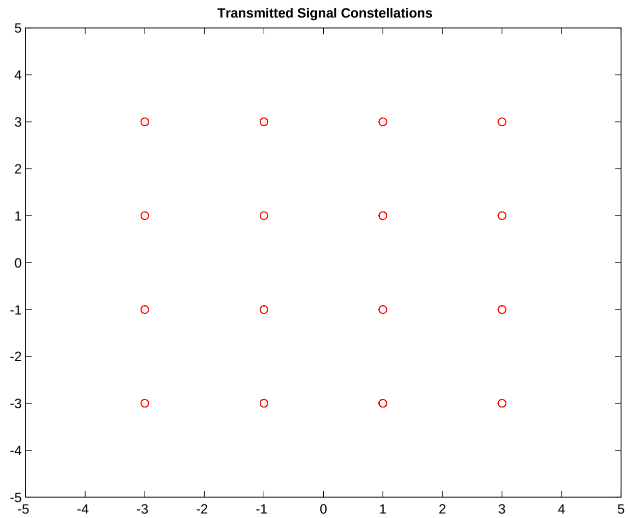

Figure 5.4: Transmitted Signal Constellation

performance. Therefore, channel estimator must be implemented at the FFT output to correct the amplitude and phase distortion caused by multipath distortion [64]. From Figure 5.6, it is clear that without channel estimator error rate at receiver is very high.

The circular pattern on the signal constellation diagram can also be observed by performing circular convolution of a multipath channel and an OFDM symbol without cyclic extension. As pointed out in [63] and [80], the cyclic extension makes the linear convolution of the channel looks like circular convolution inherent to the discrete Fourier domain, as long as the guard time duration is longer than the delay spread of the multipath channel.

Figure 5.7 shows the 16QAM signal constellation diagrams for the case of a 64-subcarrier OFDM system with a channel estimator at the receiver. Figure 5.7 shows

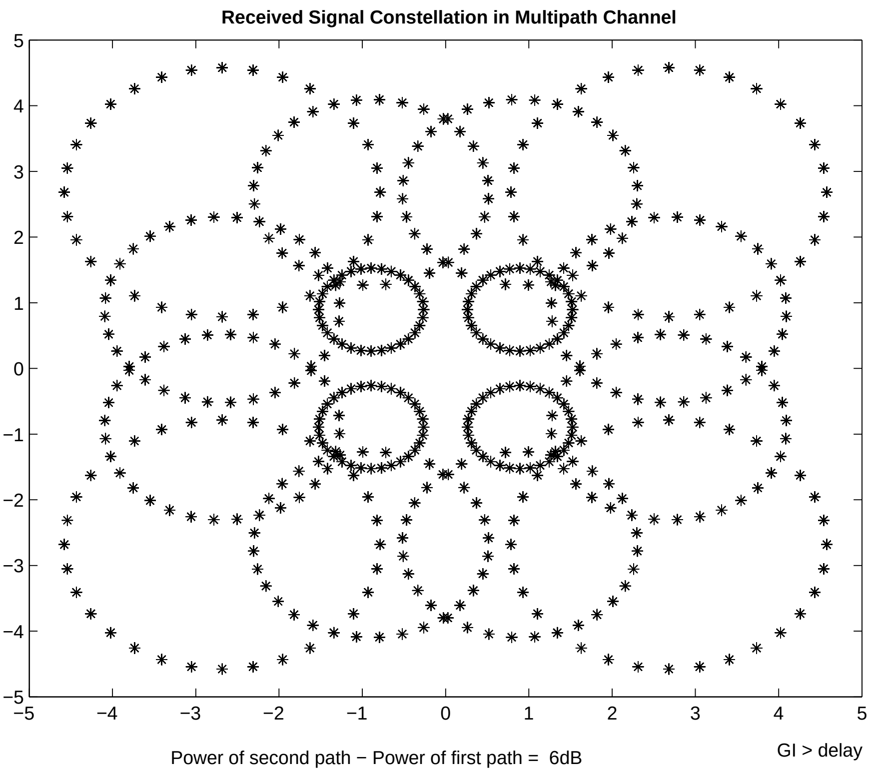

Figure 5.5: A 16QAM signal constellation diagram for a 64-subcarrier OFDM system without one tap equalizer at the receiver. The channel consists of two multipath, with the second one 6dB lower than the first one and the delay spread is less than guard time.

the 16QAM signal constellation for the case of delay spread less than the guard time duration. No distortion is observed since the delay spread is shorter than the guard time and the frequency-selective fading is compensated by the channel estimator.

Figures 5.8 and 5.9 show the constellation diagram for the case of delay spread greater than the guard time by 3.125% and 9.375% of the FFT interval, respectively. The distortion caused by ISI gets bigger as the delay spread exceeds the duration of the guard time more, resulting in higher BER.

From the constellation diagrams it is evident that if delay spread of the channel is greater than the guard interval, than error rate is quite high. Figure 5.10 shows the BER versus the maximum delay spread for an OFDM system with 64 subcarriers, and using the same channel, which have two paths, and power of the second path is 6dB lower than the first.

Simulation result shows that for the case of maximum delay spread less than guard time, no error is produced at the receiver. Once the delay spread exceeds the guard time, ISI is introduced. BER increases rapidly at the beginning and then gradually approaches an error floor as the effect of guard time to the delay spread on the performance becomes insignificant.

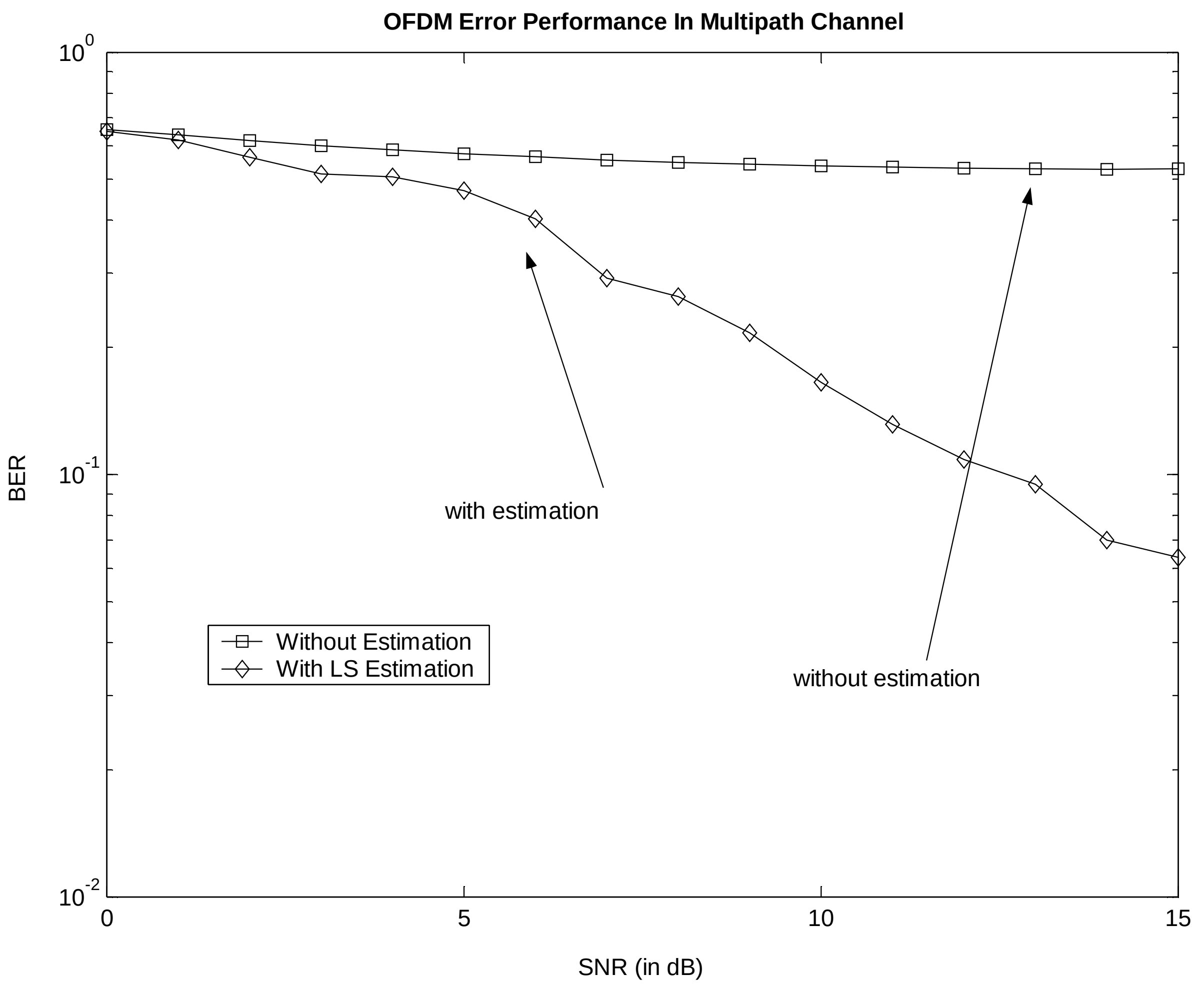

Figure 5.6: Improvement in BER because of channel estimation

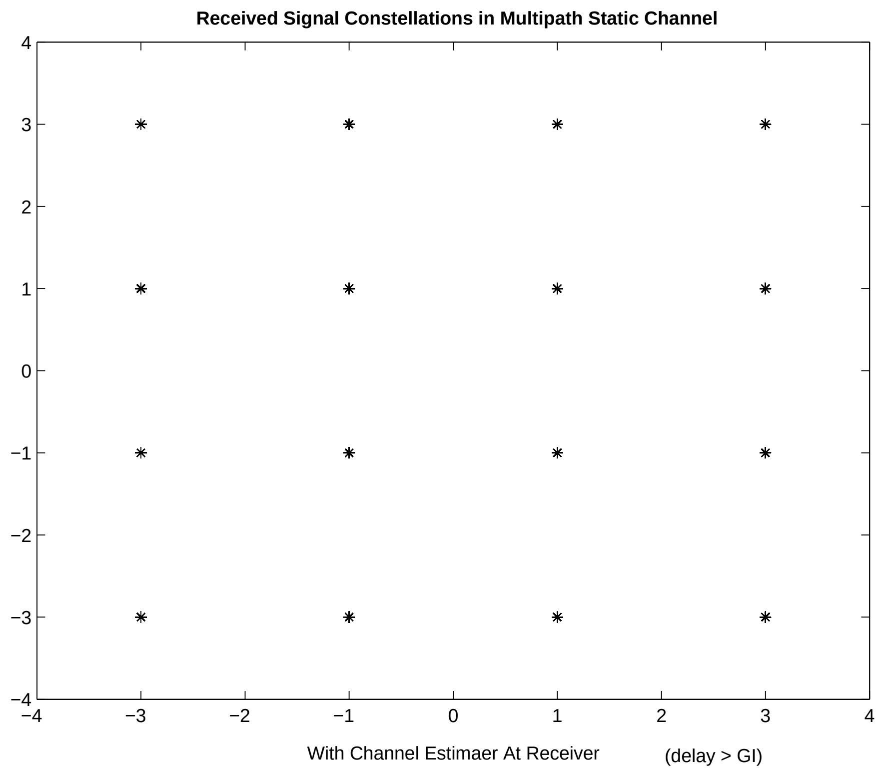

Figure 5.7: A 16QAM signal constellation diagram for a 64-subcarrier OFDM system without one tap equalizer at the receiver. The channel consists of two multipath, with the second one 6dB lower than the first one and the delay spread is less than guard time.

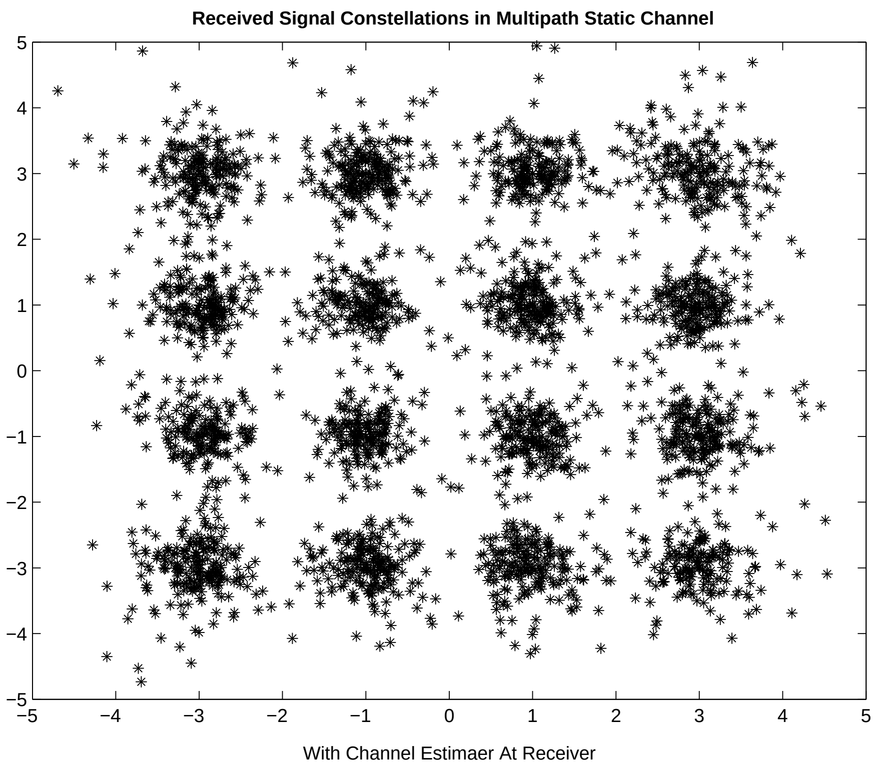

Figure 5.8: A 16QAM signal constellation diagram for a 64-subcarrier OFDM system without one tap equalizer at the receiver. The channel consists of two multipath, with the second one 6dB lower than the first one and the delay spread is greater than guard time by 3.125% of the FFT interval.

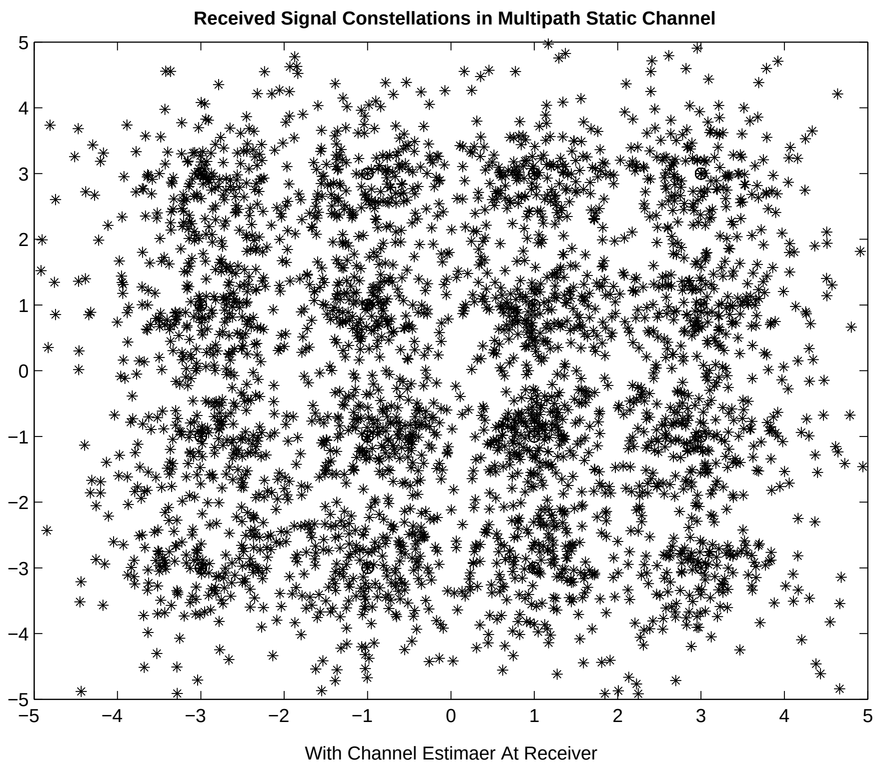

Figure 5.9: A 16QAM signal constellation diagram for a 64-subcarrier OFDM system without one tap equalizer at the receiver. The channel consists of two multipath, with the second one 6dB lower than the first one and the delay spread is greater than guard time by 9.375% of the FFT interval.

Figure 5.11 shows the BER versus OFDM systems with different number of subcarriers N. Three multipath channels with delay spread of 4, 8 and 12 samples are studied. No guard time is inserted to OFDM symbols. Simulation result shows that in all cases, BER decreases as the number of subcarriers increases. For the same signal bandwidth, increasing the number of subcarriers increases the symbol duration. The ratio of the number of distorted samples to the total number of samples per OFDM symbol decreases as the symbol duration increases and thus the BER is decreased.

5.5 Summary

In this chapter, we described basic concepts of OFDM system by presenting simulation results, including ISI, ICI, guard time and equalizers in multipath static environment. We studied the gaussian noise tolerance of OFDM. We also studied the effect of number of subcarriers and the guard time duration on the performance of OFDM systems. Furthermore, we explained the relationships between various OFDM parameters.

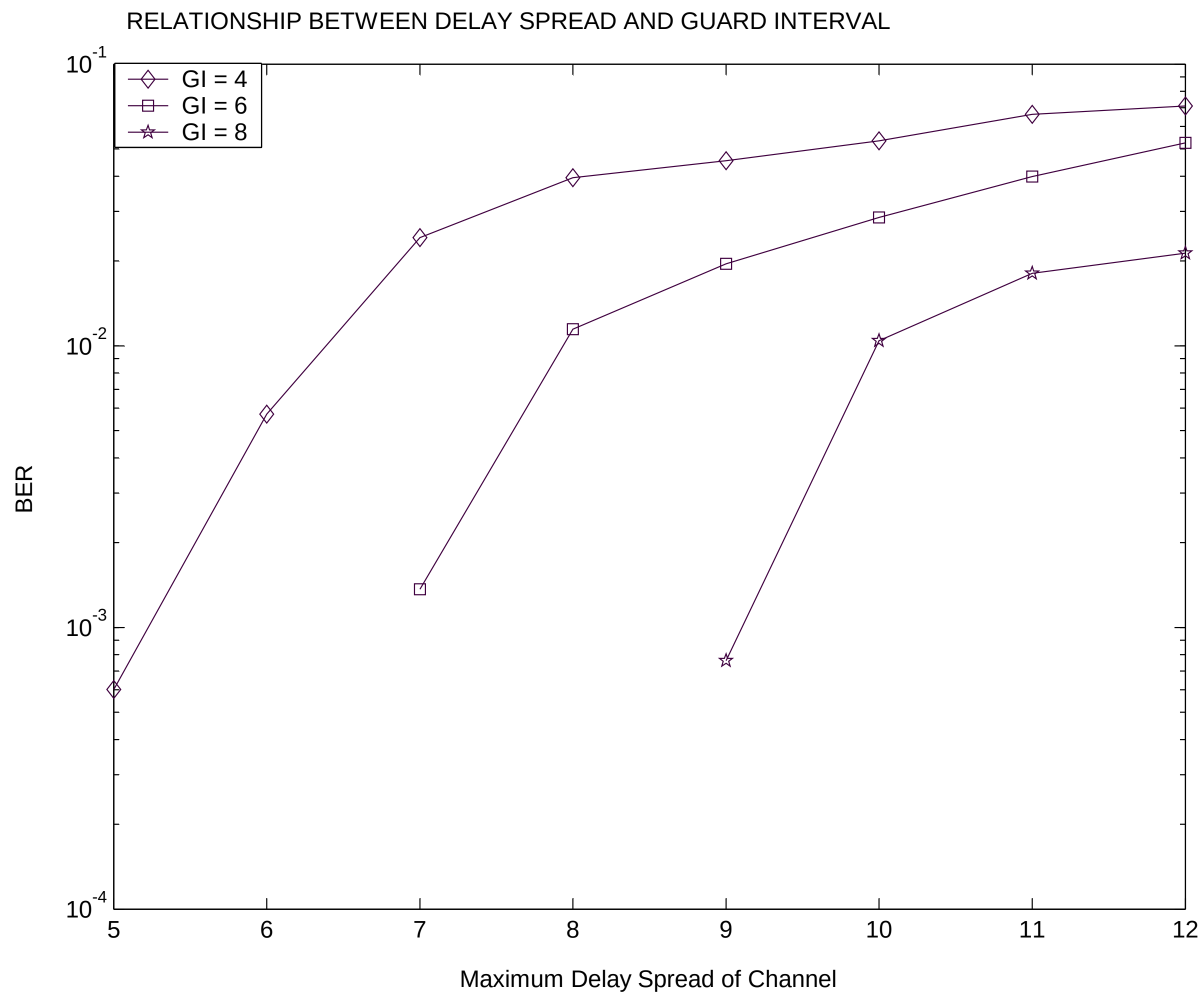

Figure 5.10: BER versus delay spread for a 64-subcarrier OFDM system with different guard time

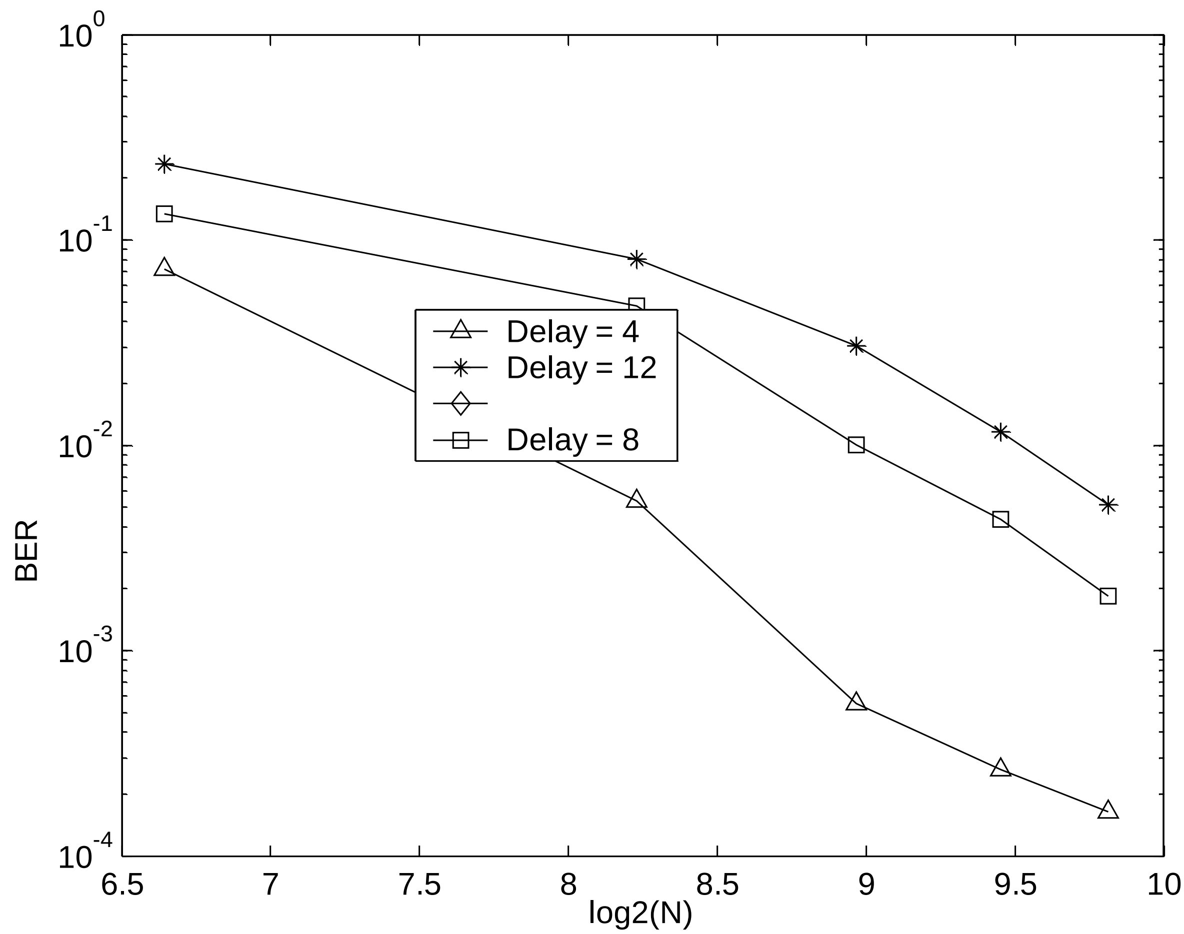

Figure 5.11: BER versus number of subcarriers with delay spread exceeds guard time

Chapter 6

Arrangement of Pilot Tones in Wireless OFDM Systems

Pilot Symbol Assisted Modulation (PSAM) is used to achieve reliable channel estimates by multiplexing pilot symbols, along with data symbols [9]. The receiver estimates channel attenuations and phase rotations at these pilot locations, and compensates for the effect of channel at data locations. A fading channel, described in previous chapter, requires constant tracking, so pilot information has to be sent more or less continuously. Decision directed channel estimation [76], can be used, but even in these schemes, pilots have to be transmitted regularly in order to mitigate error propagation. Pilot symbols are transmitted at certain locations of the OFDM time-frequency grid, and the question is how do we choose these locations.

In this chapter, we analyze different pilot patterns in terms of resulting bit error rate, and propose a new scheme of transmitting pilot symbols in wireless OFDM system. Rearrangement of the pilot pattern enables a reduction in the number of needed pilot symbols which in turns reduces the transmission overhead still retaining the same performance. The question is where and how often to transmit pilot symbols, so that the spacing between the pilot symbols shall be small enough to enable reliable channel estimates but large enough not to increase the overhead too much.

6.1 System Description

6.1.1 OFDM System Model

The basic baseband-equivalent OFDM system is shown in Figure 6.1. Each OFDM symbol consists of a packet of N data points, that are carried on N frequency tones respectively. IFFT block is used at the transmitter to transform the data sequence $X(k)$ of length N into time domain signal $\{x(n)\}$ with the following equation:

$$\begin{aligned} x(n) &= IDFT\{X(k)\} \qquad n = 0, 1, 2, ..., N-1 \\ &= \sum_{k=0}^{N-1} X(k) e^{j\frac{2\pi kn}{N}} \end{aligned} \tag{6.1}$$

Where N is the FFT length. Following IFFT block, cyclic prefix, which is chosen to be greater than the expected delay spread of the channel, is inserted to prevent ISI [60]. This cyclic prefix includes the cyclically extended part of OFDM symbol in order to eliminate intercarrier interference (ICI) [24]. As pointed out in [63], the cyclic extension makes the linear convolution of the channel looks like circular convolution inherent to the discrete Fourier domain, as long as guard time duration is longer than the delay spread of the multipath channel, as discussed in Chapter 2.

The resultant OFDM symbol is given as follows:

$$x_f(n) = \begin{cases} x(N+n), & n = -GI, -GI+1, ..., -1 \\ x(n), & n = 0, 1, ..., N-1 \end{cases} \tag{6.2}$$

where GI is the length of the guard interval. The transmitted signal $x_f(n)$ pass through the frequency selective time varying fading channel with additive white gaussian noise. The received signal is given by:

$$y_f(n) = x_f(n) \otimes h(n) + w(n) \tag{6.3}$$

where $w(n)$ is Additive White Gaussian Noise (AWGN) and $h(n)$ is the channel impulse response. Channel is assumed to be slowly fading, so it is considered to be constant during one OFDM symbol. Under these conditions we can describe our system as a set of parallel Gaussian channels, shown in Figure 6.2, with correlated attenuations $h(n)$, already discussed in Chapter 2. At the receiver, cyclic prefix is removed:

$$y(n) = \begin{cases} y_f(n), & for - GI \leq n \leq N-1 \\ y_f(n+GI) & n = 0, 1, ..., N-1 \end{cases} \tag{6.4}$$

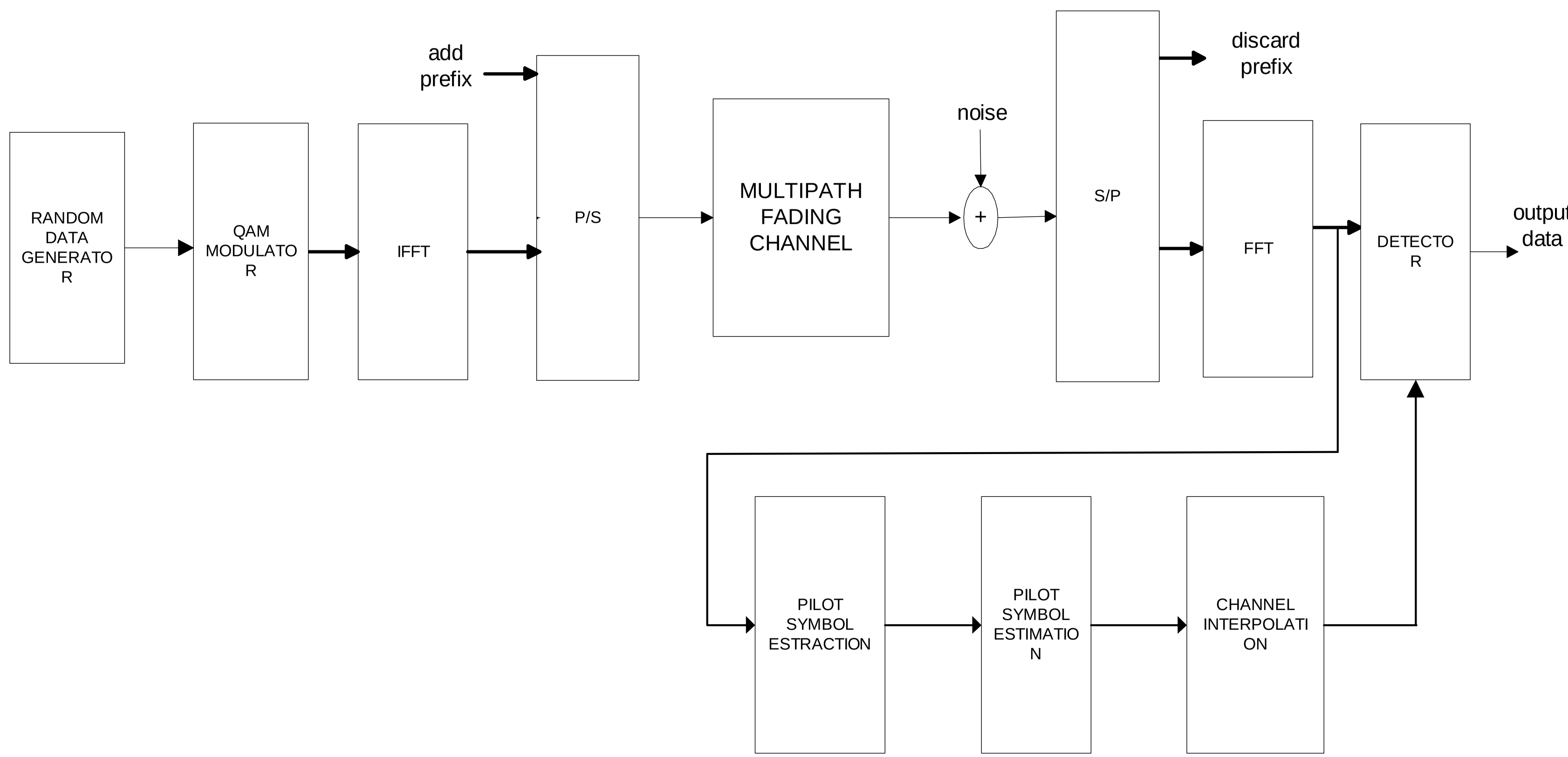

Figure 6.1: OFDM System Used in Simulations

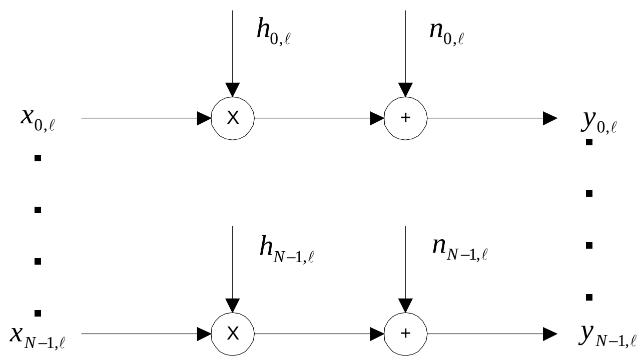

Figure 6.2: OFDM System, described as a set of parallel Gaussian channels with correlated attenuations

Then $y(n)$ is sent to FFT block for the following operation:

$$\begin{aligned} Y(k) &= DFT\{y(n)\} \\ &= \frac{1}{N}\sum_{n=0}^{N-1} y(n)e^{-j(\frac{2\pi kn}{N})} \end{aligned} \tag{6.5}$$

Assuming there is no ISI,in [81] it was shown that the relation of the resulting $Y(k)$ to $H(k)$, $I(k)$ that is ICI because of Doppler frequency and $W(k)$, with the following relation [49]:

$$Y(k) = X(k)H(k) + I(k) + W(k) \qquad k = 0, 1, 2, ..., N-1 \tag{6.6}$$

Following FFT block pilot signals are extracted and the estimated channel $H_e(k)$ at pilot sub-channels is obtained. Then the transmitted data is estimated by:

$$X_e(k) = \frac{Y(k)}{H_e(k)} \qquad k = 0, 1, 2, ..., N-1 \tag{6.7}$$

6.1.2 Channel Model Used in Simulations

We are using a fading multipath channel model [8], consisting of M paths

$$g(\tau) = \sum_{k=0}^{M-1} \alpha_k \delta(t - \tau_k T_s), \tag{6.8}$$

where α_k are Rayleigh distributed channel taps whose fading is based on Jake's model [68] as described in the previous Chapter, with an exponential power delay profile $\theta(\tau_k)$, defined as

$$\theta(\tau_k) = Ce^{-\frac{\tau}{\tau_{rms}}} \tag{6.9}$$

where τ_{rms} is the RMS-value of power delay profile. In this paper, we have used $M = 2$ paths, in which the first fading path always has a zero-delay, $\tau_o = 0$, and other fading path has delay that is always less than the length of guard interval GI.

6.1.3 Scenario

The channel estimation is based on pilots transmitted at certain positions in the time frequency grid of the OFDM system. The channel attenuations are estimated by means of interpolation between these pilots, where we assume that the channel estimator can use all transmitted pilots. This is the case in, e.g. broadcasting or in the downlink of a multiuser system. Channel attenuations in neighboring time-frequency grid points are highly correlated, a feature that can be used to channel estimation.

Our scenario consists of a wireless 16-QAM OFDM system, designed for an outdoor environment, that is capable of carrying digital video. The system operates at 500 kHz bandwidth and is divided into 64 tones with a total symbol period of 136 μs, of which 8 μs is the cyclic prefix. Our OFDM symbol thus consists of 68 samples ($N + GI = 68$), four of which are contained in the cyclic prefix ($GI = 4$). The uncoded data rate of the system is 1.9MBit/sec. We assume that $\tau_{rms} = 1$ sample for the channel considered.

6.1.4 Problem Formulation

The transmitter uses certain tones, so called pilot tones, in some particular symbols to transmit known data. The channel impulse response can be estimated using LS or LMMSE criterion, given knowledge of transmitted and received signals. The questions that arise are:

1. How many pilot tones are needed per symbol for estimation?

2. What pattern of pilot tones are better than others? Which tones should be used as pilot tones, and what is the impact of pilot tone selection on the quality of estimate ?

3. How does a scheme that uses some tones as pilot tones in each symbol (comb) can compare with a scheme that uses all tones as pilot tones (block) in some symbols? And how block and comb arrangements can be compared with the

proposed arrangement of pilots?

6.2 Number of Pilot Tones

One of the important question is about number of pilot tones per symbol needed for channel estimation.

Theorem 6.2.1 *In the absence of noise, any GI of the N available tones can be used for training to recover the channel h exactly.*

Proof: Let $\{k_1, k_2, ..., k_{GI}\}$ be the set of GI tones used for transmitting training data. The channel gains of these tones can be found exactly as $\mathbf{H}_{k_i} = \mathbf{R}_{k_i}$. Collect these gains in a vector $\mathbf{H}^p = (k_1, k_2, ..., k_{GI})^T$. Then we can write

$$\mathbf{H}^p = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & W_N^{k_1} & ... & W_N^{k_1(GI-1)} \\ 1 & W_N^{k_2} & ... & W_N^{k_2(GI-1})} \\ ... & ... & ... & ... \\ 1 & W_N^{k_{GI}} & ... & W_N^{k_{GI}(GI-1)} \end{bmatrix} \mathbf{h} \tag{6.10}$$

where

$$W_N^k = e^{-j2\pi k/N}$$

Since the matrix is a Vandermonde matrix with all parameters GI distinct, therefore it is non-singular [82], and hence h can be found exactly by inverting it. Also, with less than GI pilot tones, we have an under-determined system of linear equations, which results in a non-unique solution h.

6.3 Proposed Pattern of Pilot Tones

AS we already described in Chapter 3, channel estimation can be performed by either inserting pilot symbols into all subcarriers of OFDM symbols with a specific period (block arrangement) or inserting pilot tones into each OFDM symbol (comb arrangement) [43]. Block pilot patterns are effective in slow varying channels, and underlying assumption is, channel transfer function changes very slowly. However, comb type arrangement works well in fast varying channels, therefore, comb patterns can be used easily for tracking fast channels. In comb patterns, every OFDM symbol have some known data, i.e., pilots, in contrast to block patterns, where some specific OFDM symbols have pilots.

To minimize the bit error rate, it is desirable to spread the pilot symbols in both time and frequency, in contrast to block and comb, in which pilot symbols are transmitted in frequency and time, respectively, but not too far apart in case of fast fading. Here, we propose that instead of sending all pilots in one OFDM symbol or sending pilots in all OFDM symbols, better way is to spread the pilots in time and frequency, as shown in Figure 6.3. Simulation result shows that it works well especially in slow varying channels.

6.4 Simulation Results and Discussions

For all simulations, we assume that:

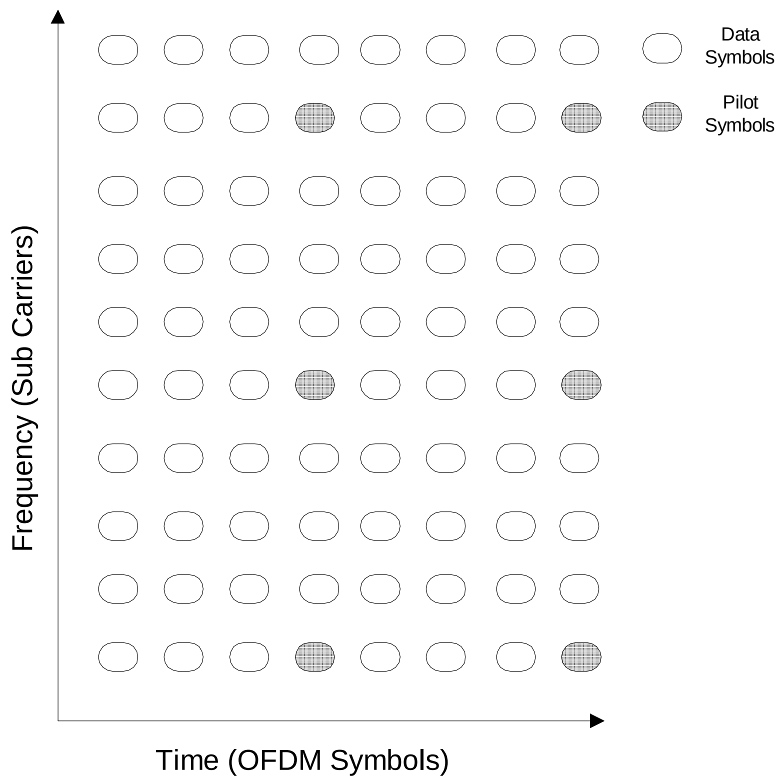

Figure 6.3: Proposed Pilot Arrangement

1. Cyclic Extension of OFDM symbols are used as guard interval.

2. The channel impulse response is shorter than the cyclic prefix to avoid ISI.

3. Transmitter and Receiver are perfectly time synchronized.

4. Channel is assumed to remain unchanged during one OFDM symbol, to avoid ICI.

All simulation parameters are shown in table 6.1.

Parameter	**Specifications**
Number of Subcarriers	64
IFFT and FFT Size	64
Length of Guard Interval	4 samples
Guard Type	Cyclic Extension
Modulation Type	16-QAM
Bandwidth	500 kHz
Pilot Ratio	1/8
Channel Taps	2
Channel Taps PDF	Exponential
Channel Model	Rayleigh fading

Table 6.1: Simulation Parameters

As mentioned earlier, we assume perfect synchronization since the aim is to observer channel estimation performance. Similarly, we consider pilot ratio of $\frac{1}{8}$, which is quite sufficient, and here our basic purpose is to compare different arrangements. For comparison purposes, same number of pilots are use in all schemes. Simulations are carried out for different signal to noise ratios and for different Doppler frequencies. For block-type pilot channel estimation, we assume that each block consists of a fixed number of OFDM symbols, which is 8 in our case. Pilots are sent at all subcarriers of the first symbol of each block and channel estimation is performed by using LS and LMMSE estimation. Channel estimated at the beginning of the block is used for all of the following symbols of the block. LMMSE estimation performs better than LS estimation, and according to [77], it gives improvement of $10 - 15$ dB in SNR. Simulation results are shown in Figure 6.4.

It is clear from Figure 6.4, for nominal value of BER for block-type pilot signal estimation, LMMSE estimation promises improvement of around $10 - 12$ dB over LS estimation.

In comb-type pilot signal estimation, channel is estimated at pilot frequencies by using LS estimation. After having estimates of channel at pilot frequencies, channel attenuations and phases at data locations can be obtained by using channel interpolation techniques, already discussed in Chapter 4. Channel is estimated at data symbols, by means of an interpolation filter, that uses the following interpolation techniques:

1. Linear Interpolation
2. Spline Interpolation
3. Cubic Interpolation
4. Low Pass Interpolation

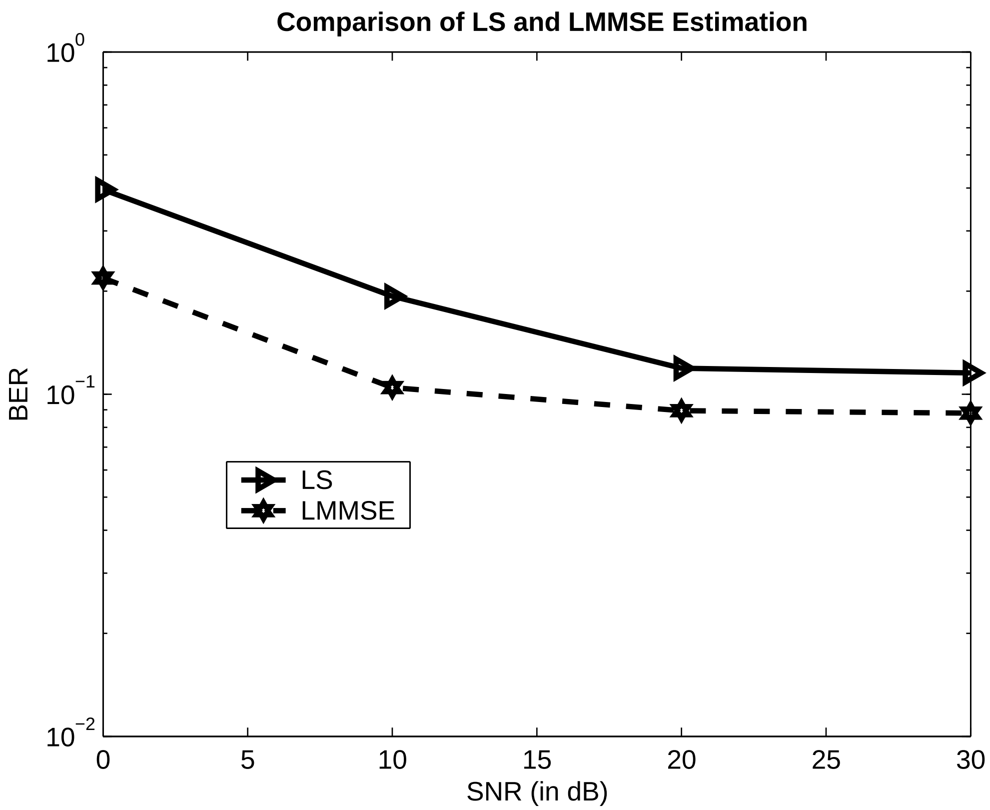

Figure 6.4: Comparison of LS and LMMSE estimation in block-type pilot Signal Estimation

Details of these interpolation methods are already presented in previous chapters. In proposed pilot pattern, we need two dimensional interpolation filter, because here pilots are spreaded in both directions i.e. time and frequency. Two dimensional interpolation filters are complex in structure, so standard technique used in multidimensional signal processing is to first interpolate in one direction, than interpolate in the other direction [77]. Concept of two dimensional interpolation is shown in Figure 6.5.

Figure 6.6 gives the BER performance of channel estimation algorithms for 2-ray Rayleigh fading channel described in Chapter 5, Doppler frequency 10 Hz, and OFDM parameters given in Table 6.1. From Figure 6.6, it is clear that among all possible pilot arrangements, proposed arrangement is best, in terms of BER, for slow varying channel. Simulation result shows that block-type estimation is around 10 dB higher than that of the comb estimation type. However, if we compare block-type estimation for different values of Doppler, from Figures 6.6, 6.7 and 6.8, it is clear that block estimation has good performance if channel is changing very slowly. This is an expected result, because at higher Doppler frequencies, channel transfer function changes so fast that there are even changes for adjacent OFDM symbols. So by observing simulation results, it can be easily verified, that block-type pilot arrangement performs well if channel is changing very slowly. In highly varying environments, channel is changing continuously, and block-type pilot patterns suffers higher values of BER.

Comb-type channel estimation with low pass interpolation achieves the best performance among all estimation algorithms for comb arrangements. The performance among comb-type channel estimation techniques usually ranges from best to worst as follows:

1. Low pass interpolation

2. Cubic Interpolation

3. Spline Interpolation

4. Linear Interpolation

These result are expected since the low pass interpolation used in simulations does the interpolation such that the mean square error between the interpolated values and there ideal values is minimized. From Figures 6.6, 6.7 and 6.8 it is clear that comb arrangement can track variations of fast varying channels. It is also consistent with the logical reasoning, because in comb arrangement pilots are transmitted in every OFDM symbol, and hence it allows tracking of fast fading channels.

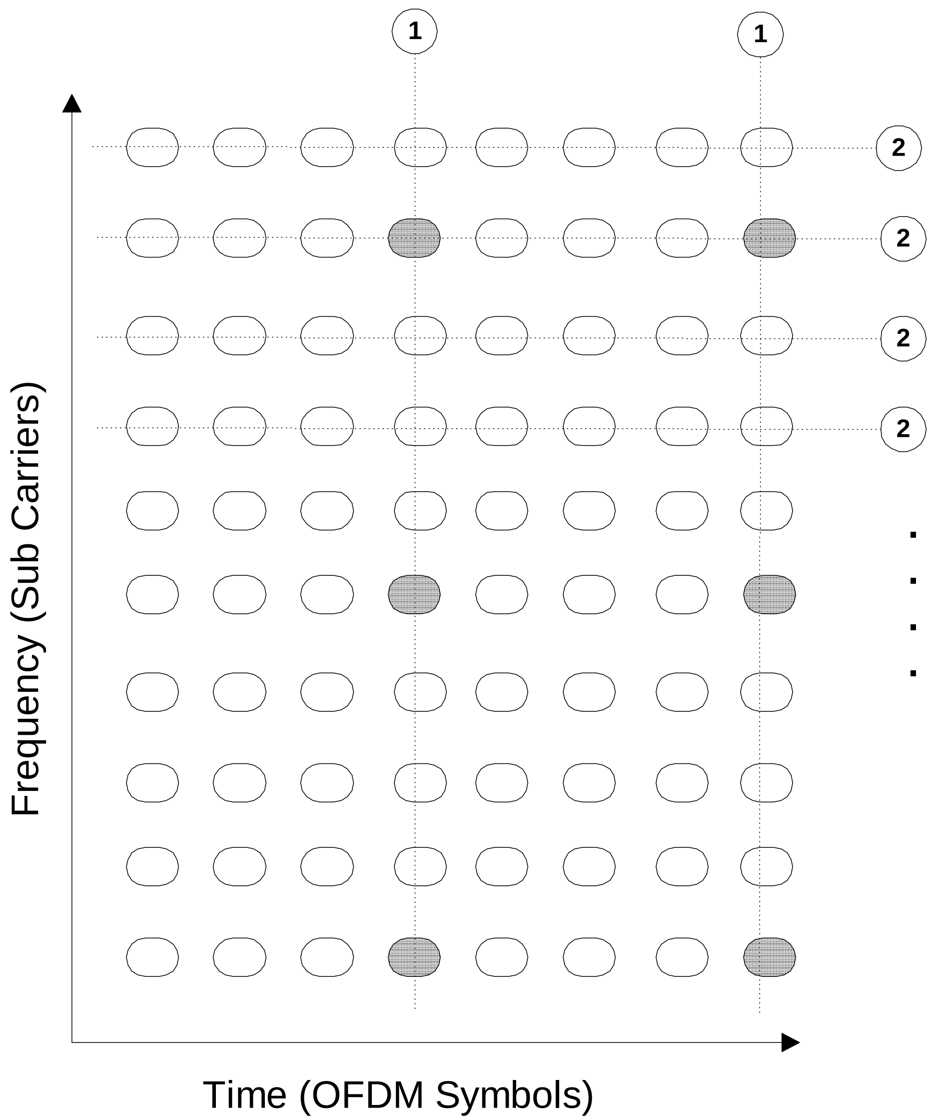

Figure 6.5: Interpolation in 2 dimensional OFDM grid

Another important observation from the simulation results is, that comb-type pilot estimation is less effected by Doppler frequency. BER should increase with increase in Doppler frequency, and BER increases but increment is very small as shown in Figure 6.9. For comb-type pilot estimation, increase in BER with Doppler frequency is very small because length of guard interval is small as compared to the number of subcarriers.

At lower values of Doppler frequencies, performance of the proposed scheme is better than any other pilot transmitting scheme, as shown in Figure 6.6. So it is recommended that in slow varying environments, it is much better to spread pilots symbols in time and frequency, instead of sending them along time or frequency separately. In slow varying channels, the proposed scheme works well for all values of SNR as shown in Figure 6.6. However, with the increase of Doppler frequency, proposed scheme works well only at higher values of SNR, as shown in Figure 6.7. For lower values of SNR, performance of lowpass-comb arrangement is better than equidistance arrangement. For highly varying channels, e.g. at Doppler frequency of 240 Hz, the performance of lowpass comb is better for all values of SNR, as shown in Figure 6.8.

From the above discussion, it is clear that proposed arrangement of pilot symbols is an excellent choice for lower values of Doppler frequencies. From Figure 6.9, the proposed pilot pattern is an optimum pilot transmission scheme in terms of BER, for Doppler frequencies less than 80 Hz. For typical carrier frequencies, in which

OFDM systems operates, channel with 80 Hz Doppler frequency is not that for a pedestrian channel. Figure 6.9 shows performance of channel estimation methods for 16-QAM modulation for different Doppler frequencies. The general behaviour of plots is that BER increases as the Doppler spread increases. The reason is the existence of ICI caused by Doppler shifts [6].

6.5 Adaptive Channel Estimation: Another Scheme

It is well known that Wiener filter provides an optimum solution for the filter weights in the mean square sense. A recursive approach to obtain the Wiener solution is the classical optimization method. This approach requires the use of a gradient vector which can be estimated in a deterministic or stochastic manner. The former is known as the method of steepest descent and the latter is known as stochastic gradient algorithm. The most common algorithm in the family of stochastic gradient algorithms is the least mean square (LMS) algorithm which is the subject of this section.

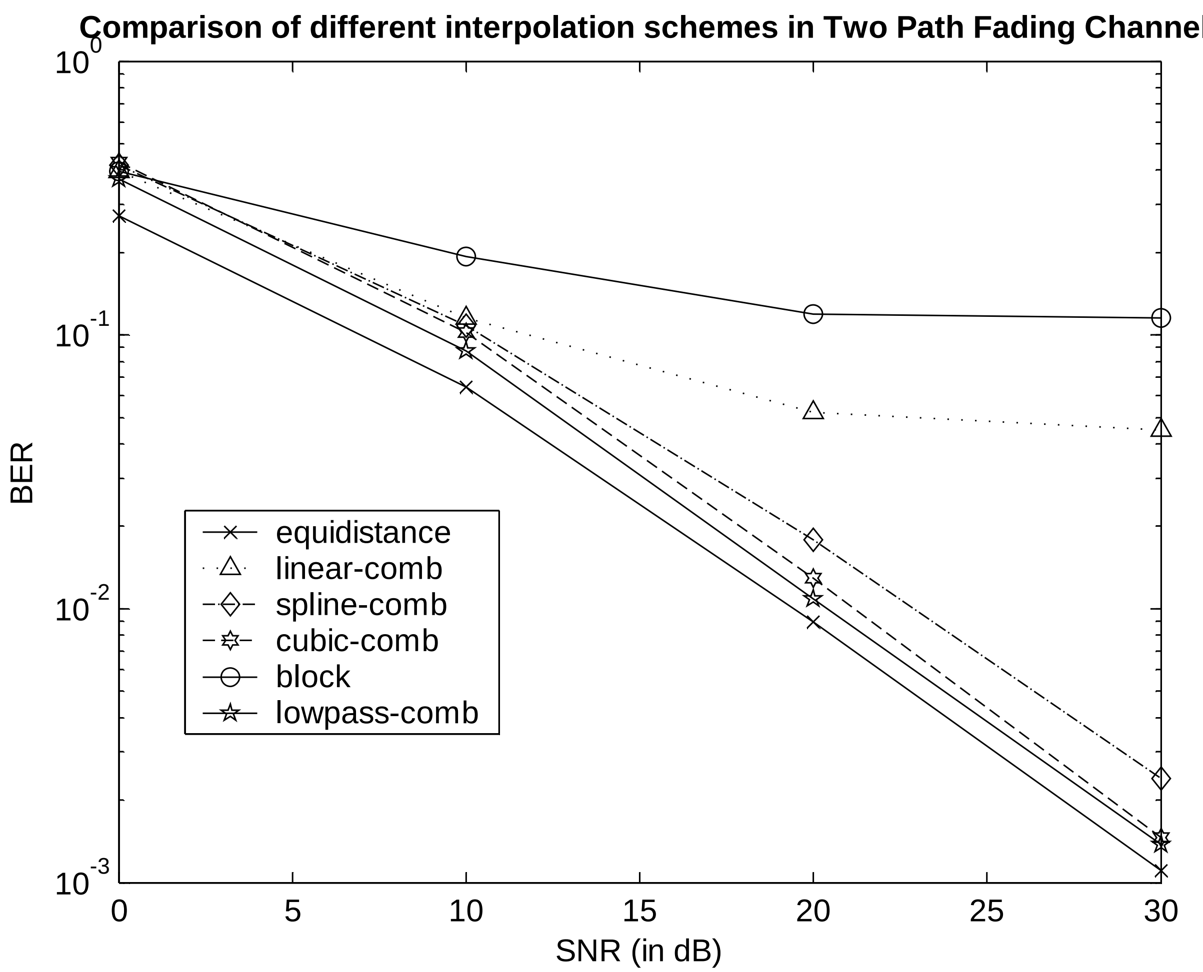

Figure 6.6: Comparison of channel estimation algorithms, for different pilot arrangements (Doppler freq. 10 Hz)

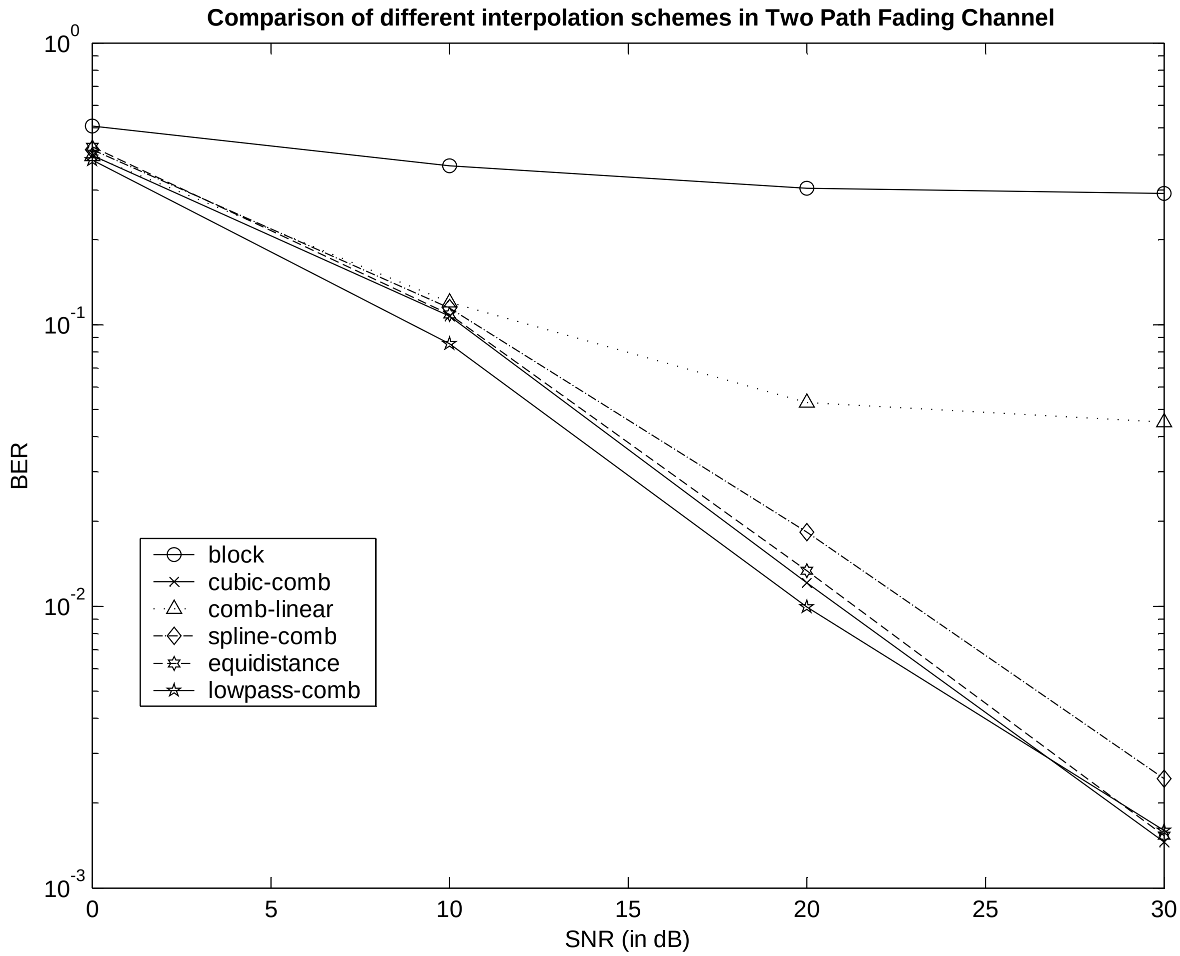

Figure 6.7: Comparison of channel estimation algorithms, for different pilot arrangements (Doppler freq. 70 Hz)

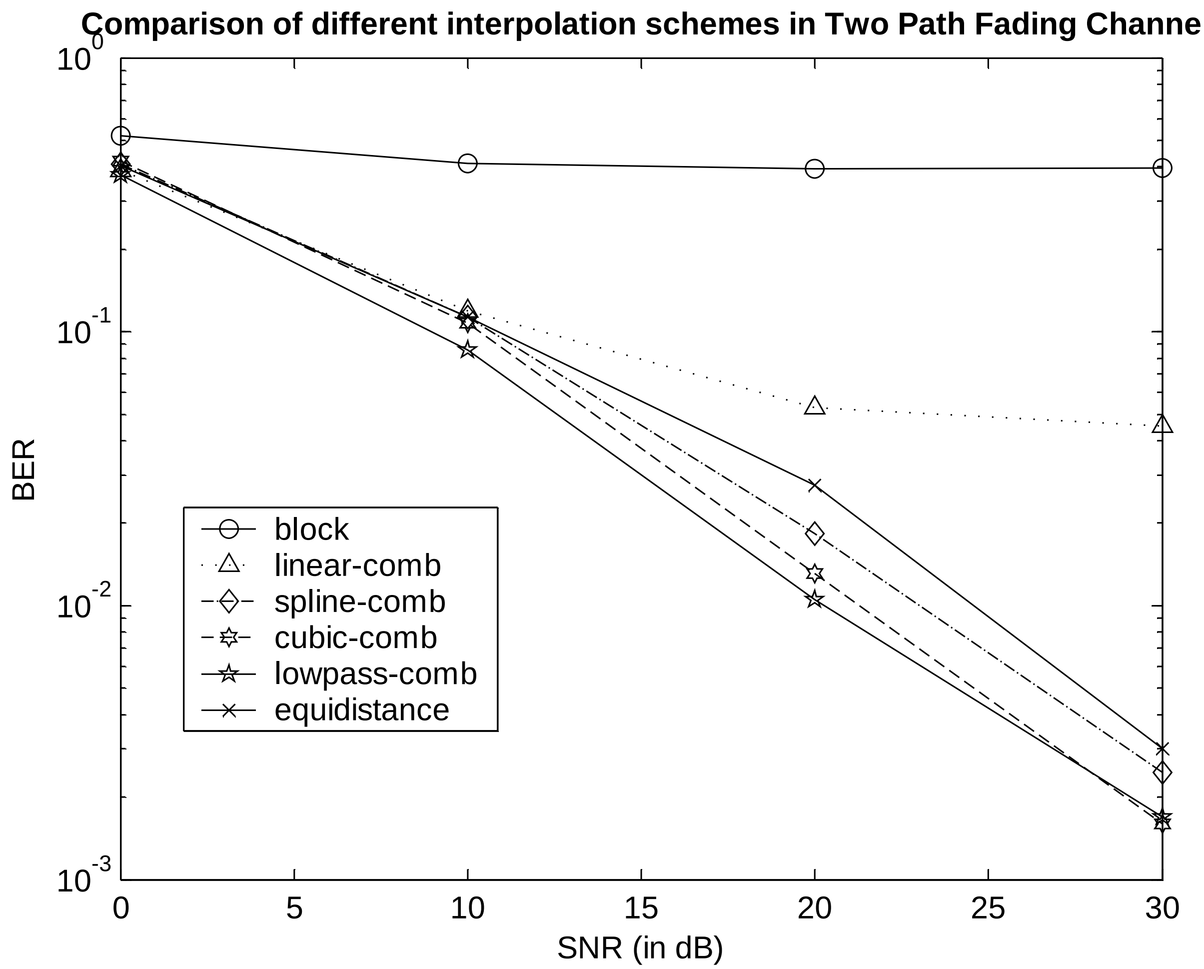

Figure 6.8: Comparison of channel estimation algorithms, for different pilot arrangements (Doppler freq. 240 Hz)

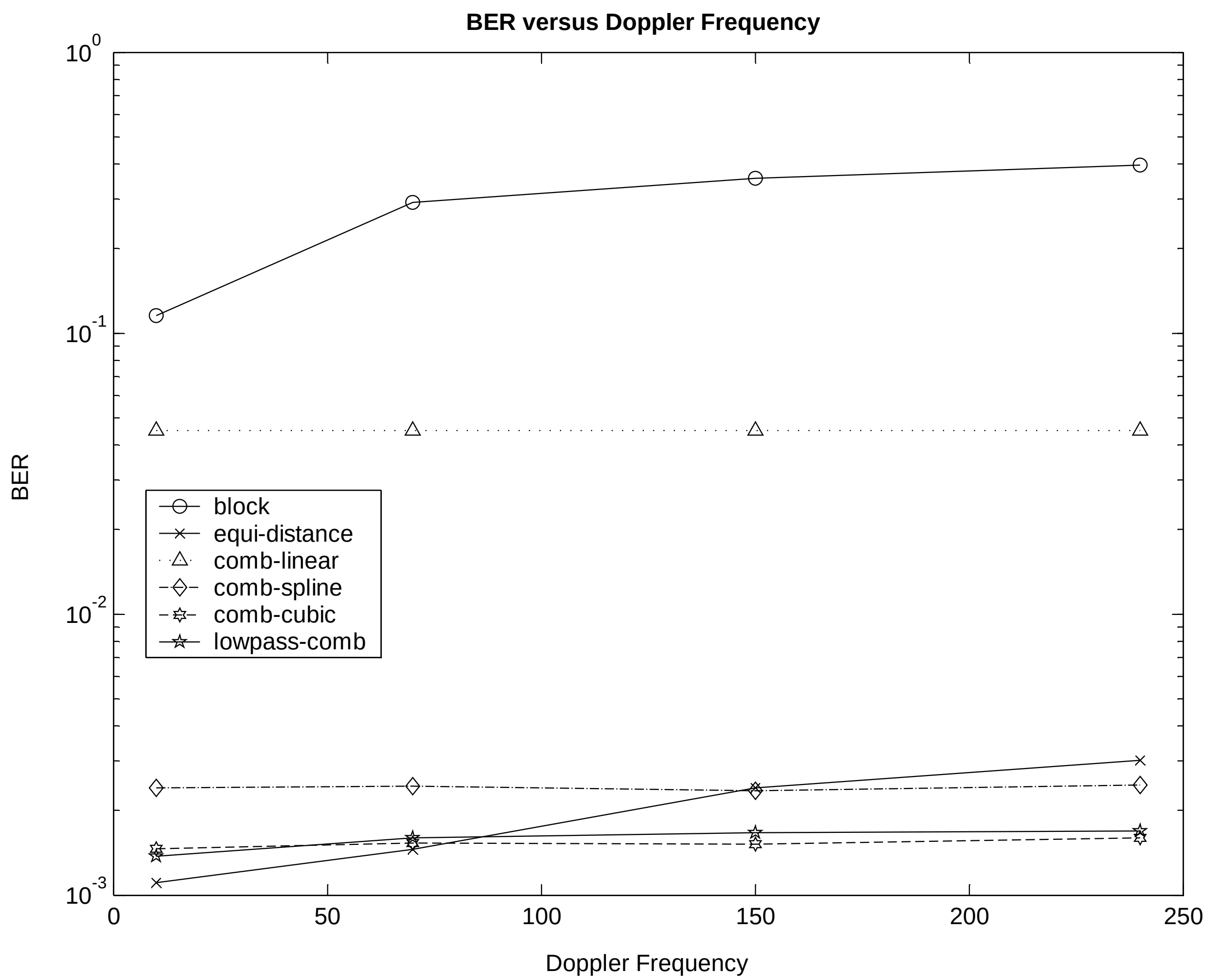

Figure 6.9: BER Versus Doppler Frequency

6.5.1 System Arrangement for Adaptive Channel Estimation

A block diagram of adaptive channel estimation in OFDM systems is shown in Figure 6.10. In the figure, we have unknown multipath fading channel, that has to be estimated with an adaptive filter whose weights or coefficients are updated based on some criterion so that coefficients of adaptive filter should be as close as possible to the unknown channel. The input to two systems, $u(n)$, are samples of known OFDM symbols (known to receiver). The output from the channel can be written as,

$$d(n) = \sum_{i=1}^{L} h_i(n)u(n - iT_s) + v(n) \tag{6.11}$$

where $v(n)$ is the sampled AWGN. Introducing vector notation, the received signal can be written as

$$d(n) = \mathbf{h}^H(n)\mathbf{u}(n) + \mathbf{v}(n) \tag{6.12}$$

where

$$\mathbf{h}(n) = [h_1(n),\ h_2(n),\ ...h_L(n),]^T$$

where $\mathbf{h}(n)$ is the vector of channel coefficients at time n, and $()^H$ stands for Hermitian transpose. The output of the adaptive filter is

$$\mathbf{y}(n) = \hat{\mathbf{h}}^H(n)\mathbf{u}(n) \tag{6.13}$$

where $\hat{\mathbf{h}}(n)$ is the vector of channel coefficients at time n. The error signal needed to update the weights of the adaptive filter is

$$
\begin{aligned}
e(n) &= d(n) - y(n) && (6.14)\\
&= \mathbf{h^H}(n)\mathbf{u}(n) + \mathbf{v}(n) - \hat{\mathbf{h}}^H(n)\mathbf{u}(n) && (6.15)
\end{aligned}
$$

For gradient based channel estimation, a general recursive equation to update the weights is

$$\hat{\mathbf{h}}(n+1) = \hat{\mathbf{h}}(n) + \mu - \nabla J(n) \tag{6.16}$$

where $J(n)$ is a cost function and $\nabla J(n)$ is the gradient vector with respect to $\mathbf{h}^*$. The most commonly used cost function is the Mean Square Error (MSE)

$$J(n) = E[|e^2(n)|] \tag{6.17}$$

reducing (6.16) to

$$\hat{\mathbf{h}}(n+1) = \hat{\mathbf{h}}(n) + \mu E[e(n)\mathbf{u}^*(\mathbf{n})] \tag{6.18}$$

6.5.2 Channel Estimation by LMS Algorithm

The LMS algorithm established itself as the workhouse of adaptive signal processing for two primary reasons [83]:

- Simplicity of Implementation and computational efficiency.
- Robust Performance

Beside this, it has some shortcomings, e.g. it is sensitive to the eigenvalue spread of the correlation matrix of the input signal. This problem does not cause any harm in channel estimation because the input data is usually white. However, this is not true in case of equalization where the input data to the adaptive filter is the channel output that introduces some correlation in the input signal. Thus, from the channel estimation prospective the eigenvalue spread of the input correlation matrix is usually close to 1. In fact, this is an ideal condition for LMS algorithm and justifies its suitability for this problem. Moreover, the speed of convergence of the LMS algorithm depends upon the proper selection of step size.

The application of LMS algorithm to track a time-varying system results in an error surface with time-varying minima. This the tracking algorithm has to track or approach towards a time-varying Wiener solution. If the LMS algorithm is used to track the channel then the weight update equation for the filter coefficients or taps is

$$\hat{\mathbf{h}}(n+1) = \hat{\mathbf{h}}(n) + \mu \mathbf{u}(n) e^*(n) \tag{6.19}$$

where instantaneous estimates are used as an stochastic approximation for the gradient vector of Equation (6.17).

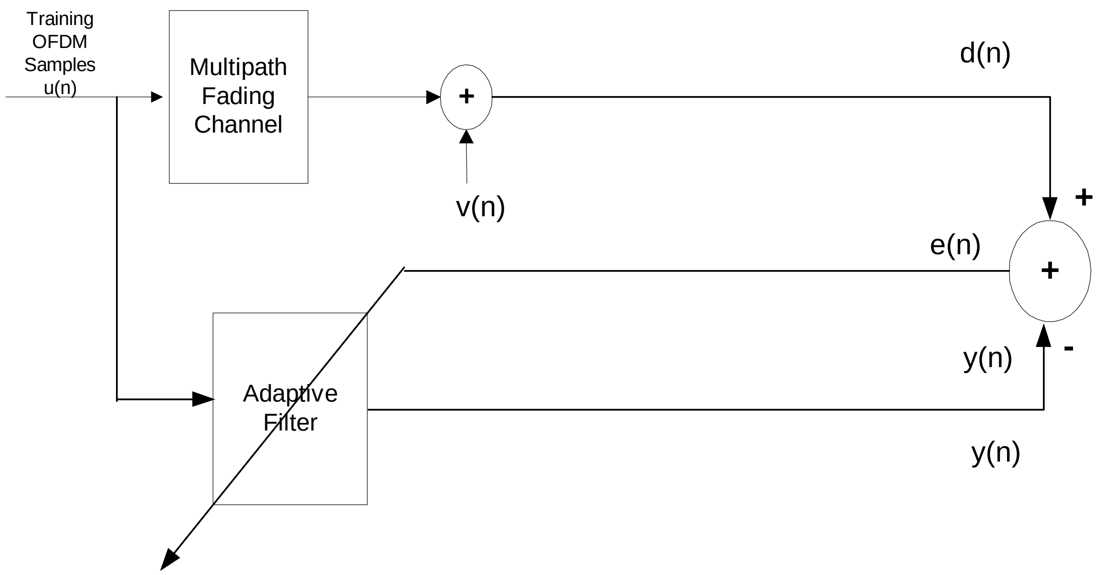

Figure 6.10: Scheme for Channel Tracking with Adaptive Filter

The Mean Square Error (MSE) is shown in Figure 6.11, in which we assume channel is changing slowly, so after sending some training data, filter converges, and then we can transmit OFDM symbols safely. If the channel is fast varying, than we need to transmit training data periodically, depending on how fast channel is changing.

6.6 Proposed Method for Compensation of Channel Phase

In an OFDM link, the subcarriers are perfectly orthogonal only if transmitter and receiver uses exactly the same frequency. Any frequency offset immediately results in ICI [28]. Before an OFDM receiver can demodulate the subcarriers, it has to perform at least few tasks, one of them is to estimate and correct carrier frequency offsets of the received signal.

There are also some timing offsets, which we are not discussing here, because we already assumed that transmitter and receiver are perfectly time synchronized. For single carrier systems, frequency offsets give only a degradation only in the received signal to noise ratio (SNR), rather than introducing interference. This is the reason that the sensitivity to frequency offsets are often mentioned as disadvantages of OFDM relative to the single carrier systems.

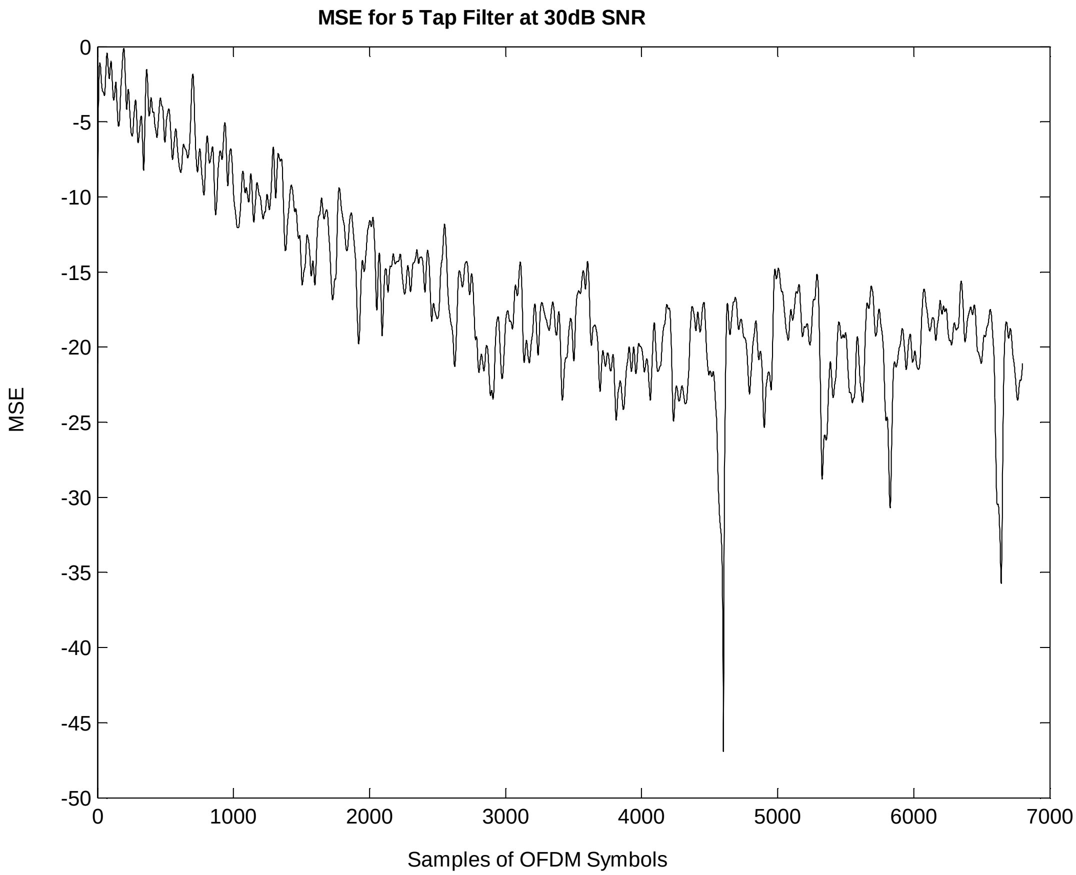

Figure 6.11: MSE for proposed adaptive channel estimation scheme

There are different techniques proposed in literature, that how this degradation can be kept to a minimum [28]. A most common technique to achieve frequency synchronization is by using cyclic prefix or special OFDM training symbols, as proposed by Fredrik in [84].

6.6.1 Sensitivity to frequency Offsets

In Chapter 2, we already explained that OFDM subcarriers are orthogonal if they have a different integer number of cycles within the FFT interval. If there is a frequency offset, than the number of cycles in the FFT interval is not an integer anymore, with the result that ICI occurs after the FFT. The most prominent effect of these frequency offsets is the increase in BER. The FFT output of each subcarrier will contain interfering terms from all other subcarriers, with an interference power that is inversely proportional to the frequency spacing [28].

We assumed that channel remains constant during one OFDM symbol, constant phase of the channel produces a drift in the frequencies of all subcarriers, as shown in Figure 6.12. This drift in frequencies causes ICI, and high error rates. In order to demonstrate the effect of channel phase on overall performance of the system, we simulate the system by assuming that channel produces only amplitude distortions in OFDM samples. Simulations results are shown in Figure 6.13. From the simulation results, it is very clear that in all pilot patterns, effect of channel phase is very dominant. If, in some way, we estimate the phase of the channel and then

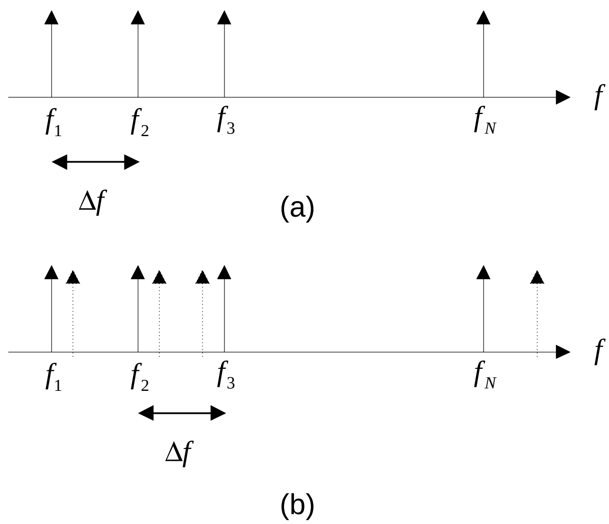

Figure 6.12: Drift in Subcarriers Frequencies due to channel Phase

compensate it before taking FFT, than the overall error performance of the system will be improved.

6.6.2 Proposed Algorithm

We assume channel is changing slowly, and estimate the phase of the channel by using an adaptive filter. We identify the channel, as shown in Figure 6.14. Adaptive filter adapts itself, by using an adaptive control algorithm [83] as the channel changes.

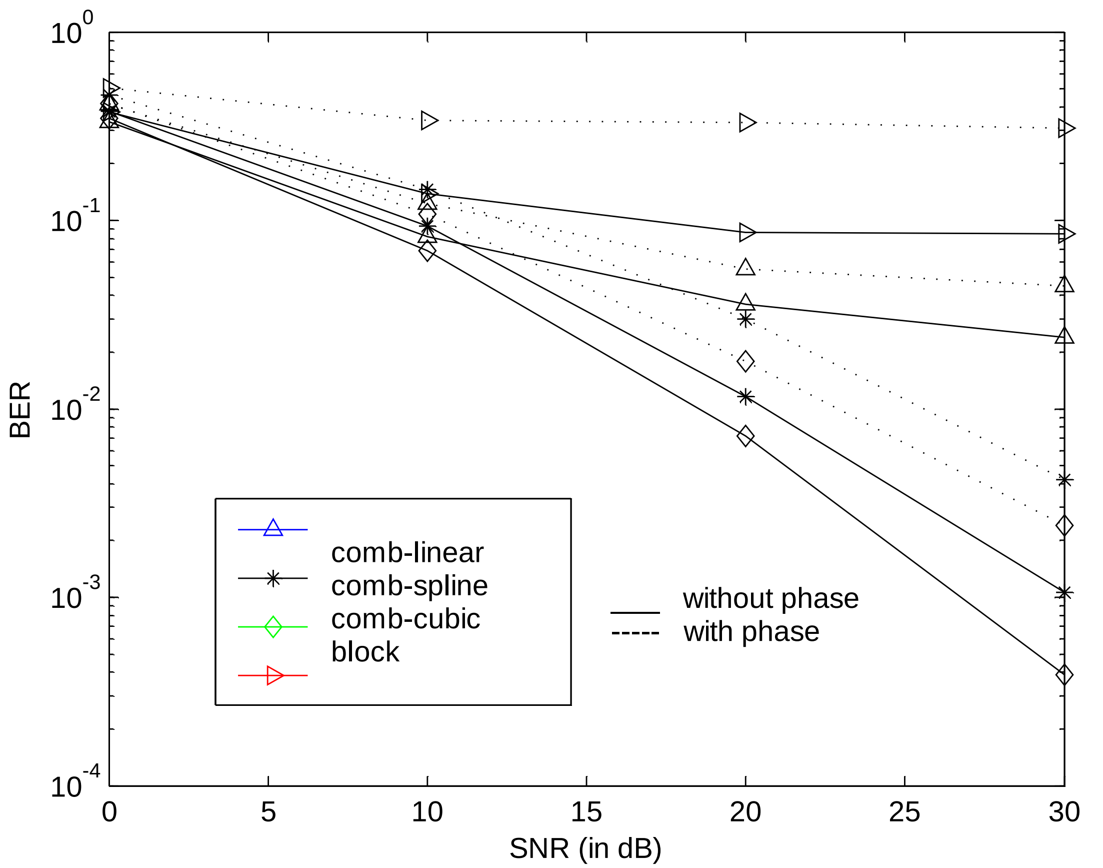

Figure 6.13: Effect of Channel Phase on BER performance

In this thesis, we are using LMS algorithm for adaptation, which is described in Equation (6.20).

$$\hat{\mathbf{w}}(n+1) = \hat{\mathbf{w}}(n) + \mu \mathbf{u}(n) e^*(n) \tag{6.20}$$

where

$$\begin{aligned} \hat{w}(n+1) &= Estimate\ of\ tap\ weight\ at\ time\ n+1 \\ \mu &= step\ size\ coefficient \end{aligned}$$

Simulation results shows that this method works well, if the channel is changing slowly. MSE for the filter taps coefficients is shown in Figure 6.15. From Figure 6.15, it is clear that adaptive filter converges, but with sufficient mean square error, this is a limitation of the proposed algorithm. Error rate is high, because we are using a single tap adaptive filter, which is not sufficient to identify the channel very accurately. So by designing the phase compensator in a more generalized way, we can reduce the MSE significantly, and overall performance improves. Here, we are just identifying the phase of channel, and in phase compensator, we compensate the phase of channel, simply by multiplying $e^{-j\phi}$ where ϕ is the phase of channel.

Performance improvement in BER, by applying proposed scheme of phase compensation, is shown in Figure 6.16 and Figure 6.17. From the simulation results, it is obvious that proposed scheme works well and give almost $5-6$ dB improvement in SNR.

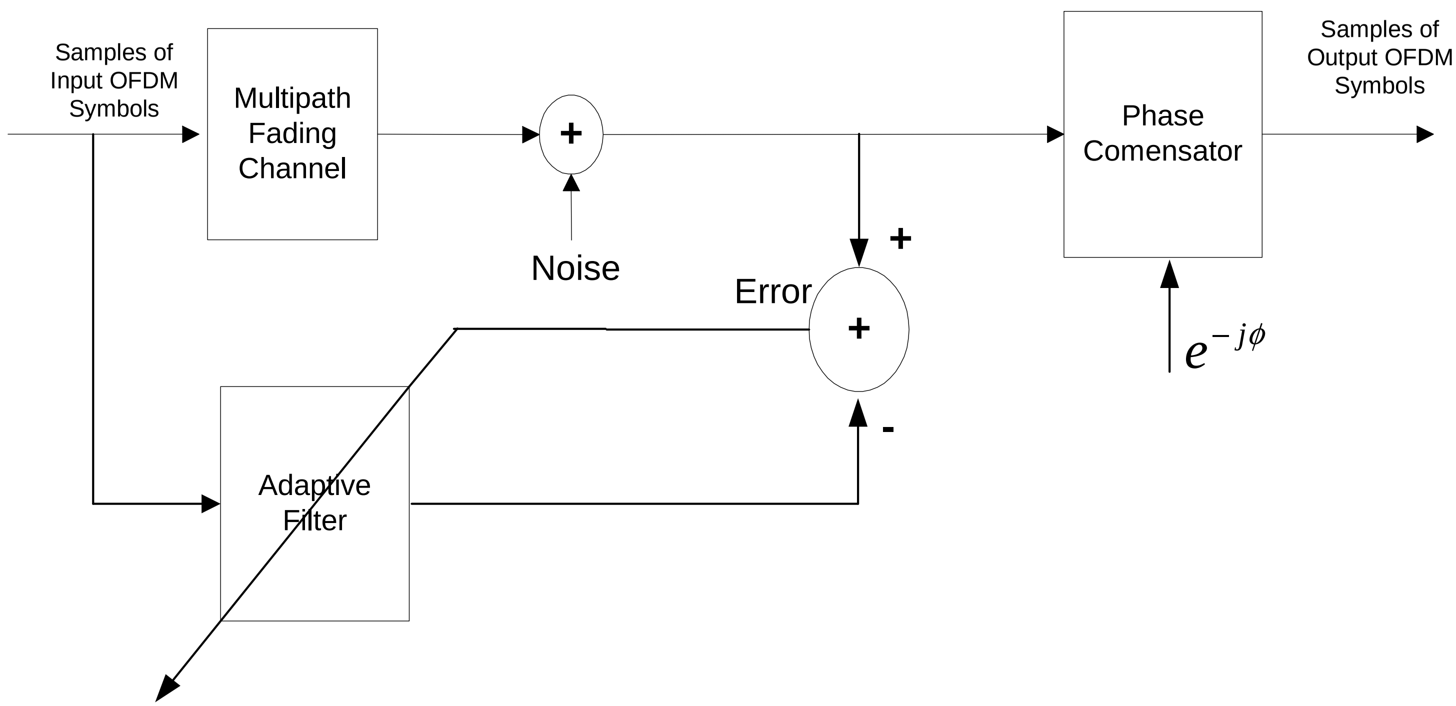

Figure 6.14: Proposed Scheme of Phase Compensation in OFDM Systems

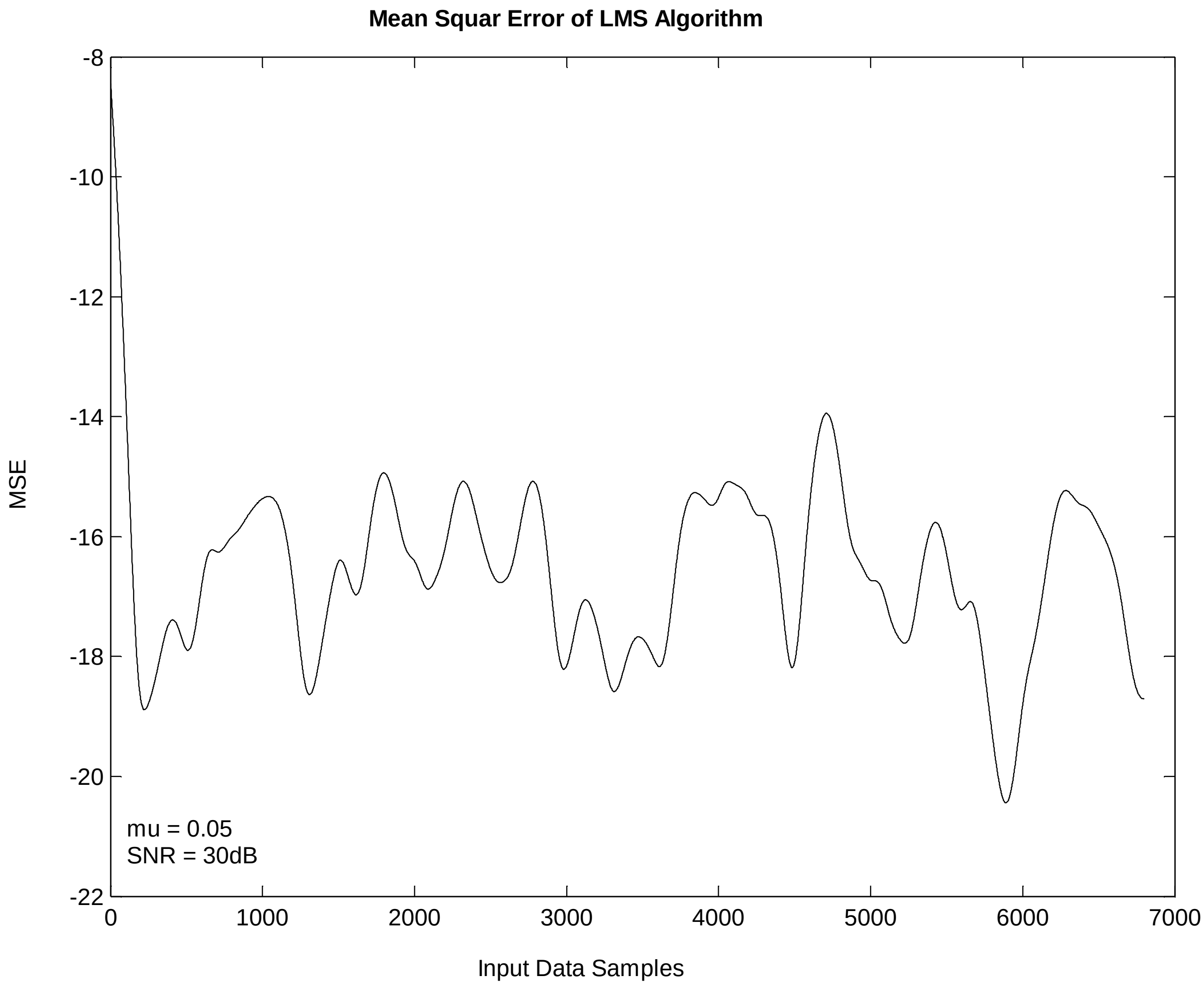

Figure 6.15: MSE curve of Adaptive Filter Taps

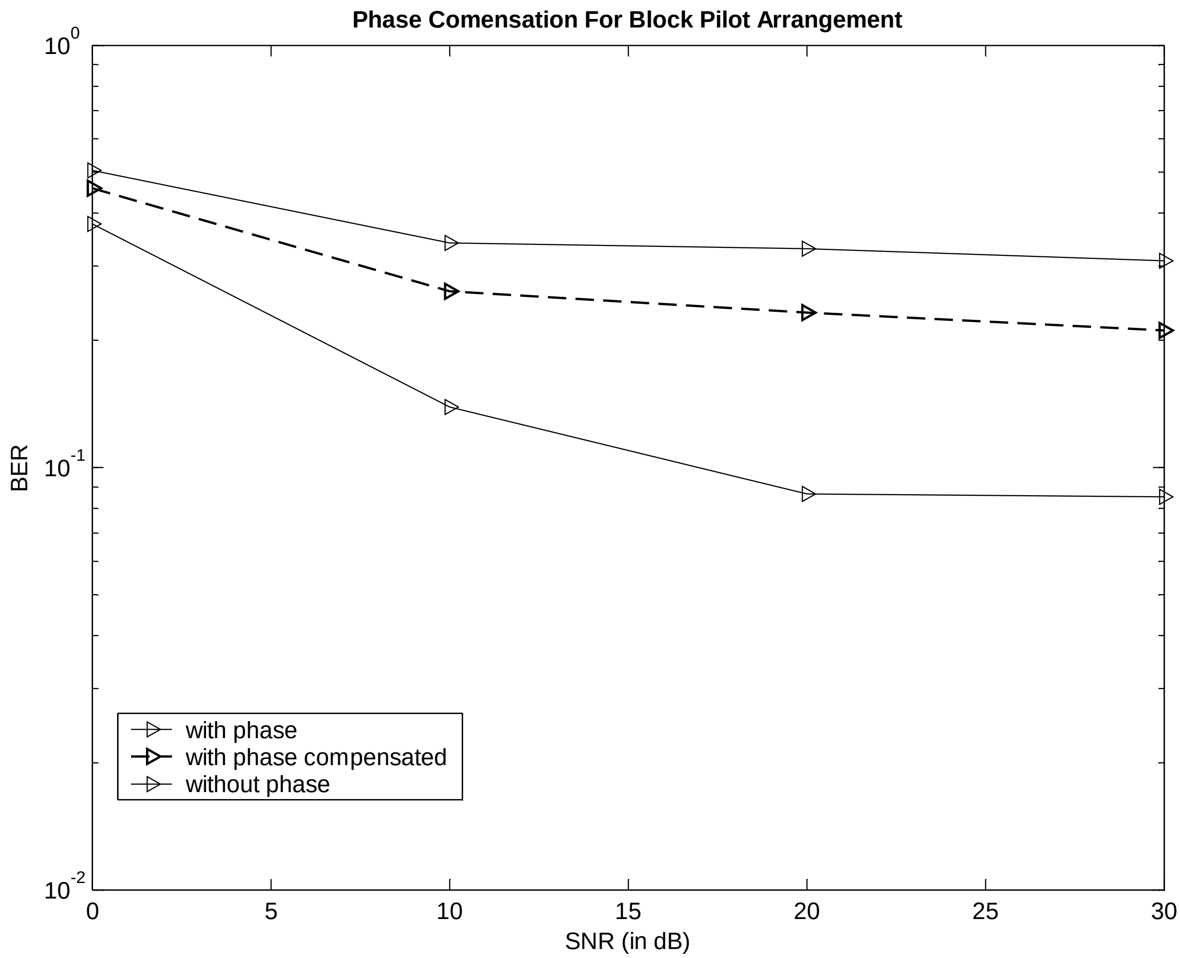

Figure 6.16: Improvement in BER for block arrangement

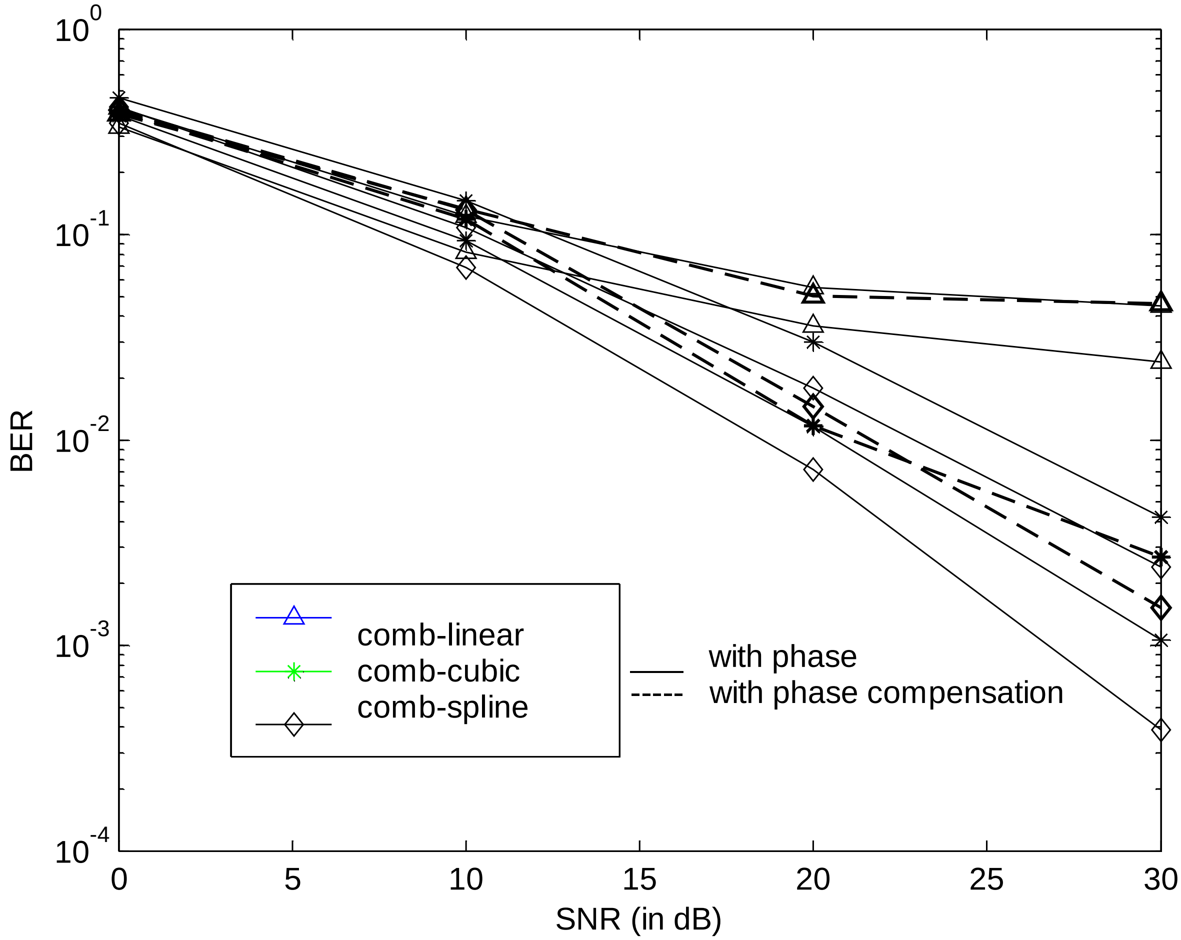

Figure 6.17: Improvement in BER for comb arrangement

Chapter 7

Future Work and Conclusion

7.1 Conclusions

Orthogonal Frequency Division Multiplexing (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, we investigated the performance of OFDM systems in detail, and via simulations we demonstrated effectiveness of Orthogonal Frequency Division Multiplexing, over single carrier systems. One of the major advantages of OFDM systems is its robustness against multipath delay spread of the channel. Hence, its typical applications are in tough radio environments. From the simulation results we discussed in Chapter 3, it is clear that if the length of guard interval is chosen properly, then OFDM systems exhibits robustness against multipath propagation, eliminates ISI and hence can be used for transmission at higher data rates. OFDM is also suitable for single frequency networks, since the signals from other transmitters can also be viewed as echoes i.e. multipath propagation. This means that it is favorable to use OFDM in broadcasting applications like Digital Audio Broadcasting (DAB) and Digital Video Broadcasting (DVB).

As in any digital communication systems, there are two alternatives for modulation: coherent or differential. The European DAM system uses differential QPSK, while the proposed scheme for DVB is coherent QAM. Differential PSK is suitable for low data rates and gives simple and inexpensive receivers, which is important for portable consumer products like DAB receivers. However, in DVB the data rate is quite high and low bit error rates are difficult to obtain with differential PSK. A natural choice for DVB is multiamplitude schemes. Due to the structure in OFDM, it is easy to design efficient channel estimators and equalizers. This is one of the appealing properties of OFDM which should be exploited to achieve high spectral efficiency. For coherent modulation, a dynamic estimation of channel is necessary for the demodulation of OFDM signals. In wideband mobile channels, pilot based signal correction schemes has been proven a feasible method for OFDM systems.

Channel estimation can be performed by many ways: either inserting pilot tones into all of the subcarriers of OFDM symbols with a specific period or inserting pilot tones into each OFDM symbol. In this thesis, we explored these two pilot arrangements in detail for different doppler frequencies. Channel estimation based on block type pilot arrangement is presented, and it is shown that this type of arrangement performs better when the channel is changing slowly. 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. It was shown that block-type channel estimation gives $10 - 15$dB higher than comb type channel estimation. This is expected because the structure of comb arrangement allows tracking of channel better than block arrangement. 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 again expected since the comb type channel estimation allows the tracking of fast fading channels and low pass interpolation does the interpolation in such a way that mean square error is minimized. We proposed a new scheme for the insertion of pilots, and call it equidistance pilot arrangement. Simulation results shows that equidistance arrangement works well for slow varying channel. However, for fast varying channels, performance of comb arrangement with low pass interpolation is much better than equidistance arrangement. So depending on the cell site, we have different pilot patterns and now the trick is how to choose a suitable pilot pattern, under different conditions of channel.

We also proposed an adaptive scheme for channel estimation in OFDM systems, which tracks the multipath fading channel by using LMS algorithm. In this scheme, we identified the channel by using an adaptive filter and simulation results shows that after sending some training OFDM symbols, the filter converges and then we can send OFDM data symbols, with sufficient knowledge of channel. One of the main disadvantage of OFDM systems is, that they are much sensitive to frequency offsets. Subcarriers will remain orthogonal if and only if there are no frequency offsets. When OFDM symbols passes through the channel, then the channel phase introduces frequency drifts in subcarriers. We propose a scheme to combat such frequency shits, estimate the phase of the channel, and then compensate it by using a phase compensator. Simulation results shows that the proposed scheme works fine if the channel is changing slowly.

7.2 Future Work

There are few suggestions regarding the future work. In the proposed equidistance scheme of pilot insertion, it was concluded that equidistance pattern works well for low Doppler frequencies upto around 80Hz, while at higher values of Doppler frequencies comb patterns with low pass interpolation works fine. 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.

Similarly, in the proposed scheme of phase compensation, we are limited with the use of single tap adaptive filter for estimating the phase of channel. This work can be extended to design a phase compensator in a more generalized manner, and use more taps of adaptive filter for better performance of the system.

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