SEMI-BLIND CHANNEL ESTIMATION FOR MULTIUSER OFDM-IDMA SYSTEMS

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December 2014

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Submitted in completion of the academic requirements for the degree of M.Sc. Engineering

School of Engineering: Electrical, Electronic and Computer Engineering

University of KwaZulu-Natal, Durban, South Africa

December 2014

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DECLARATION

I, Adisa J. Adeola, declare that:

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ii. This thesis has not been submitted for any degree or examination at any other University.

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Name: Professor Stanley H. Mneney

Signed: ________________________________

Date: ________________________________

DEDICATION

To the giver of knowledge

ACKNOWLEDGEMENT

All praise to God for His mercy endureth forever.

First and foremost, I would like to express my sincere appreciation and gratitude to my honorable, highly esteemed supervisor Professor Stanley Mneney, for his constant motivation and support during the course of my work in the last one year. I truly appreciate and value his fatherly role, esteemed guidance and encouragement from my first day to the end of this thesis.

My deepest appreciation to my exceptional co-supervisor, Dr. Olutayo Oyerinde. He has been a source of many ideas, enthusiasm, guidance and encouragement which made this research possible. Working with him has been my greatest pleasure and I am glad to have work under his co-supervision. I appreciate the time he spared for me.

I would like to express my gratitude to my colleague and friends here for their encouragement and support. I cherish all the wonderful time we had together. I am deeply indebted to Balogun Muyiwa, Alonge Akintunde, Adejumobi Babatunde and Sefunba Maria, for their eminent support and skills, imperative to the compilation of this dissertation. Thanks to all for the time and effort.

Through it all, I owe the greatest debt to my parents and siblings for the active role they took in my education. They have provided me invaluable support through their prayers and love.

My heartful appreciation goes to Mr Ibrahim Awal; you are the selfless man I have ever known, with the whole of my heart I say thanks and appreciate all your kind gestures.

All the wonderful people in my life, this thesis would have been impossible if not for their perpetual moral support. Thanks for your love and care.

Special appreciation to Telkom South Africa, Alcatel-Lucent and THRIP for the financial support through the Center of Excellence in the School of Engineering: Electrical, Electronics and Computer Engineering, University of KwaZulu-Natal, Durban. Special thanks to Dr Davidson and the entire Center of Engineering Postgraduate Study (CEPS) for the financial support.

ABSTRACT

Over the last decade, the data rate and spectral efficiency of wireless mobile communications have been significantly enhanced. OFDM technology has been used in the development of advanced systems such as 3GPP LTE and terrestrial digital TV broadcasting. In general, bits of information in mobile communication systems are conveyed through radio links to receivers. The radio channels in mobile radio systems are usually multipath fading channels, which cause inter-symbol interference (ISI) in the received signal. The ability to know the channel impulse response (CIR) and Channel State Information (CSI) helps to remove the ISI from the signal and make coherent detection of the transmitted signal at the receiver end of the system easy and simple. The information about CIR and CSI are primarily provided by channel estimation.

This thesis is focused on the development of multiple access communication technique, Multicarrier Interleave Division Multiple Access (MC-IDMA) and the corresponding estimation of the system channel. It compares various efficient channel estimation algorithms. Channel estimation of OFDM-IDMA scheme is important because the emphasis from previous studies assumed the implementation of MC-IDMA in a perfect scenario, where Channel State Information (CSI) is known.

MC-IDMA technique incorporates three key features that will be common to the next generation communication systems; multiple access capability, resistance to multipath fading and high bandwidth efficiency.

OFDM is almost completely immune to multipath fading effects and IDMA has a recently proposed multiuser capability scheme which employs random interleavers as the only method for user separation. MC-IDMA combines the features of OFDM and IDMA to produce a system that is Inter Symbol Interference (ISI) free and has higher data rate capabilities for multiple users simultaneously. The interleaver property of IDMA is used by MC-IDMA as the only means by which users are separated at the receiver and also its entire bandwidth expansion is devoted to low rate Forward Error Correction (FEC). This provides additional coding gain which is not present in conventional Multicarrier Multiuser systems, (MC-MU) such as Code Division Multiple Access (CDMA), Multicarrier-Code Division Multiple Access (MC-CDMA) systems, and others. The effect of channel fading and both cross-cell and intra-cell Multiple Access Interference (MAI) in MC-IDMA is suppressed efficiently by its low-cost turbo-type Chip-by-Chip (CBC) multiuser detection algorithm.

We present the basic principles of OFDM-IDMA transmitter and receiver. Comparative studies between Multiple Access Scheme such as Frequency Division Multiple Access (FDMA), Time Division Multiple Access (TDMA), CDMA and IDMA are carried out.

A linear Minimum Mean Square Error (MMSE)-based estimation algorithm is adopted and implemented. This proposed algorithm is a non-data aided method that focuses on obtaining the CSI, remove ISI and reduce the complexity of the MMSE algorithm. However, to obtain a better and improved system performance, an improved MMSE algorithm and simplified MMSE using the structured correlation and reduced auto-covariance matrix are developed in this thesis and proposed for implementation of semi-blind channel estimation in OFDM-IDMA communication systems. The effectiveness of the adopted and proposed algorithms are implemented in a Rayleigh fading multipath channel with varying mobile speeds thus demonstrating the performance of the system in a practical scenario. Also, the implemented algorithms are compared to ascertain which of these algorithms offers a better and more efficient system performance, and with less complexity. The performance of the channel estimation algorithm is presented in terms of the mean square error (MSE) and bit error rate (BER) in both slow fading and fast fading multipath scenarios and the results are documented as well.

TABLE OF CONTENTS

LIST OF ABBREVIATIONS

1G	First Generation Network
2G	Second Generation Network
3G	Third Generation Network
4G	Fourth Generation Network
ADC	Analog-to-Digital Converter
ADSL	Asymmetrical Digital Subscriber Line
AMPS	Advanced Mobile Phone Systems
APP	*a Posteriori Probability*
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
BS	Base Station
CA	Code Aided
CATV	Community Antenna Television
CDMA	Code Division Multiple Access
CFO	Carrier Frequency Offset
CIR	Channel Impulse Response
CP	Cyclic Prefix
CSI	Channel State Information
CTF	Channel Transfer Function
DA	Data Aided
DAB	Digital Audio Broadcasting
DAC	Digital-to-Analog Converter
DDCE	Decision Directed Channel Estimation
DEC	Decoders
DFT	Discrete Fourier Transform
DS-SS	Direct Sequence Spread Spectrum
DVB –T	Terrestrial Digital Video Broadcasting
ESE	Elementary Signal Estimator
EVD	Eigen Value Decomposition
FDM	Frequency Division Multiplexing

FDMA	Frequency Division Multiple Access
FEC	Forward Error Correction
FFT	Fast Fourier Transform
GSM	Global System for Mobile Communication
HDSL	High Speed Digital Subscriber Line
HOS	Higher Order Statistical
HPA	High Power Amplifier
ICI	Inter Channel Interference
IDFT	Inverse Discrete Fourier Transform
IDMA	Interleave Division Multiple Access
IFFT	Inverse Fast Fourier Transform
IMT	International Mobile Telecommunications
ISI	Inter Symbol Interference
ITU	International Telecommunication Union
LAN	Local Area Network
LLR	Logarithm Likelihood Ratio
LMS	Least Mean Square
LO	Local Oscillator
LOS	Line of Sight
MAC	Multiple Access Channel
LTE	Long Term Evolution
MAS	Multiple Access Scheme
MAI	Multiple Access Interference
MAN	Metropolitan Area Network
MC	Multicarrier
MMAC	Multimedia Mobile Access Network
MMSE	Minimum Mean-Squared Error
MSE	Mean Square Error
MU	Multicarrier
MUD	Multiuser Detection
NDA	Non Data Aided
NMT	Nordic Mobile Telephone
OFDM	Orthogonal Frequency Division Multiplexing
PAPR	Peak-to-Average Power Ratio
PDC	Personal Digital Cellular

PN	Pseudo random Noise
QOS	Quality of Service
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
SOS	Second Order Statistical
TACS	Total Access Communication System
TDMA	Time Division Multiple Access
VLSI	Very Large Scale Integration

LIST OF NOTATIONS

C_B	Channel coherent bandwidth
τ_d	Delay spread
λ	Signal wavelength
f_d	Doppler Shift
d	Distance
v	Mobile velocity
$\Delta\emptyset$	Phase change
l	Path length
D_S	Doppler Spread
T_C	Coherence Time
P_R	Received power
P_T	Transmitted power,
f	Carrier frequency
G	Power gain
P_l	Path loss
B_S	Signal bandwidth
T_S	Symbol duration
B_D	Total available bandwidth
B_{GD}	Guard band
P_G	Process gain
S_{BW}	Spread spectrum bandwidth
S_{TW}	Transmitted signal bandwidth
T	Sampling period
T_G	Length of cyclic prefix
SC	Subcarriers
f_{max}	Maximum frequency
$x(t)$	Transmitted signal
δf	Frequency shift
θ	Signal phase offset
D_S	Doppler Spread
α	Inversely proportional
βn	Propagation paths amplitude
$h_k(n)$	Fading channel coefficient

d(n)	Additive White Gaussian Noise
$ב_k(n)$	Multiuser interference
$r(n)$	Received signal
N	Number of sub-carriers
^{H}	Hermitian transpose
$E\{\cdot\}$	Expected value
$Var\{\cdot\}$	Variance
$Re\{\cdot\}$	Real part

LIST OF FIGURES

LIST OF TABLES

CHAPTER 1

INTRODUCTION

1.1 Evolution of Wireless Communication Systems

There was an explosive growth and high demand for high quality voice, data and video services in the 20th century. The carrier of these services has largely been wireless links. Wireless communication services thus made a way through into our society in an unprecedented scale. In recent times, there has been an increase in the demand for data transmission and bandwidth efficiency in wireless transmission due to the phenomenal and exponential growth of wireless communication. The demand in communication calls for technologies that will make efficient use of the limited available electromagnetic resources and provide high quality of service (QoS) and broadband data access in the most rational way. In the past decade, new methods and products for wireless communication systems have resulted from gradual development of different technologies that evolved in the public, military and commercial sectors, sharing the available radio resources.

A chronological evolution of wireless communication systems is summarised as follows:

1.1.1 First Generation (1G) Wireless Communication Systems

The last few decades have seen an increase in the amount of information transmitted over the air. Transmission of voice signals employing analog frequency division multiple access (FDMA) was used in first generation of wireless telecommunication technology. The standard used in this early system varies from one country to another. The United State of America used the Advanced Mobile Phone Service (AMPS), while United Kingdom and Scandinavia used the Total Access Communication System (TACS) and the Nordic Mobile Telephone (NMT) respectively, [1]. Each user in this generation was assigned a unique frequency band and the signals from different users were separated in the frequency domain using the multiple access scheme called FDMA. Though data security is extremely important in any wireless communication but in the first generation systems, there was compromise in data security because advanced encryption methods were not authorised. Only a simple technique could be utilised to intercept conversations.

1.1.2 Second Generation (2G) Wireless Communication Systems

In the second generation system, there was improvement in technologies that supported Integrated Circuits (IC). Digital technology was adopted and it was more economical and practical than its predecessor that utilised the analog technology. Code Division Multiple Access (CDMA) and Time Division Multiple Access (TDMA) are the multiple access digital technologies used in the second generation systems. The second generation system supported speech service and low data-rate, data services. It provided better data services, high spectral efficiency, more advanced roaming and 3-times increase in user capacity. In 1991, the second generation standard was used in the production of the first digital cellular service and its commercial operation commenced in Europe [2]. Many of the cellular systems used at present time depend on 2G.

1.1.3 Third Generation (3G) Wireless Communication Systems

October 1st 2002, mark the inception of the operation of 3G communication system in Japan. The 2G complexity was augmented by the 3rd generation system. When compared to 1G, it achieved 10-fold increase in system capacity by using CDMA or TDMA which are more complex multiple access techniques. Improved quality images and video stream transmission, data services with expeditious internet access are the 3G standards supported services. The International Telecommunication Union (ITU) released an interdependent set of recommendations which were used to construe International Mobile Telecommunications-2000 (IMT-2000), that served as the global standard for 3G wireless communication. By linking many varied systems' based networks, IMT-2000 provided access for worldwide wireless communication. It combined current services and networks with the aim to achieve a unified generation network, as shown in fig 1.1.

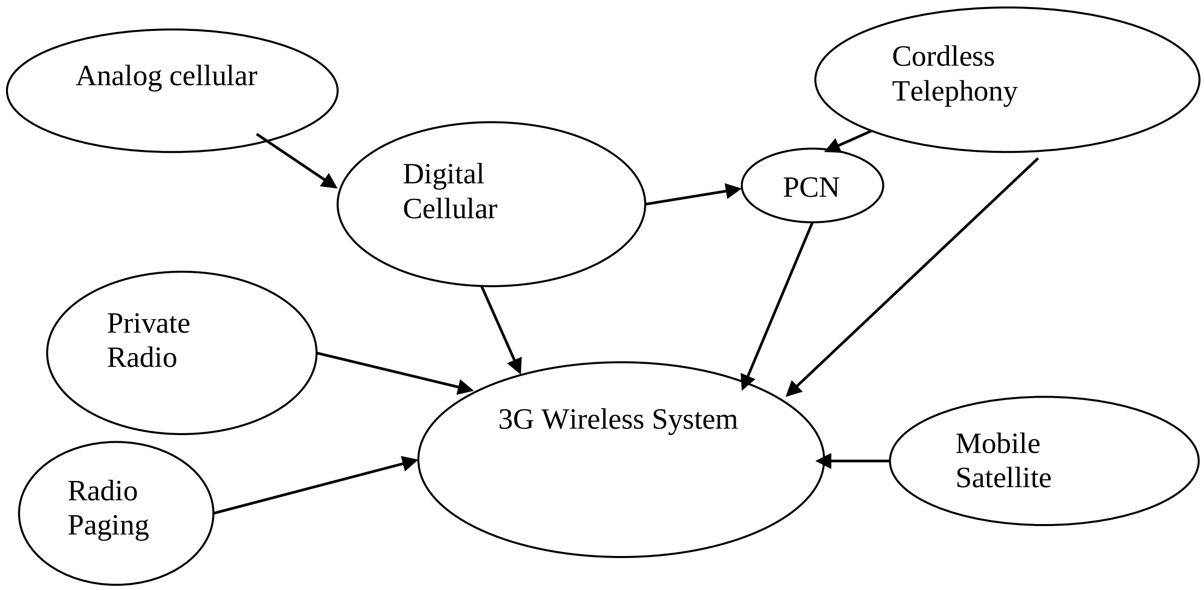


Figure 1.1: A unified third generation network.

1.1.4 Fourth Generation (4G) Wireless Communication Systems

Researchers and industries triggered by the evolution of new technologies and the ever increasing demand for high data rate services by users in the mobile systems, came up with an exhaustive illustration of the imminent fourth generation (4G) mobile communication system even before the deployment of 3G systems. The recommendation of ITU-R M.1645 “Framework and overall objectives of the future development of IMT-2000 and systems beyond IMT-2000” was approved by ITU in June 2003 [3]. 4Gs essential characteristics include:

- The support of data rates for high mobility up to 100 Mbps and low mobility up to 1 Gbps for mobile access and local access, respectively.
- The increase in system security and reliability because of its packet based architecture.
- The optimization of its spectrum and interoperability capabilities.
- The fast and high data rate, improved network capacity and it is absolute integration with wired line backbone networks.

Several challenges in communication system are yet to be overcome in spite of the 4G essential characteristic. These include the request for high transmission rate for data (up to 1Gbps), quality of service (QoS) management, a perfect global roaming, high user capacity, incorporation of 3G and next-generation components and there compatibility. High bit rate data transmission is required for high quality audio, video, and mobile integration in this current and future mobile communication system. Data transmission over a radio channel at very high bit rates can cause the Channel Impulse Response (CIR) to spread across several symbol periods, thus causing Inter Symbol Interference (ISI) in the system. ISI can be mitigated using a multicarrier system, Orthogonal Frequency Division Multiplexing (OFDM), which will be discussed later.

The research on 4G communication systems specifically focused on spectral efficiency and system complexity. Thus different multiple access schemes were considered to achieve this goal. OFDM-based multiple access schemes and associated hybrid technologies have become popular and have been the focus of recent mobile communications research due to their inherent advantages which include efficient and reliable high data-rate transmission as well as low system complexity.

1.2 Fundamental of the Wireless Channel

The wireless communication channel has undesirable effects on transmitted signals that pass through it due to changes of its physical properties. The transmission through the radio channel propagation contains multipath reception. The complexities of the radio channel environment interact with the transmitted signal. These complexities arise from the physical properties of the channel. The distorted, delayed and phase shifted received signal is the result of the combination of the effect of scattered buildings and other structures, reflected, diffracted, and scattered copies of the signal transmitted that arrive at different times with differing attenuation level at the receiver. Figure 1.2 shows the effect of reflection and diffraction on the received signal. Reflection takes place if the signal hits an object with a relatively large surface compared to the wavelength of the signal. Diffraction takes place when the transmitted signal encounters an opaque body with a larger surface compared to the wavelength of the signal. It then leads to bending of a wave around an obstacle. Scattering happens if the transmitted signal strikes an object, and the size of the object compared to the signal wavelength are the same or the signal wavelength is smaller. A perfect understanding of the fundamental characteristics and challenges confronting radio channel systems helps in the design and implementation of cellular communication schemes. The parameters that characterize wireless channels are discussed below.

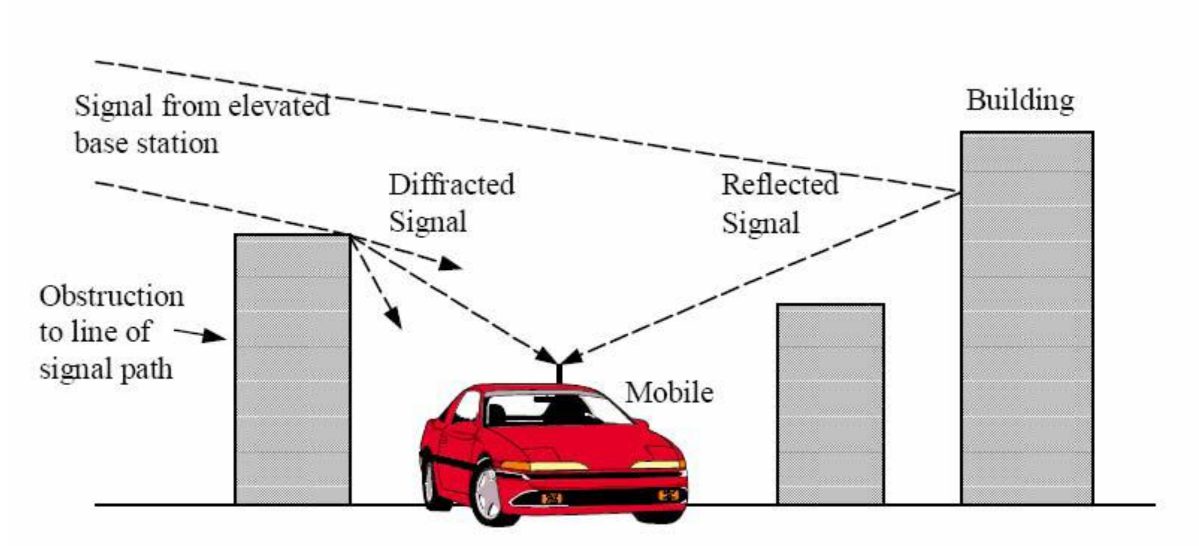


Figure 1.2: Effect of Reflection, Refraction and Diffraction on signal.

1.2.1 Parameters of Fading Channels

1.2.1.1 Delay Spread

Delay spread (τ_d) describes the inherent feature of the channel according to the time variance. As stated earlier, the received signal is the combination of the effect of reflections of object like hills, mountains, buildings and other infrastructures along the transmission path. It takes an extra time for the reflected signals to arrive at its destination due to the extra path lengths they follow compared to the direct signal. There is thus a slight difference in the time of arrival for the multipath signals and as a result, the received energy is spread. Delay spread can be described as the maximum time that occurs between the longest and the shortest multipath propagation arrivals time of the received signal. This "delay spread" is known to be the spread in arrival times, and it causes symbols of the transmitted signal to overlap thus producing ISI in mobile communication. The performance of a system is degraded by the presence of ISI in the system and thus increases the Bit Error Rate (BER) of the system. Delay spread is a parameter that determines the coherence frequency and must be taken into consideration when designing a communication system.

1.2.1.2 Coherence Bandwidth

Coherence bandwidth (C_B) depends upon fading parameter which is the delay spread. The frequency of the channel response is measured statistically by the coherence bandwidth and it is inversely proportional to delay spread and is given by the mathematical expression [4]

$$C_B = \frac{1}{2\tau_d}, \tag{1.1}$$

where, τ_d represents the channel delay spread. Frequency-selective characteristics and the flatness of a communication channel is determined by the coherence bandwidth. *Rappaport et al* described coherence bandwidth as the frequency range where the amplitude of the two frequency components of the radio channel becomes correlated [5].

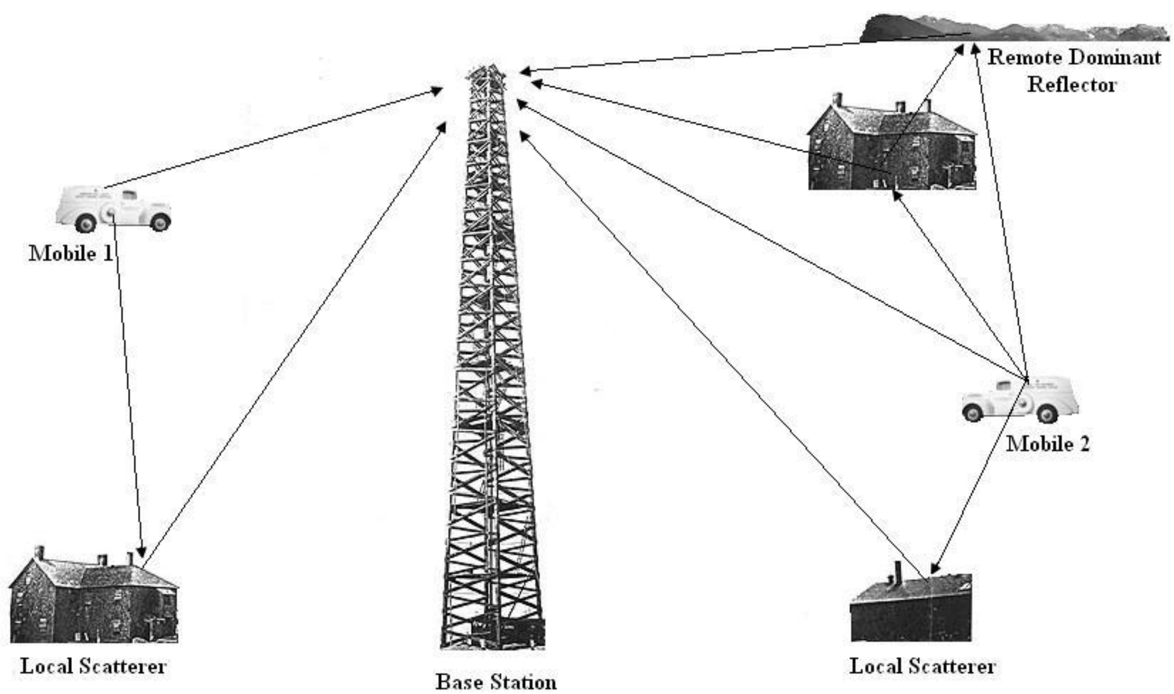


Figure 1.3: Multipath propagation in wireless communication channel [6].

1.2.1.3 Doppler Shift

In the doppler shift (f_d), the frequency of the originally transmitted signal differs from the frequency of the received signal. This is caused by the relative motion that exists in the mobile channel between the transmitter and the receiver. The changes in the transmitted and received signal frequencies is called Doppler shift. In Figure 1.4, a mobile subscriber moving at velocity v receives a signal from a remote Base Station (BS) S. The wave that travelled from the source S to the mobile at point x and y has a relative difference in path length $l = d\ cos\emptyset$, where the distance $d = v\Delta t$, the time required for the mobile to move from point x to y is Δt and λ is the signal wave length. The difference in path length causes a change in phase, and [7] gives the phase change as:

$$\Delta\emptyset = 2\pi l/\lambda = \frac{2\pi d cos\emptyset\ .}{\lambda} \tag{1.2}$$

Hence, the apparent Doppler shift or frequency change is given as

$$f_d = \frac{1}{2\pi}\ X \frac{\Delta\emptyset}{\Delta t} = \frac{v}{\lambda} cos\emptyset\ . \tag{1.3}$$

If the mobile, using a particular communication technique, has high relative speed, then the carrier is susceptible to carrier frequency offsets caused by Doppler shift. This is the case for communication systems using OFDM. The relative movement between a transmitter and a receiver can either be away or toward each other. The movement of the transmitter and receiver towards each other causes each wave to take lesser time to reach the observer. The arrival time of the wave at the observer is reduced, thereby causing an increase in the frequency, hence the Doppler frequency shift has a positive value. On the contrary, if the transmitter is moving away from the receiver, there is reduction in frequency because the receiver is positioned farther away thus increasing the arrival time. In this case the doppler frequency shift has a negative value. This phenomenon is called the Doppler's effect and it becomes significant when developing mobile radio systems [7, 8].

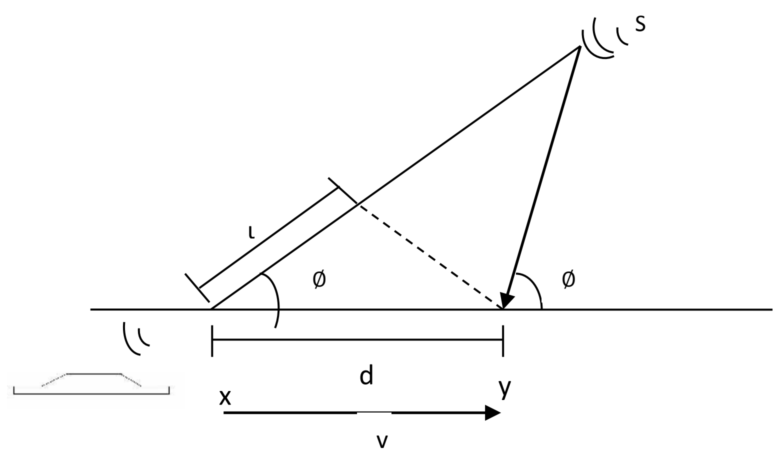


Figure 1.4: Doppler Effect in mobile communication.

1.2.1.4 Doppler Spread

The nature of the channel with respect to varying time is described by Doppler spread (D_S) and coherence time unlike delay spread and coherence bandwidth where the dispersive nature of the channel was considered. In wireless communication, as stated earlier, the transmitted data may overlap and a large number of radio signals arrive at the receiver at different times. Thus the number of changes in frequency due to Doppler shift depends on the number of waves that arrive. This is what causes Doppler spread. The highest Doppler spread occurs when the transmitter and receiver have constant speed and are in relative motion. The multipath signal components that contribute to a single fading possess different Doppler shifts. Doppler spread can therefore be expressed to be the difference in signal component Doppler shifts.

1.2.1.5 Coherence Time

Coherence time refers to a statistical measurement of the time interval when impulse response of the channel is constant. It is the particular period of time when the amplitude correlation of two received signals is strong. Mathematically, the Doppler spread and coherence time are conversely equivalent to one another. That is,

$$T_c = \frac{1}{\max f_d}, \tag{1.4}$$

where, T_C represent the coherence time, $\max f_d$ is the maximum Doppler shift given as $\frac{v}{\lambda} cos\emptyset$.

1.2.1.6 Path Loss

The received signal in free space has a direct line of sight (LOS) in the absence of obstructions such as buildings, hills and any other obstructive object that exists between the transmitter and receiver. There is a gradual loss in the signal power because of the transmitter and receiver space interval and it is referred to as attenuation. The average loss of signal power in free space is called path loss because it is attributed to transmitter and receiver physical distance. As the distance increases, there is a corresponding increase in attenuation. In free space, the received power is inversely proportional to the square of the carrier frequency and square of the distance. In [7] the free state propagation is given as

$$P_R \ \alpha \ \frac{G * P_T}{f^2 * \ d^2} \quad , \tag{1.5}$$

where, P_R is the received power, P_T represent the transmitted power, d represent the interval from transmitter to receiver while f is the carrier frequency. The product of power gains of the transmitter and receiver antennas is G.

The path loss can therefore be expressed as the ratio of transmitted power to received power as

$$P_l = \frac{P_T}{P_R}. \quad \text{(assuming no other losses)} \tag{1.6}$$

1.2.1.7 Shadowing

As stated above, path loss occurs when the receiver is in a direct line of sight (LOS) with the transmitter. Contrary to path loss, shadowing occurs when there is an obstacle in the medium between the sender and the receiver. In this case the signals received at the receiver combines the direct line sight (LOS) signal and the multi-path components. Therefore, the character and nature of the medium is the factor that determines the attenuation experienced by the signal power. Thus, shadowing is the fluctuation experienced along the path of transmission when there are obstructions such as buildings, hills, and any such which act as obstacles between the transmitter and the receiver.

1.2.2 Classification of Fading Channels

1.2.2.1 Non- Frequency Selective (Flat) Fading

It is a type of multipath time delay fading and it occurs when the transmitted signal has a bandwidth (B_S) that is smaller than the coherent bandwidth (C_B). That is $B_S < C_B$. In this situation, at the receiver, the spectral shape of the transmitted signal remains constant or preserved, while the gain changes. Flat fading responds to the channel parameter delay spread, and thus, the channel reacts to the fluctuations due to the multipath effect. It then causes the received signal power to change with respect to time. The changes in gain can lead to deep fades. Different diversity techniques such as space diversity, time and frequency diversity and effective coding are various ways to leverage and control fading distortion. Also, the multipath time delay τ_d of the channel is small compared to the symbol duration T_S. It means that all the multipath received signals have relatively small delay compared to the symbol duration, $\tau_d < T_S$. Hence, the entire frequency spectrum experiences equal fading due to a constant channel transfer function and appropriately named flat fading.

1.2.2.2 Frequency Selective Fading

Frequency selective fading occurs when the channel coherence bandwidth is significantly smaller in comparison to the transmitted signal bandwidth, $C_B < B_S$. The interference and degree of fading is applied more to some of the frequency components of the transmitted signal across the channel, hence, the name frequency selective fading. The transmitted signal has a symbol period that is smaller than the delay spread. In that situation, the received signal is comprised of a multiple version of the transmitted signal which are faded and are delayed for a longer time (high delay spread), $T_S < \tau_d$. A distorted signal is received as a result of the ISI within the radio channel [4, 8]. The challenge of frequency selective fading is greater when compared to flat fading and it can be effectively addressed using multicarrier systems such as the OFDM techniques.

1.2.2.3 Fast Fading and Slow Fading

These two types of fading depend on the Doppler spread. Here, the correlation of the pace of change between the channel impulse response of the mobile radio and the transmitter is important while classifying the fade to be either fast or slow. A situation when the impulse response of the channel changes at a very slow rate compared to the pace of change of the transmitted signal is called slow fading. The channel is considered invariant over one symbol period. In the frequency domain the bandwidth of the transmitted signal (B_S) is greater than Doppler spread (D_S), that is $B_S > D_S$ and the mobile radio channel has a transmitted signal with a smaller symbol period (T_S) than the coherence time (T_C) of the mobile radio channel, that is $T_S < T_C$. The other instance is when the channel impulse response of the channel changes at a swift rate within the symbol period. The transmitted signal possesses a greater symbol period than the coherence time of the mobile radio channel, that is $T_S > T_C$, [8, 9], and the transmitted signal has a bandwidth that is less than the Doppler spread, that is $B_S < D_S$. In fast fading, there is signal distortion which increases as the Doppler spread increases relative to the transmitted signal [9].

Figure 1.5 below Illustrates the 4 different type of fading experienced by a signal either as a function symbol period or baseband signal bandwidth. The following symbols were used:

T_S: Reciprocal bandwidth which is the symbol period.

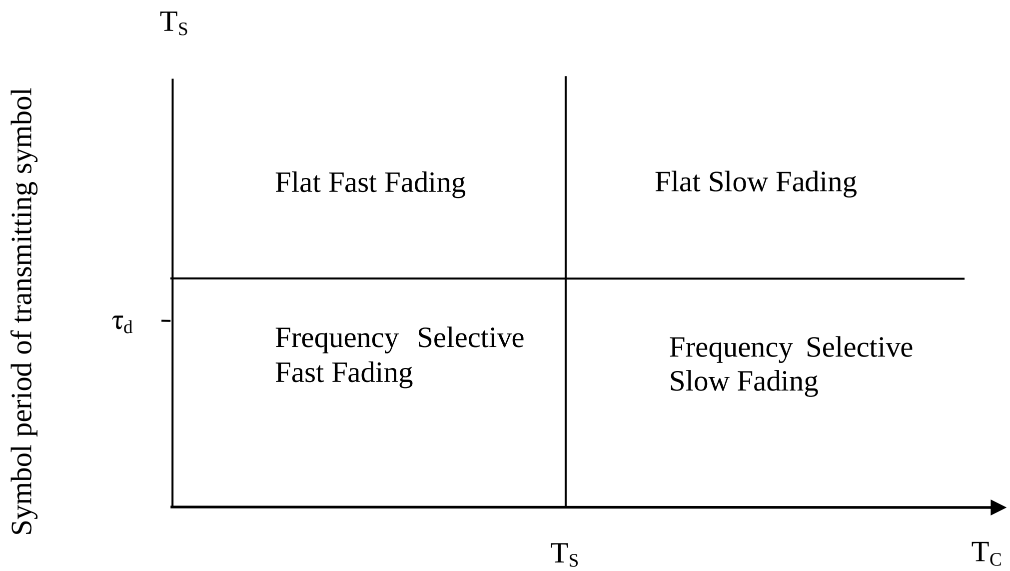
$T_{C:}$ Coherence time.

B_S: Transmitted signal bandwidth.

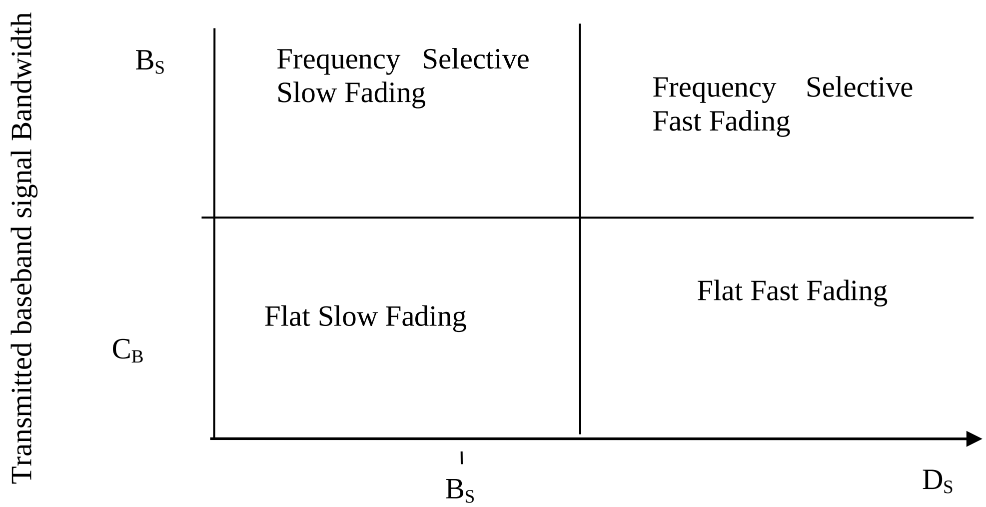
τ_d: The rms delay spread.

C_B: Coherence bandwidth.

D_S: Doppler spread.



(a) Representation in the time domain.



(b) Representation in the frequency domain.

Figure 1.5: Fading experienced as a function of Symbol period and baseband signal bandwidth.

Figure 1.5, above illustrate the relationship between the different multipath parameters such as T_S: Reciprocal bandwidth which is the symbol period, $T_{C:}$ Coherence time, B_S: Transmitted signal bandwidth, τ_d: Delay spread, $C_{B:}$ Coherence bandwidth, D_S: Doppler spread. It shows the impact of symbol period and signal bandwidth on signal.

1.3 Research Motivation and Objectives

Wireless communication has been facing various inherent challenges based on the near unpredictable nature of the wireless environment. These challenges include multipath fading, co-channel interference and Doppler effect among others. Taking into account the plethora of applications and users, there is high demand for data transmission at high speeds. The ability to guarantee broadband mobile data access, at a very high speed, poses a big challenge to future wireless communication. There is also a constant need to ensure that the available spectrum is distributed fairly among the numerous users. The demand has inadvertently led to the evolution of various multiuser communication systems.

In the first to third generation wireless communication systems, FDMA, TDMA, and CDMA were the most commonly used techniques for multi-user systems [10, 11]. However, the combination of OFDM with multiple access techniques formed a hybrid scheme called multicarrier-multiuser systems (MC-MU). MC-MU has brought a lot of gains to multi-user wireless communication systems in terms of spectral efficiency and high data transmission rates. Orthogonal Frequency Division Multiplexing-Time Division Multiple Access (OFDM-TDMA), Orthogonal Frequency Division Multiple Access (OFDMA), Orthogonal Frequency Division Multiplexing-Code Division Multiple Access (OFDM-CDMA), and the most recently introduced Orthogonal Frequency Division Multiplexing-Interleave Division Multiple Access (OFDM-IDMA) [12] are the multiuser techniques which are OFDM based.

Users in OFDM-TDMA wireless communication systems are distinguished by means of time slots. All the sub-carriers in the system are allocated to a specific user for usage at a particular OFDM symbol duration. At the transmitter, all symbols from all the users are combined to form an OFDM-TDMA frame. OFDM-TDMA was adopted by the IEEE 802.16 standard as an option for transmission at the 2.11 GHz band [13]. Though, there are advantages of power saving, simplicity in resource allocation and ease of implementation associated with OFDM-TDMA system, the system exhibits relatively high latency, frequency-re-use factor greater than index factor of 3, and very poor flexibility [14]. Despite all that, there is also the problem of inherent performance degradation for delay constrained systems [14].

The multiple access scheme called OFDMA is based on OFDM transmission technique. The idea of OFDMA is initially proposed in [15] for the return channel in Community Antenna Television (CATV). In OFDMA all the available sub-carriers are divided into non-overlapping subsets and are arranged into different sub-channels which in turn are assigned to distinct users. Interference between users is avoided due to orthogonality among the sub-carriers, thus achieving higher system flexibility and efficiency in the allocation of system resources. Orthogonality is achieved if the condition that the frame length is smaller compared to the cyclic prefix length. OFDMA support unlimited spectrum sharing among users. Some properties of OFDMA are as follow [16]: A maximum of one user is allocated to each of the subcarriers and the selected length of cyclic prefix must be longer than the delay spread. It also utilises DFT and IDFT implementation to reduce computational cost and complexity. It achieved parallel transmission by using the IFFT to modulate each of the coded bits on a subcarrier. Likewise, the signal experience frequency selective fading with a better performance of BER in fading environments. In this scheme, sub-carrier spacing is quite small and it is therefore sensitive to frequency offset and phase offset which is highly pronounced at high mobile speed due to Doppler effect, the amplitude is large and when the signal goes through the amplifier non-linearity the BER increases. To mitigate this problem MC-CDMA is deployed. It is made up of OFDM-CDMA.

In MC-CDMA each user data symbol is spread over the different subcarriers and is multiplied by the spreading codes for onward transmission. As a result, multiple copies of the same data symbol are transmitted on different subcarriers. This implies signal spreading takes place in the frequency domain and the system attains frequency diversity. The individual merit of OFDM and CDMA contribute to the inherited advantages of MC-CDMA. However, the demand for power control in MC-CDMA is a disadvantage [14]. Although the rate loss is avoided in MC-CDMA systems, Multiple Access Interference (MAI) is the major challenge associated with this scheme. MAI can be overcome at the receiver by using Multi-User Detection (MUD). The use of MUD is a costly option because of its high computational complexity [14].

In order to circumvent the problem of high complexity in MUD associated with MC-CDMA systems, Ping *et al* [17] around 2002, introduced a multiple access scheme which is interleave-based and obtained a system with enhanced performance, receiver with less complexity and a system with high spectral efficiency. In 2006 Mahafeno proposed the hybrid MC-MU system called OFDM-IDMA that is MC-IDMA [12]. In this scheme, ISI and MAI over a multipath channel are dealt with by MC-IDMA. It made use of iterative chip-by-chip MUD system which is applicable to as many users as possible. It means that MC-IDMA is not user limited. The scheme utilised all the inherent advantages associated with the widely used multiple access scheme which are OFDMA, CDMA, IDMA. It discards their disadvantages in the process. The receiver has a simple structure with less complex MUD incorporated, and the number of users do not affect the cost. MC-IDMA achieves optimal performance. It avoids matrix operations. Correlation of signal from different users is avoided by employing chip-level interleaving.

MC-IDMA wireless communication system is a new multiusers technique. Extensive studies on this scheme bear witness to a system that is not susceptible to error -an almost perfect system. Perfect systems are still elusive though. The system must be affected by its environmental. The channel impulse response is not known at the receiver. It then makes detection of the message information difficult. There is thus a need for the channel to be estimated. In MC-IDMA, channel estimation deserves more attention as only a few studies have been conducted in this area. A soft based decision directed channel estimation was used by the authors in [18, 19] to implement the least mean square (LMS) algorithm for their proposed channel impulse response (CIR) estimation in the time-domain. However, the limiting factors associated with the use of LMS algorithm are excess mean square error (MSE), data and slow dependence convergence problems.

Furthermore, channel estimation scheme based on pilot-assisted symbol was proposed in [20]. The authors employ concurrent transmission of both pilot and information symbols. Unfortunately, the use of pilot symbols is known for its wastage in the scarce communication bandwidth. Besides, the channel estimation employed by such scheme makes use of interpolation techniques at data points and the pilot symbols exclusively. The estimation of the channel at data point cannot be 100 percent perfect. An estimation scheme, employing the combination of linear algorithm-based CIR estimation and adaptive algorithms-based CIR prediction is proposed for MC-IDMA systems in [21]. In the proposed scheme, the channel estimator exchanges soft information with the MUD only. However, the complexity of the proposed scheme is high.

A robust channel estimation technique, better than what has been reported in [18-21], need to be investigated. From the literature few works exist that estimate the MC-IDMA system employing semi-blind channel estimator. The newly proposed MC-IDMA systems is estimated using the semi-blind channel algorithm bearing in mind both optimum performance and energy efficiency of the multi-user systems is the motivation for this study, the aim and objectives of this thesis include :

- Development of algorithm for semi-blind channel estimation for OFDM-IDMA based multiuser systems.
- Development of simulation and generation of comprehensive results in terms of bit error rate (BER) and mean square error (MSE).
- Performance analysis for the developed semi-blind channel estimation schemes.

1.4 Original Contributions

The effectiveness of channel estimation and analysis of the performance of the recently proposed OFDM-IDMA systems based on combination of OFDM and Interleave Division Multiple Access (IDMA) is the main contributions in this thesis. MC-IDMA uses interleavers as the only means of user separation and devotes most of the bandwidth expansion to the Forward Error Correcting codes (FEC). It supports a large number of users with enhanced BER performance.

- In this thesis, semi-blind channel estimation for a fast fading and slow fading channel is investigated. The former is represented by mobile speeds of 120km/hr and the latter 5km/hr.
- MMVE algorithm developed in [22] and used for channel estimation in OFDM system is herein adopted for the case of Multi-user MC-IDMA system.
- Thereafter, MMSE estimator is modified using the assumption of a finite length response by reducing the estimator matrix to develop semi-blind channel estimation algorithm tagged “Modified Minimum Mean Value Estimator (MMMVE)”.
- Finally, semi-blind channel estimation using an algorithm that combines MMSE and eigen decomposition is developed and this is named as “Simplified Mean Value Estimator (SMVE)”.

An efficient channel estimation algorithm is proposed which is based on the concept of linear MMSE-based algorithm and orthogonal pilot sequences to form a semi-blind estimation technique. It is utilised to obtain the channel state information (CSI) instead of the conventional pilot symbols which dedicates an entire block to each user for training. The approach is therefore appropriate for

MC-IDMA systems operating in high mobility channels. In order to achieve a better and improved system performance, using the structural properties of the transmitted symbol, an improved algorithm is also developed from the MMSE algorithm, the Eigen Value Decomposition (EVD) algorithm. The performance of these three algorithms is tested for different number of users. The performance of the improved algorithm is compared with the performance of MMSE algorithm simulated on an OFDM-IDMA scheme. To the best of our understanding, the proposed algorithms have never been utilized for channel estimation in the MC-IDMA system. To prove its effectiveness in practical scenario the algorithm is applied to Rayleigh fading multipath channels and the mobile speed is varied. Computer simulations results are presented. The algorithms are further comprehensively compared to ascertain which of the algorithms offers a better and more efficient system performance. The BER is evaluated to predict the performance of MC-IDMA system. Analytical and simulation results show that the improved algorithm perform better.

1.5 Expected Publications

From the research works amassed in the course of study, the following articles are under preparation and will be submitted for possible consideration for publication in conference proceedings and journal.

1. Juliana A Adisa, Olutayo O. Oyerinde and Stanley H. Mneney, "Semi-blind channel estimation employing Minimized Mean Value-MMSE Estimator (MMVE) algorithm for MC-IDMA systems," to be submitted to *Southern Africa Telecommunication Networks and Applications Conference (SATNAC 2015)*.

2. Juliana A Adisa, Olutayo O. Oyerinde and Stanley H. Mneney, "Modified Minimized Mean Value-MMSE algorithm-based Semi-blind channel estimator for MC-IDMA systems," to be submitted to *IEEE AFRICON* 2015.

3. Juliana A Adisa, Olutayo O. Oyerinde and Stanley H. Mneney, "Combined Simplified Mean Value-based MMSE and Least Square algorithms for semi-blind channel Estimation in MC-IDMA wireless communication systems," to be submitted to the Transactions of the SAIEE journal.

1.6 Thesis Organization

The organization of this thesis follow the structure below.

Chapter 1 is an investigation into wireless system and the propagation aspects of the radio channel. Understanding the characteristics of radio channel is critical to investigation of channel estimators. The general introduction serves mainly as a refresher for a reader who is familiar with the area of wireless communication systems.

Chapter 2 describes OFDM basic principles, its working model, properties, parameters, and applications. Furthermore, different ways in which multiple users can share the available radio channel are reviewed. IDMA operation is presented and an iterative detection algorithm is described with rigorous mathematical details and comprehensive comparisons between Multiple Access Scheme (MAS).

Chapter 3 present an introduction to MC-IDMA system and channel estimation in MC-IDMA. It provides a simplified and efficient channel estimation algorithm for the uplink transmissions.

Chapter 4 explore the proposed improved channel estimation algorithms and interrogates there corresponding computer simulated result.

Chapter 5 describes OFDM-IDMA computer simulations and results using semi-blind time domain estimation technique. It also compare the complexity and BER performance between the different algorithms.

Conclusions and the possible future work are given in chapter 6.

CHAPTER 2

MULTILPE ACCES SCHEMES, OFDM AND IDMA OVERVIEW

2.1 Introduction

Recently, a number of ways have been proposed to utilize and manage the available radio spectrum among the users of the wireless communication systems efficiently. Multiple simultaneous users in a system are supported with users utilizing any of the multiple access techniques which are related to the frequency band, the time band and the code band. The use of Orthogonal Frequency Division Multiplexing as a prominent technology in high rate data transmission is also a global convergence. The fourth-generation (or the professed 3.9 G) mobile systems, WiMax, WiBro, WiFi etc are all wireless local network systems that are OFDM based. Digital multi-carrier modulation is the scheme used by OFDM. It is used in data transmission at high data rates with frequency-selective channels.

This chapter gives a broad view of the OFDM process. It introduces IDMA as a best multi-user system to be combined with the OFDM system (thus forming the MC-IDMA) in order to obtain energy efficient wireless communication systems. This chapter also introduces the study of different channel estimation techniques that exist. The estimation of the channel using the pilot symbols (Pilot-assisted channel estimation method), estimation based on inherent information in received signal, and intrinsic information in the unknown data symbols, respectively. The last channel estimation techniques called decision directed channel estimation is also discussed. The importance of semi-blind channel estimation schemes and the reason why it is better than other estimation schemes is highlighted here.

The layout of the chapter is summarised as follows. Section 2.2 gives an Introduction to Multiple Access Schemes (MAS) and it details the different techniques. A brief history and principle of OFDM is described in Section 2.3. IDMA principles with its chip-by-chip mathematical analysis are given in Section 2.4, while Section 2.5 compares the various MAS techniques. Section 2.6 presents the various channel estimation techniques. The chapter is summarized and concluded in section 2.7.

2.2 Multiple Access Scheme (MAS)

In wireless services, a finite amount of spectrum is distributed and the ability to send information simultaneously by different users is desired. The limited spectrum can be shared effectively among as many users as possible. In general, a multiple-access scheme (MAS) is a scheme in communication where the limited available channel bandwidth is shared and allocated to various users resulting in efficient and high capacity usage of the limited spectrum. In wireless communications design, MAS ensures that the channels allocated to different users are well managed in spite of the limited spectrum. The sharing of the channel bandwidth among the various users is based on time, frequency, or codes, as the case may be. In the section below, the major multiple access scheme are discussed.

2.2.1 Frequency Division Multiple Access (FDMA)

The bandwidth is assigned to various users in the frequency domain and each user is allocated a finite portion of bandwidth for constant and permanent usage. The channel bandwidth is divided into a number of smaller non-overlapping channels separated by guard bands to prevent interference, (Fig. 2.1) Users are said to be frequency-orthogonal in FDMA. It stand to follow then that a particular user (ideally) cannot have impact on other user [23]. FDMA by nature is a narrowband MAS system which is usually executed in narrowband systems. However, since a finite portion of the bandwidth is permanently allocated to a specific user, the bandwidth becomes a wasted resource when the user is idle. As stated earlier, United State of America utilized Advance Mobile Phone Service (AMPS) as the first analog cellular system based on FDMA. According to [9], the mathematical expression for the number of simultaneous users that can share the limited channel bandwidth in FDMA scheme is given as

$$N = \frac{B_D - 2B_{Gd}}{C_B} \quad , \tag{2.1}$$

where, B_D is the total available bandwidth, B_{GD} the guard band to mitigate multipath delays and interference and C_B represent the channel bandwidth.

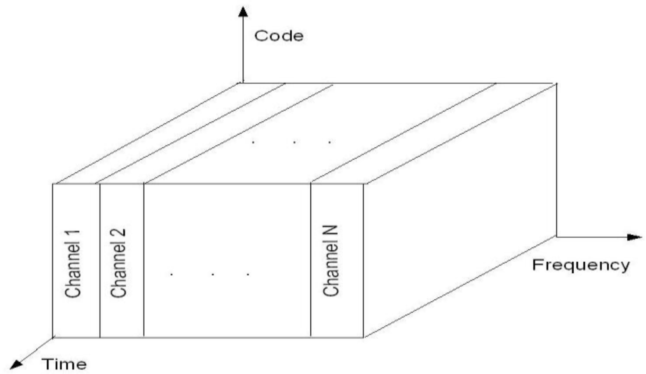


Figure 2.1: Specrum sharing with respect to frequency.

2.2.2 Time Division Multiple Access (TDMA)

In this scheme, the whole bandwidth of the channel is available to all the users but only for a limited period of time. This means that TDMA is constrained in the time domain [4]. Since continuous transmission is not required in digital systems, TDMA transmission occurs in bursts. The same carrier frequency is useed by all the users but non overlapping time slots are available where the user can either transmit or receive radio signals. In TDMA, users are time-orthogonal, since transmission by different users is disjoint in time, fig. 2.2, [24]. In most cases, the accessible bandwidth is split into a small number of channels in contrast to FDMA and the users are assigned different time slots at which they can have the whole channel bandwidth for their transmission processes. In this modulation technique, the available bandwidth affect the number of time slots per frame. In the upper link of TDMA, the users transmit through different channels and have different delays. Synchronization of the transmitter in the up-link, to maintain orthogonality of the received signal, is important. Therefore, introduction of guard bands between TDMA channels to compensate for multipath and synchronization error is required. Two guard bands are used, each of the guard bands are positioned at the two different ends of the frequency assigned. TDMA is used in the GSM system. Mathematical expression for the total number of simultaneous users that can share the limited channel bandwidth in TDMA scheme is given by [9]

$$N = \frac{n(B_D - 2B_{Gd})}{C_B}, \tag{2.2}$$

where, the maximum number of users allowed per radio channel is n, while B_D, B_{GD}, and C_B are the total available bandwidth, the guard band and the channel bandwidth respectively.

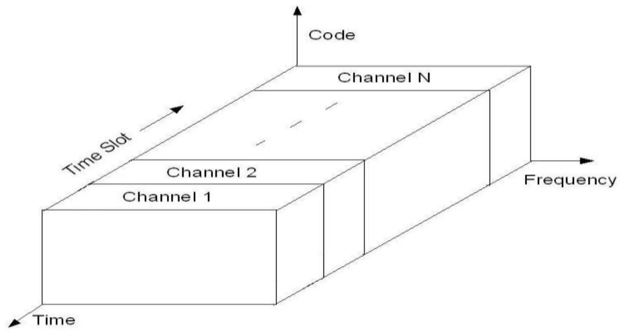


Figure 2.2: Specrum sharing with respect to time.

2.2.3 Code Division Multiple Access (CDMA)

A type of Direct Sequence Spread Spectrum (DS-SS) is the CDMA system, in which the uncorrelated spreading signal with a broad bandwidth is used to disseminate the narrowband message signal. As a result, the transmission of the input signal is over a large bandwidth. A spreading code which is also known as a Pseudo Random Noise Code (PN code) is used in CDMA to multiply the message signal. PN-code appears random noise-like but, there is a precise mathematical rule that can be used to generate them [23]. The pseudorandom code of a particular user is orthogonal to the code of other users. Therefore, all the users in CDMA systems occupy the same bandwidth and transmit concurrently without interfering with each other. It is required at the receiver end to know the code-word used at the transmitter in order to detect the data of the right user. The knowledge of the processing gain of the system enables one to determine the number of users permitted to transmit in a CDMA system. The processing gain is given by

$$P_G = \frac{S_{Bw}}{S_{Tw}}, \tag{2.3}$$

where, P_G represent the process gain at the receiver, which is the relationship between the spread spectrum bandwidth (S_{BW}) and transmitted signal bandwidth (S_{TW}). The difficulty to jam or detect the transmitted signal and the amount of multi-path effect reduction are determined with the CDMA processing gain. The reduction of cross-cell interference, robustness against fading, high flexibility, transmission in asynchronous form, ease of cell planning, reliable data encryption and dynamic channel sharing are among good quality properties of CDMA [25].

At the receiver, multiple access interference (MAI) problems surface because the existing orthogonality among the spreading sequence of simultaneous users is lost. MAI is the result of the random time offsets between signals and it occurs between direct sequences users. CDMA has attractive features but MAI limits its capacity and performance and degrades the system. It is important that the users' data signals are recovered at the receiver. This is the major problem with CDMA because of the cost and complexity involved in Multi–User Detection (MUD) techniques [26, 27]. Figure 2.3 shows the sharing of the available bandwidth with respect to PN code.

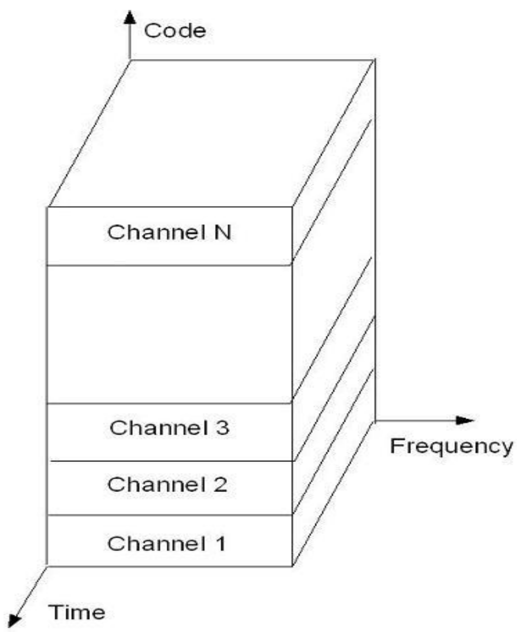


Figure 2.3: Specrum sharing with respect to code, CDMA.

2.3 Fundamental Operations of OFDM

Orthogonal frequency division multiplexing is a modulation technique which is suited for high-data-rate transmission in a delay dispersive environment. In OFDM, the whole channel is split into numerous narrow band sub-channels, also referred to as subcarriers. This splits the high data-rate stream into many low data-rate parallel streams that are transmitted over parallel narrow band multiplexed sub-channels that are mutually orthogonal. There is thus, no interference between subcarriers. It then becomes easier to separate signals carried by each sub carrier because of the absence of interference. OFDM is not just a modulation technique; it can also be referred to as a multiplexing technique. The OFDM technique allows for spectrum overlapping, thus ensuring the efficient use of the available bandwidth.

2.3.1 History of OFDM.

Literature review on the history of OFDM is found in [28]. Origin of OFDM technique can be traced back to the work of Chang [29] in the mid 60's when he published a paper on synthesis of bandwidth. The idea presented in his paper was on how to simultaneously transmit data signals via a linear band limited channel in the absence of multipath interference which causes Inter-Symbol Interference (ISI) and Inter-Carrier-Interference (ICI). His work was patented in 1966. Following this was the work of Saltberg [30] where the performance of signal transmission in parallel form was discussed and its effectiveness analysed. He opined that instead of perfecting the individual channels, effort should be concentrated on how to reduce cross talk between adjacent channels. Baseband modulation and demodulation using Discrete Fourier transform (DFT) is the subsequent work on OFDM system carried out by Weinstain and Elbert [31]. The consecutive transmitted symbols were separated in time domain with guard intervals using a raised-cosine window to combat ISI and ICI. The orthogonality between subcarriers in the system used by Weinstein was not perfect over a time dispersive channel. Another significant contribution was introduction of cyclic extension by Paled and Ruiz [32]. They used the cyclic extension of the OFDM transmitted symbol to creatively fill the empty guard interval introduced by Weinstein. OFDM systems are widely used in most popular applications and hence there is need for it to perform better. Consequently, there is need for more work to be done in the research of OFDM systems. Robustness against sampling frequency errors, timing error, phase noise and carrier frequency offset could be achieved through synchronization [33]. A precise channel state Information that improved OFDM system performance is obtained by channel estimation [34-36]. Techniques such as clipping and peak windowing to reduce the relatively high PAPR are presented in [37].

Over time, various improvement were made on the OFDM techniques and in 1985, Cimini presented OFDM as a solution to wireless communication problems and it was adopted with its suitable coding and used as a wireless technology standard in the new Terrestrial Digital Video Broadcasting (DVB-T), Digital Audio Broadcasting (DAB), high data-rate wireless LAN standard [38, 39], Multimedia Mobile Access Communication (MMAC), as well as the IEEE 802.16a Metropolitan Area Network (MAN) standard. OFDM has a good potential to be deployed in future wireless systems, particularly, the 4^{th} generation mobile wireless systems [40].

2.3.2 Serial and Parallel Data Transmission

In a common serial data transmission system, the spectrum of each data symbol being transmitted occupies the whole available bandwidth and the symbol transmission is continuous. In this serial system, when fading occurs, several adjacent symbols may be totally destroyed because of the bursty nature of the Rayleigh channel. The system is more susceptible to delay spread impairments if the symbol interval is decreased. A multiplexed data system is also referred to as a parallel system which proffers an alternative solution to the problems confronted with serial transmission. In a parallel system, continuous streams of data are transmitted concurrently, so that large numbers of data elements are transmitted at any point in time. This system is designed such that the individual data element spectrum can only use a small fraction of the total bandwidth that is available [41].

2.3.3 FDM Scheme Versus OFDM Scheme

Frequency-division multiplexing (FDM) is a scheme in which the total available bandwidths are divided into smaller non overlapping frequency bands. In a conventional FDM system, filters and demodulators are used at the receiver to recover information in each carrier. There is introduction of guard bands among all the various carriers. The use of guard bands results in inadequate use of the spectrum that is scarce and expensive.

In a regular FDM system, a single fade or interference can lead to failure of the whole link. However, in a multicarrier system, only a small fraction or 0.01% of the subcarrier will be affected and hence can be corrected with error correction coding [42]. Although, a conventional parallel data transmission can occur in FDM, there is inefficient use of bandwidth. Fig. 2.4 shows a conventional FDM and OFDM system. Fig 2.4 (A) represents the FDM system; the channels are spaced apart to prevent Inter-carrier-interference (ICI), and the signal message can be recovered with conventional filters and demodulators. It can be seen from Fig. 2.4 (B) that 50% out of the total bandwidth is conserved by the multicarrier modulation OFDM overlapping technique.

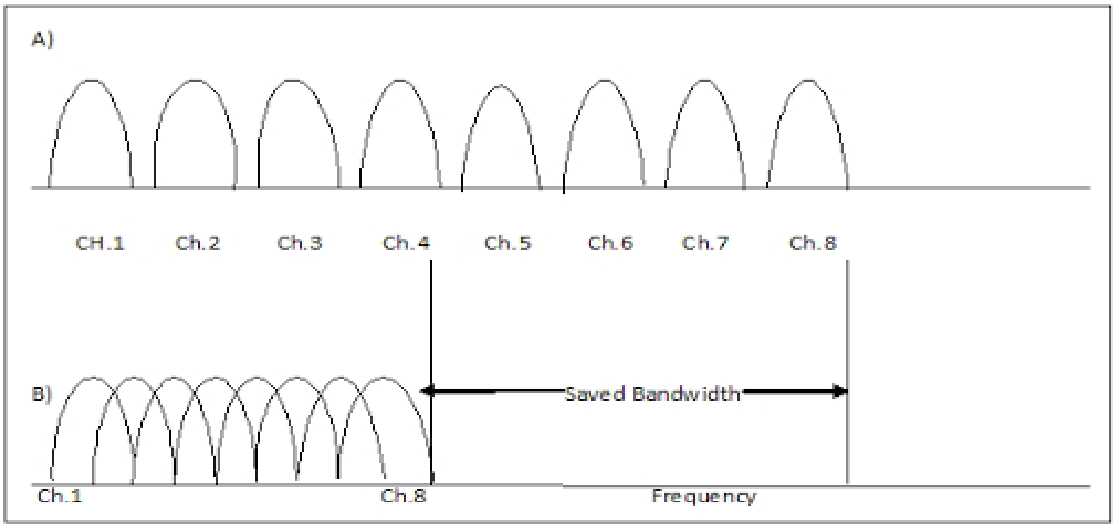


Figure 2.4: (A) and (B) conventional FDM and OFDM.

The word OFDM begins with letter "O", which stand for orthogonal. Orthogonality differentiates the commonly used FDM from OFDM. OFDM inherited all the advantages of FDM. Orthogonality is maintained between subcarriers by carefully choosing the frequency range between the subcarriers. It can be proven mathematically that when the dot product of two signals is zero, the two signals are said to be orthogonal. In fact, the subcarriers are spaced apart by distance of 1/T, where the duration of an OFDM system is represented by T. Figure 2.5 shows the frequency spectrum of subcarrier in an OFDM transmission. From the diagram, it can be seen that the spectrum of subcarriers significantly overlaps across each other, unlike in the conventional FDM operation. For a large number of subcarriers associated with parallel transmission, a large number of oscillators [43] or arrays of sinusoidal generators and the coherent demodulator are needed. This makes the parallel systems costly. Implementing OFDM is therefore complex and costly. However, in 1971, [31] an effective way to reduce the OFDM implementation complexity was proposed by Weinstein and Ebert, who employed Discrete Fourier Transform (DFT) for effective baseband modulation and demodulation. DFT uses basis functions which are sinusoidal and cosinusoidal and are harmonically related functions. Thus, sinusoidal generators or oscillators are eliminated, thus, reducing complexity significantly. The work in [31] focuses on how to introduce sufficient processing, thus eliminating the bank of subcarrier oscillators. Inter-symbol-interference (ISI) and inter-carrier-interference (ICI) are mitigated using a raised-cosine windowing.

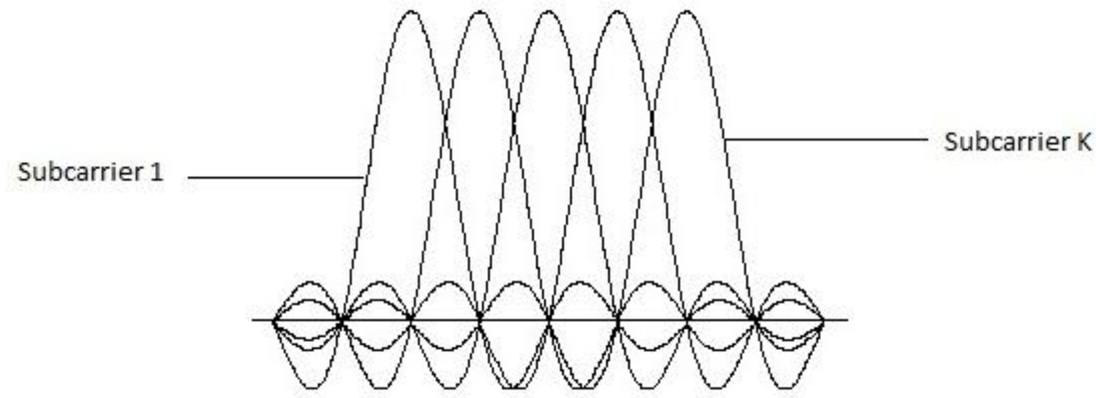


Figure 2.5: Frequency spectrum of an OFDM transmission.

2.3.4 OFDM system block diagram

Briefly, the original data to be transmitted is split into streams of N parallel symbols, each of which has a lower data rate. The modulated data is assigned to subcarriers based on subcarrier assignment information obtained from subcarrier level sensing [44]. Each subcarrier is modulated in phase and amplitude by the N parallel data streams which has a lower data rate. Figure 2.6 illustrates the OFDM transmission process that occurs at both the transmitter and receiver.

In more detail, the principle of operation of OFDM systems is as follows:

On the transmitter side, the binary serial data input is first channel coded. Application of channel coding on the input data ensures that the BER (bit error rate) is kept to the lowest value thereby improving the system performance. It is also used to lower the peak-to–average power ratio (PARP) as proposed in [45]. The encoded data is interleaved; this is assigning adjacent data bits to non-adjacent bits to reduce the burst symbol error. The bits are then mapped into symbols depending on the type of modulation used; QPSK, 16/64 QAM or BPSK. The data symbols are used to modulate each sub-carrier in phase and amplitude, the number of symbols used depends on the type of modulation process.

The serial modulated data streams are changed to parallel and fed into the Inverse Fast Fourier Transformation (IFFT) which transforms the data stream from frequency domain to time domain. The original serial data is the effective Fourier Transform of multitone data signal and an inverse Fourier generator bank of coherent demodulators [31]. It can be seen that OFDM scheme is a complete digital modem built to perform Fast Fourier Transformation (FFT) and IFFT (the implementation of Discrete Fourier Transform (DFT) and Inverse Discrete Fourier Transform

(IDFT)). It eliminates subcarriers (SC) oscillators and coherent demodulators used in FDM, thereby reducing the cost of the OFDM processor. When the signal is in the time domain, the DFT is used to compute its samples in the frequency domain. T is used to represent the sampling period, and N represent the number of points to be sampled as related to the frequency domain. The DFT basic frequency can be given as $\frac{1}{NT}$. An integer multiple of the basic frequency is the component of each frequency for each subcarrier. According to Nyquist sampling theorem, time domain signal sampling can generate a maximum frequency that is represented by $f_{\mathrm{max}} = \frac{1}{2T}$. The center of the DFT point is where the carrier frequency is located. The operation carried out by IDFT is the opposite of DFT operation. The signal defined in frequency domain is converted to the time domain. The IDFT time signal has a time duration which is equivalent to *NT*. In essence, both are a reversible pair. DFT and IDFT can be used at transmitter and receiver side respectively. The parallel output of IFFT is converted to serial because the input of IFFT is made up of N samples (the symbols for the different subcarriers), therefore the output of IFFT also consist of N values. These N values need to be transmitted one after the other as temporal samples [39].

The output time-domain OFDM symbol is extended by a cyclic prefix (CP). Cyclic prefix (CP) is a special type of guard; it is a duplicate of the last part of the entire OFDM symbol which is pre-fixed to the transmitted symbol. The CP length T_G is such that it is selected to be longer in length than the delay spread experienced by radio channel, τ_d, i.e., $T_G \geq \tau_d$. Thus, the essential component of the signal received is perceived as the OFDM transmitted symbol convolved with the channel impulse response. The neighbouring OFDM symbols partially overlap due to dispersion. The cyclic prefix function is to preserve orthogonality among the subcarriers resulting in zero ICI. This only occurs when the CP length is maintained and is more than the delay spread. Also, within the FFT interval, an integer number of cycles must be produced by the delayed copy of the OFDM symbol. This is established by the cyclical extended portion of the OFDM symbols. An integer number of cycles are produced by individual sub-carrier when an OFDM symbol with no CP attached and is equivalent to the size of the IFFT employed to produce the signal.

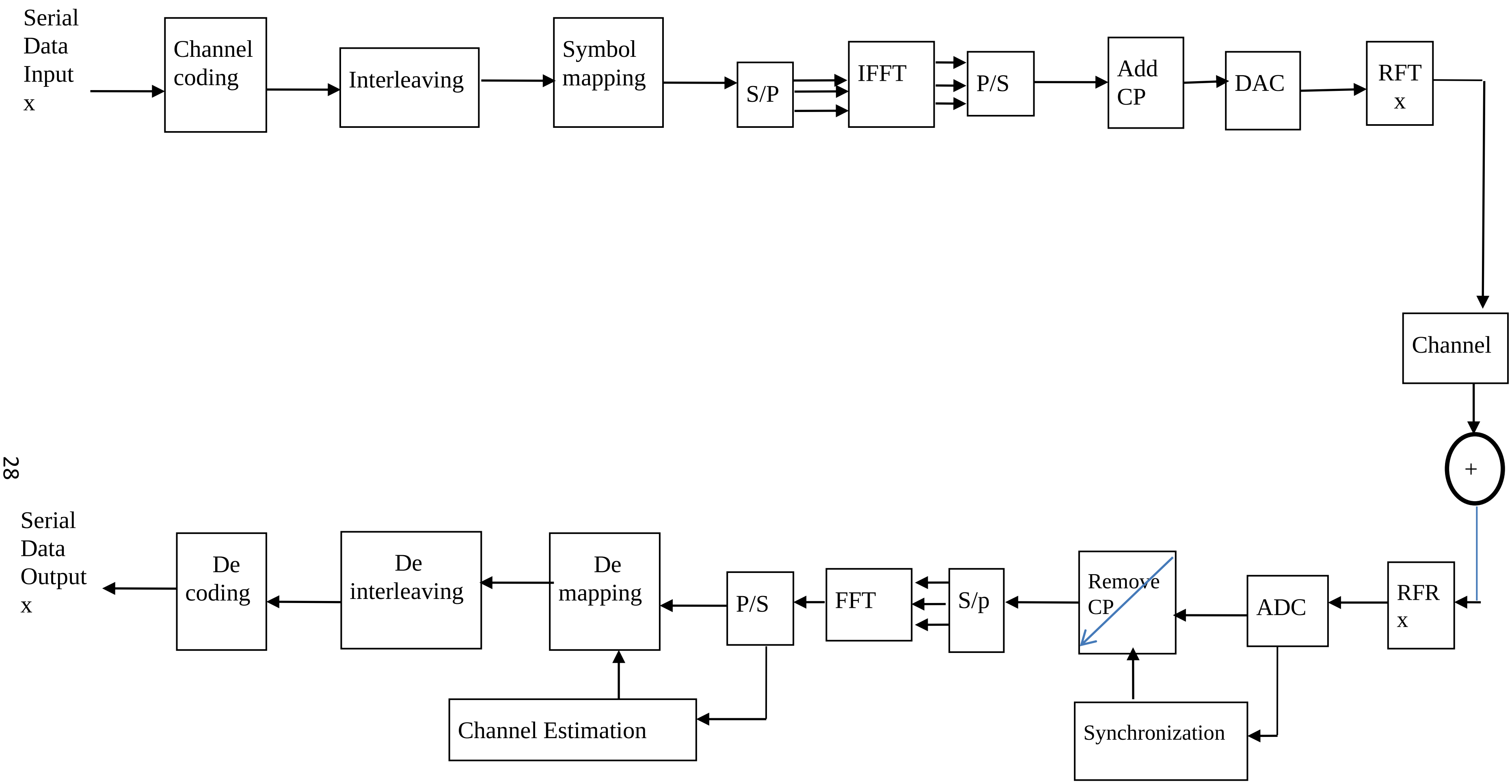


Figure 2.6: Block diagram of a baseband OFDM transceiver.

The end-to-end placement of transmitted symbols copies lead to uninterrupted signal transmission with no disruption at ends. Addition of CP makes the system to have a longer symbol time because the end of the symbol is copied and appended to it in front. Figure 2.7 shows an OFDM system with the insertion of a cyclic prefix. The symbol period in the samples is represented by T_S, which is the aggregate symbol period $T_S = T_G + T_{FFT}$, where CP length in samples is given as T_G and the estimated size of IFFT is used by T_{FFT} to produce an OFDM signals. CP of not more than 10 per cent of the symbol's duration is employed thus leading to signal to noise ratio loss of 0.5 to 1 dB. The analogue OFDM symbol can be obtained using the digital-to-analog converter (DAC) before the signal is amplified and converted to the desired centre frequency before transmission in the frequency selective fading channel. Power consumption, complexity and cost, must be considered before using DAC.

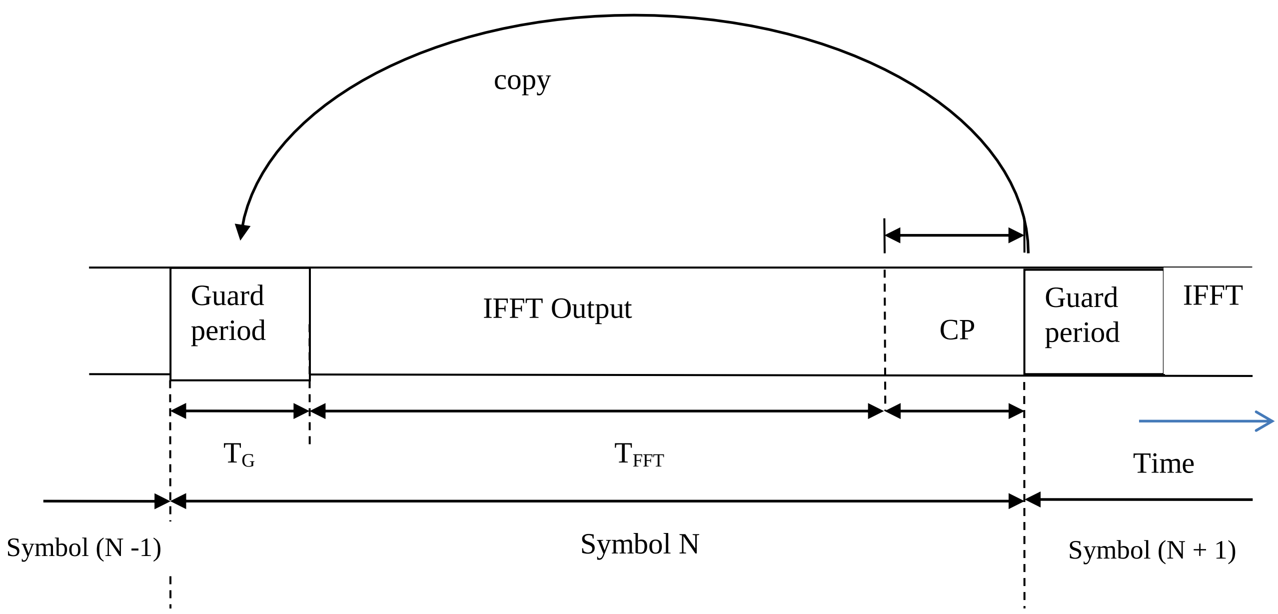


Figure 2.7: Cyclic prefix fitting.

The other part of the OFDM transceiver is the receiver end of the system where the transmitted symbols are passed through the analog to digital converter, removal of CP and at this point synchronization process takes pace such as symbol, carrier, frequency and sampling frequency synchronization. Serial to parallel conversion occurs to allow the FFT operation process which converts the time domain signal to the frequency domain representation. Channel estimation is also done either in the time domain or frequency domain and the information is used to detect or rececover the transmited data.

2.3.5 The significance of OFDM systems

From the OFDM history, there are various transformations in OFDM operation through the years. The main OFDM systems contributions are:

- Transformation from analog multicarrier systems to their digital implementation. Prior to this time, sinusoidal generators and coherent demodulator were required for modulation and demodulation processes. Presently, baseband modulation and demodulation is carried out using DFT (discrete Fourier transform), therefore eliminating the subcarrier oscillators and coherent demodulators. This transformation has led to a reduction in cost of OFDM systems. Also, for highly competent processing, the complete execution of DFT-based frequency-division multiplexing is not by bandpass filtering but by digital baseband processing. The number of arithmetic operations in DFT can be reduced from N^2 to $N\log_2 N$ (N is FFT size) using the fast Fourier transform algorithm (FFT). Yet another breakthrough in Very Large Scale Integration (VLSI) made the commercial accessibility of FFT chips possible in large scale size and relatively high-speed.
- ISI can completely be eliminated in a system. The length of CP in an OFDM system determines the level of ISI elimination. When the CP period is made long compared to the channel impulse response length, then ISI will be completely removed. When performing cyclic convolution over a time dispersive channel orthogonality between subcarriers is maintained. At the receiver part, there is an energy loss that is proportional to the CP length. This loss occurs as a result of CP removal from the received signal and is the cost of the zero ICI that result. This is the second major contribution to OFDM systems.

2.3.6 The challenges of OFDM Systems

The potential for a transmission link with high speed is one of the attractive properties of OFDM as a modulation scheme. Nevertheless, its large Peak-to-Average Power Ratio (PAPR) is a limitation to its operation. A robust OFDM receiver is designed with synchronization consideration in mind. In order to know when transmission of OFDM symbols start and end, time and frequency synchronization is neccesary and the modulator and demodulator's local frequencies are matched, respectively. Should any of the synch steps not match each other with utmost accuracy, the whole concept of orthogonality among SCs is defeated. It follow that ISI and ICI will be introduced. Synchronization (frequency and time offset) and how to properly estimate the channel and ensure that the correct channel state information (CSI) are available at the point of demodulation is a major challenge and the focus of this thesis. The OFDM system challenges are discussed below in more detail.

2.3.6.1 Peak-to-average power ratio (PAPR)

PAPR of an OFDM emerged when the received signal is considered to be higher than the average amplitude. The reality that the OFDM signal is the superposition of a number of subcarrier sinusoidal signals [39] is the origin of the PAPR problem. According to papers [46, 47], the expression of PAPR of a transmitted signal x(t) as the ratio of the maximum instantaneous power to the average envelope power of the signal and is given below as

$$PARP = \frac{max|x(t)|^2}{E[|x(t)|^2]}, \tag{2.4}$$

where, $E[|x(t)|^2]$ is the signal average power.

In an OFDM wireless system the high power amplifier (HPA) must have a large linear range which is efficiently used and when a non-linear HPA is used, the non-linearity destroys the orthogonality between subcarriers, there is out of band emission, and there is an increase in the bit-error rate (BER) due to distortion and interference to other users [44].

RF power amplifier with reduced efficiency, and increased complexity in the D/A (digital to analog) and A/D (analog to digital) conversion processes are some of the disadvantages associated with large PAPR. The drawbacks associated with PAPR necessitates the search for a viable technique to combat PAPR. Todate, there are several PAPR reduction techniques which include the following; coding and phase adjustment [44, 45], clipping and filtering [48, 49], peak windowing and interleaver technique [50, 51]. In [52] the review of some major techniques is presented. In the review paper, it was stated that in spite of the fact that reduction of capacity due to PAPR, increased power, the complexity and increased BER are issues considered for PAPR technology selection. However, the cost for reducing extra PARP complexity must be lower than the cost of power inefficiency. In [51] a data randomization technique is presented, where it is submitted that by interleaving a data frame, the related OFDM signal crest can be suppressed. In all of these techniques, PAPR reduction is basically carried out at the transmitter. The PAPR can be reduced significantly with moderate increase in BER using the clipping method and other method such as peak windowing, and peak cancellation [50].

2.3.6.2 Synchronization

OFDM systems are highly sensitive to synchronization error. Therefore, processes such as symbol, carrier, frequency and sampling frequency synchronization are necessary. Many synchronization schemes have been proposed [33, 53, 54]. Inaccurate synchronization lead to

- Timing error: timing error lead to subcarrier phase error and if the offset is large to ISI.
- Carrier Phase noise: which is considered as the shift in frequency δf and the equivalent received signal phase offset θ.
- Frequency error: this is as a result of differences in transmitter and receiver oscillator frequencies, non-linear channels introduced Doppler shift and phase noise. There are two damaging effects for frequency offset; signal amplitude is reduced and ICI is introduce. This means that there is a shift in sinc functions and these are no longer sampled at the peak. Also (ICI) by other carriers [39] is introduced. Figure 2.8 below shows the effect of frequency offset in a system.

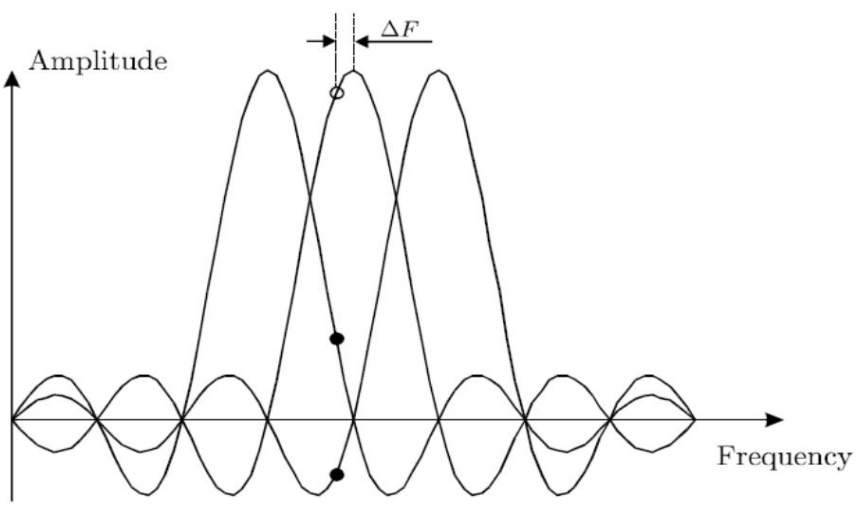


Figure 2.8: Frequency offset ΔF: in a signal amplitude (o) and intercarrier interference(▪)

Some of the synchronization schemes exploit the redundancy in the cyclic prefix [33, 54]. The data-aided synchronization technique is engaged in [53, 55]. A summary of time and frequency synchronization is given in [54]. The received serial output from the synchronised signal is changed to parallel, and FFT operation takes place, the parallel output of FFT is changed back to serial and is de-mapped.

2.3.6.3 Channel Estimation

A reliable and accurate estimation of the channel state information (CSI) is expedient to be known before the demodulation of the transmitted signals. Whether the signal is distorted or not the radio channel is frequency selective and time varying in nature [7]. Also, accurate signal demodulation request for correct and up to date estimate of the channel. There are three techniques which are used in channel estimation and these are discussed in section 2.6.

2.3.6.4 Radio frequency amplifiers

There is need for radio frequency (RF) amplifiers with a high PARP because the amplitude of the OFDM signal at the receiver is noisy and the range is very large.

2.3.6.5 Addition of guard bands

Addition guard bands lower the symbol rate and hence the overall spectral efficiency of the system is compromised.

2.3.7 Advantages of OFDM systems [44]

- The spectral overlapping among sub-channels serve as a means to cancel equalization, combating impulsive noise, and making complete use of the available bandwidth [31].
- It is robust against frequency selective fading channel because the data is split into N parallel symbols. Each in the transmitted symbol has a bandwidth narrower than the bandwidth of sub-channel; hence OFDM experiences only flat fading.
- Narrowband interference has little effect on few subcarriers in OFDM because of OFDM is resistant to it.
- It uses cyclic prefix to eliminate ISI and ICI.
- Very large scale integration (VLSI) technologies have facilitated adaptation of OFDM by overcoming problems of high-speed cached-memory architecture [41]
- Impulsive parasitic noise and co-channel interference are avoided.

- The use of steep band pass filter is removed due to spectral overlap among subcarriers. Orthogonality will ensure the separation of subcarriers at the receiver.
- There is an immunity embedded in OFDM which effectively deals with delay spread, a channel parameter that need attention in wireless communication.
- Multi-taps equalizer/detector is not used in OFDM receiver, thus reducing complexity.
- OFDM systems are cost effective and computationally efficient implementing modulation and demodulation with FFT techniques.
- It is less sensitive to sample timing offset compared to the single carrier system. A single carrier system is more sensitive to sample timing offset than OFDM

2.3.8 Applications of OFDM

A Simple list of application include [38, 39]

- Digital Audio Broadcasting (DAB) is a standard in the European market and OFDM is used for the basis of this standard.
- Asymmetric Digital Subscriber Line (ADSL), High-speed Digital Subscriber Line (HDSL), and Digital Video Broadcasting (DVB) are all global standard that exploit OFDM for their basis.
- OFDM technology is used in the current wireless point-to-point and point-to-multipoint configurations that is Wireless Local Area Networks.
- IEEE 802.11a was published as a supplement to IEEE 802.11 standard and the use of OFDM in the 5GHz band was outlined.

2.4 Interleave-Division Multiple Access (IDMA) system

In order to circumvent the problem of high complexity in MUD associated with CDMA systems, an interleaved-based multiple access scheme was recently studied by Ping *et al* [17] around 2002, proposed and investigated as a multiuser scheme in 2003. It was an extension and special type of well recognized multiple access communication technique CDMA. It has low complexity at the receiver, capable of asynchronous transmission [56], robust against multipath, offers diversity against fading, has high spectral efficiency and improved performance compared to CDMA. A spreading code that is unique is assigned to each user for encoding the information-bearing signal in CDMA systems. The correlation property of the spreading sequences is used to decode the desired signal at the receiver. As the number of users in the system increases, the performance of CDMA degrades rapidly due to multiple access interference (MAI). In order to accommodate more users in the system, there is need for complex techniques to cancel interference which are difficult to implement [57]. IDMA is a promising system that inherits most of CDMA associated advantages such as: mitigation of interference from other cell users in particular and diversity against fading. The various difficulties embedded with CDMA systems are overcome by IDMA. More users are accommodated and its receiver's structure is simple because a very simple and effective chip-by-chip (CBC) and turbo-like iterative MUD strategy is used. The MUD is more relevant and produces efficient detection when the number of users in the system is large [58]. The IDMA chip-by-chip detection is facilitated with interleaver processing to ensure that the codes are spread out and adjacent chips are uncorrelated, thereby preventing strong signals from other users and ensure it does not interfere with the decoding process of a specific user. CBC has a very low computational cost [58], and independent on the users number unlike in the CDMA, OFDM-CDMA and OFDMA schemes [58], it treats both MAI and ISI effectively and the required quality of service is not compromised. The various user-signals are separated at the receiver using the unique interleaver assigned to each. Thus it is named Interleave-Division Multiple Access (IDMA). IDMA scheme dedicates its entire bandwidth to coding using a low rate code for forward error correction (FEC), thus combining coding and spreading operation. According to [10], when the entire bandwidth is dedicated to coding, a multiple access channel (MAC) with perfect capacity is obtainable, hence IDMA has a larger coding gain.

2.4.1 The Transmitter Structure

The component of the conventional IDMA structure constitute the transmitter and receiver (transceiver) part as shown in Fig 2.9. Taking into consideration the transmitter part, with U^{th} number of users transmitting data simultaneously, the concurrent input information bit from diverse users denoted by $user_u$ is first encoded by low rate forward error correction (FEC) code, thus generating a coded sequence $[\mathrm{s}_{\mathrm{U}} = s_u(1), \ldots s_u(n), \ldots s_u(N)]^T$, where N is the subcarrier length.

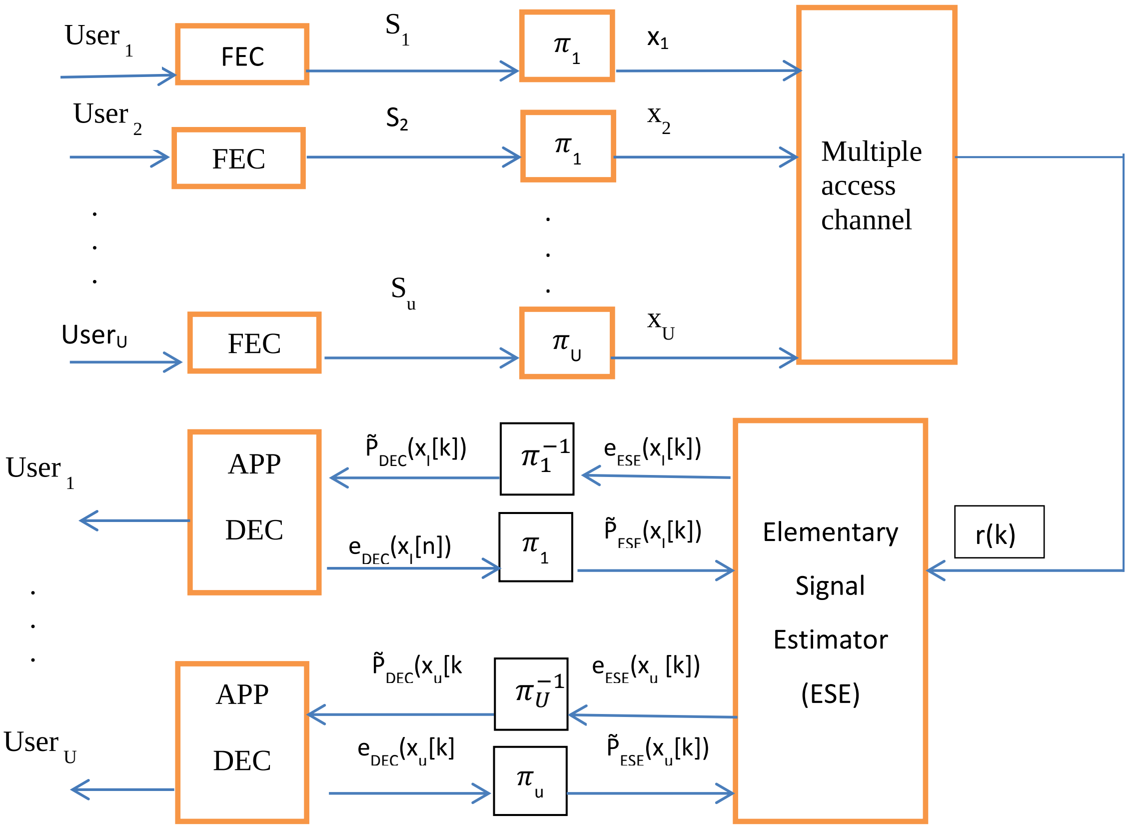


Figure 2.9: IDMA transmitter and receiver structure.

The FEC operation increases the coding gain and control errors of the input data propagating over the fading channel. The output of FEC coding is permutated by a user specific interleaver. These interleavers are different for all users because they are generated at random and independently thereby serve as the means of user separation at the receiver end. They generate a chip sequence $X_{\,u} \equiv \left(x_u(1),\ldots,x_u(n),\ldots,x_u(N)\right)^T$[59]. The interleaved chip sequence X_u, is transmitted to the receiver through a multiple access channel.

2.4.2 The Receiver Structure

According to [60], the receiver structure of IDMA is unique and sub-optimal in nature because it avoids the multiple access and FEC code constraint. The IDMA part of the receiver structure comprises of the Elementary Signal Estimator (ESE) and *a posteriori probability* decoders (DECs). Let $x_u(n)$ represent the nth chip transmitted by user-u and $h_u(n)$, is the fading channel coefficient for an active user-u, The transmitted chip passes over a multipath channel. The received signal $r(k)$ is given as

$$r(k) = \sum_{u=1}^{U} x_u\,(k) \otimes h_u(k) + z(k), \qquad\qquad k\ = 1,2,\ldots,N, \tag{2.5}$$

z(k) is the additive white Gaussian noise (AWGN) with zero mean and variance σ^2. For a particular user-u , the eqn(2.5) becomes:

$$r(k) = x_u(k) \otimes h_u(k) + \beth_u(k)\,, \tag{2.6}$$

where, symbol $\beth_k(k)$ represents distortion which is a combined effect of additive noise z_k and interference due to other users with respect to user u and can be expressed as:

$$\beth_u(k) = r(k) - x_u(k)h_u(k)\ \ , \tag{2.7}$$

$$\beth_u(k) = \sum_{u\,\neq u'} x_{u'}\,(k)h_{u'}(k) + z(k)\,. \tag{2.8}$$

Elementary signal estimator (ESE) and U number of a posteriori probability (APP) decoders (DECs) are the major components at the receiver part of the IDMA system. Each user is entitled to one APP decoder and the operation of ESE and APP decoder in IDMA system is iterative in nature. The process involved in ESE and the APP decoders is explained in section 2.4.3. These are of paramount importance to the IDMA receiver structure. The mean and variance of the signal is determined in the ESE and APP decoder.

2.4.3 The Elementary Signal Estimator and the APP Decoders

The elementary signal estimator (ESE) receives the output $r(k)$ from multipath channel and a coarse soft-in-soft-out chip by chip (CBC) estimation detection is carried out for the initial eradication of interference among simultaneous users in the system. According to [61], the computational cost required for CBC is small and complexity involved is low. At the receiver, the estimated probabilities values of the transmitted chips x(k) are the outputs of the ESE [61]. They are arranged separately with respect to simultaneous users, and later sent to the APP DECs. The ESE-APP DECs operation is iterative and it process its extrinsic information in a turbo-like manner [60]. The output from ESE is fed to the APP DECs which perform APP decoding to remove FEC code constraint. The refined probabilities of the transmitted chips are fed back into ESE. The output of the ESE is improved. At the last iteration, the APP decoders produce hard decisions based on the refined estimations by the ESE and the logarithm likelihood ratio (LLR) estimate of the ESE and the APP decoders is obtained as [60]

$$p\big(x_u(k)\big) \equiv \ log\left[\frac{Pt\ (x_u(k)\ =\ +1)}{Pt\ (x_u(k)\ =\ -1)}\right]\ \ , \tag{2.9}$$

The return of the ESE and the DECs above are estimated probabilities of the LLRs values of the transmitted signals at the receiver [61], which will be distinguished by $\tilde{p}_{ESE}(x_u(k))$, and $\tilde{p}_{DEC}(x_u(k))$, respectively, depending on whether they originate from the ESE or the APP decoders at the receiver. Pt denotes probability. Therefore, considering the multipath channel denoted by the coefficient $h(k)$, the ESE operation employs the received signal $\{r(k)\}$ and $\{\tilde{p}_{ESE}(x_u(k))\}$ the corresponding a posteriori LLR for its operation, so that the resulting output is obtained as [60]

$$p\big(x_u(k)\big) = log\left[\frac{Pt\ \big(x_u(k)\ =\ +1|\ r(k)h_u(k)\big)}{Pt\ \big(x_u(k)\ =\ -1|\ r(k)h_u(k)\big)}\right]\ \ , \tag{2.10}$$

using Baye's theorem

$$p\big(x_u(k)\big) = \ log\left[\frac{Pt\ \big(r(k)|\ x_u(k)\ = +1, h_u(k)\big)}{Pt\ \big(r(k)|\ x_u(k)\ = -1, h_u(k)\big)}\right] + \quad \tilde{p}_{ESE}\big(x_u(k)\big)\ , \tag{2.11}$$

from the first part of (2.11), we have

$$e_{ESE}(x_u(k))\ =\ log[\ \frac{\mathrm{Pt}\ \big(\mathrm{r(k)}|\ \mathrm{x_u(k)}\ = +1, \mathrm{h_u(k)}\big)}{\mathrm{Pt}\ \big(\mathrm{r(k)}|\ \mathrm{x_u(k)}\ = -1, \mathrm{h_u(k)}\big)}\quad . \tag{2.12}$$

Since r(k) is Gaussian, using (2.5), we write (2.11) as

$$Pt\ \big(r(k)|\ x_u(k)\ = +1, h_u(k)\big) = \frac{1}{\sqrt{2\Pi Var(ב_u(k))}} * exp\left(-\frac{\big(\ r(k) - E\big(ב_u(k)\big) - h_u\big)^2}{2Var\big(ב_u(k)\big)}\right), \tag{2.13}$$

similarly

$$Pt\ \big(r(k)|\ x_u(k)\ = -1, h_u(k)\big) = \frac{1}{\sqrt{2\Pi Var(ב_u(k))}} * exp\left(-\frac{\big(\ r(k) - E\big(ב_u(k)\big) + h_u\big)^2}{2Var\big(ב_u(k)\big)}\right), \tag{2.14}$$

using (2.13) and (2.14) in (2.11)

$$e_{ESE}\big(x_u(k)\big) = 2h_u(k).\frac{r(k) - E\big(ב_u(k)\big)}{Var\big(ב_u(k)\big)} \quad . \tag{2.15}$$

The mean and variance of the operation is denoted by E[.] and Var [.], respectively. The central limit theorem is employed and the interference $ב_u(n)$ in (2.6) can be estimated by a Gaussian variable with mean and variance given as [60]

$$E(ב_k(n)) = E\big(y(n)\big) \; - \; E\big(x_k(n)\big)\,, \tag{2.16}$$

and

$$Var\big(ב_k(n)\big) = Var\big(y(n)\big) - Var\big(x_{k'}(n)\big). \tag{2.17}$$

These, i.e, (2.16) and (2.17) can be substituted into (2.15) and we have

$$e_{ESE}\big(x_u(k)\big) = 2h_u(k).\frac{r(k) - E\big(r(k)\big) + E\big(x_u(k)\big)}{Var\big(r(k)\big) - | \;\; h_u|^{\,2} Var\,\big(x(k)\big)} \;. \tag{2.18}$$

The estimated mean and variance of the received signal based on (2.6 and 2.7) is therefore obtained as [60]

$$E\big(r(k)\big) = \sum_{u'=1}^{U} h_{u'}(k)\, E\big(x(k)\big)\,, \tag{2.19}$$

$$Var\big(r(k)\big) = \sum_{u'=1}^{U} | \;\; h_u(k)|^{\,2} Var\,\big(x_{u'}(k)\big) + \sigma^2 \;\; , \tag{2.20}$$

where, $e_{ESE}\big(x_u(k)\big)$ is the extrinsic LLR about the transmitted signal x_u(k) based on the characteristics of the multipath channel and the *a priori* information of simultaneous users in the system [60].

2.4.4 Decoder

Similarly, $p_{ESE(x_u(k))}$ represent the inputs to the DEC for user-u, with the FEC code constraint. Each user is assigned a single decoder, and the extrinsic LLR of each user is fed to its decoder. The decoder carries out the a posteriori probability (APP) decoding using deinterleaved output of the ESE and produce e_{DEC} (x_u(k)) extrinsic LLR output. The a posteriori LLRs about the transmitted chip is given as

$$e_{DEC}\big(x_u(k)\big) = \; \log\left(\frac{Pt[x_u(k) \; = \; +1| \; C, e_{ESE}(x_u(k))]}{Pt[x_u(k) \; = \; -1| \; C, e_{ESE}(x_u(k))]}\right) - e_{ESE}\big(x_u(k)\big), \tag{2.21}$$

$$exp\ (e_{DEC}(x_u(k))) = \ \log\left(\frac{Pt[x_u(k) = +1]}{Pt[x_u(k) = -1]}\right). \tag{2.22}$$

This implies that

$$Pt[x_u(k) = +1] = \frac{exp\ (e_{DEC}(x_u(k)))}{1 + exp\ (e_{DEC}(x_u(k)))} \quad , \tag{2.23}$$

and

$$Pt[x_u(k) = -1] = \frac{1}{1 + exp\ (e_{DEC}(x_u(k)))} \quad , \tag{2.24}$$

since

$$E\big(x_u(k)\big) = (+1)Pt[x_u(k) = +1]\ +\ (-1)Pt[x_u(k) = -1]\ , \tag{2.25}$$

$$E\big(x_u(k)\big) = \ \frac{exp[e_{DEC}(x_u(k))/2] - exp[-e_{DEC}(x_u(k))/2]}{exp[e_{DEC}(x_u(k))/2] +\ exp[-\,e_{DEC}(x_u(k))/2]} \quad , \tag{2.26}$$

$$E\big(x_u(k)\big) = \ tanh(e_{DEC}(x_u(k))/2)\,. \tag{2.27}$$

and

$$Var[(x_u(k))] = E[(x(k) -\ E[x_u(k)])^2] \quad , \tag{2.28}$$

$$Var[(x_u(k))] = 1 -\ (\ E[x_u(k)])^2 \quad . \tag{2.29}$$

2.4.5 The CBC Detection Algorithm

The CBC detection algorithm given by [62]:

Initialization

Set E [x_u(k)] = 0, Set Var [x_u(k)] =1

E [r(k)] = 0 and Var [r(k)] = M + σ^2

Main process

For i =1 to iteration

For m = 1 to M

1. Evaluate $E(\beth_u(k)) = E\big(r(k)\big)\ -\ E\big(x_u(k)\big)$

   And $Var\big(\beth_u(k)\big) = Var\big(r(k)\big) - Var\big(x_{u'}(k)\big)$
2. Update $E(x_u(k)) = \ tanh\ \ (\tilde{p}_{ESE}(x_u(k))/2)$

   And $Var(x_u(k)) = 1\ - (E(x_k(n)))^2$

3. Update $E(u(k)) = \sum_{u'=1}^{U} h_{u'}(k)\, E(x(k))$

   And $Var(r(k)) = \sum_{u'=1}^{U} |\ h_n(k)|\ ^2 Var\ (x_{u'}(k)) + \sigma^2$

   end

   end

2.4.6 Chip-by-chip detection flow chart for IDMA system

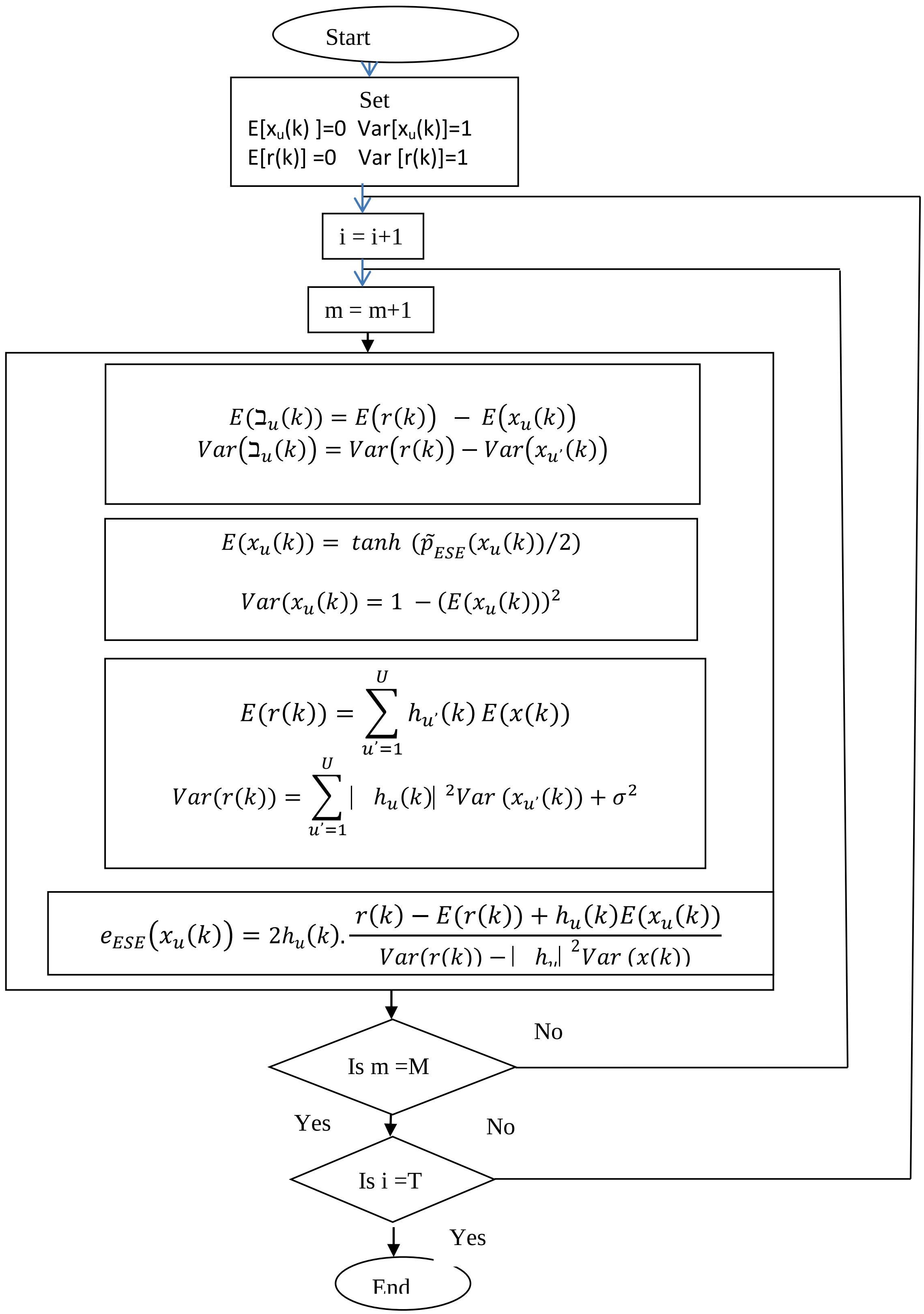


Figure 2.10: Chip-by-chip detection flow chart for IDMA system

2.5 Multiple access techniques comparison

The table below shows the performance comparison between IDMA system and other multiple access schemes. This indicates the potential of IDMA scheme for practical use according to [61].

	TDMA	FDMA	OFDMA	CDMA	IDMA
Multi-user separation by	Time slot	Frequency	Orthogonal frequency	Signature sequence	Interleaver
Methods to combat ISI	Equalization	Cyclic prefix [63]	Cyclic prefix	Rake receive	CBCdetection
High single-user rate can be achieved with		High order modulation		Difficult	Superposition coding rate
Elimination of Intra-cell interference		Not required		Iterative Multi-User-Detection	
inter-cell interference		Sensitive		Mitigated	
Synchronization required		Needed		Not required	

Table 2-1: Comparison of different multiple access scheme.

Comparative analysis among the various multi users is shown in Table 2.1. The table shows the differences in operation of the MAS. The use of reduced spreading factor or adopting multicode CDMA lead to high data rates achievement in CDMA systems. In contrast, FEC code allotment with high coding rate lead to high data rate transmission in IDMA system. CDMA and IDMA system both utilise MUD to mitigate interference. However, CDMA is limited by the number of users and the high cost, while IDMA is not. In a similar manner, IDMA uses the CBC detection algorithm to combat intra-cell interference. The advantage is that the complexity associated with CBC does not depend on the number of users. It achieves multi-user gain in the case of each user with a rate constraint. It supports asynchronous transmission; therefore no complicated synchronization is required on the transmitted data while frame synchronization is required to maintain orthogonality in schemes like TDMA, FDMA, and OFDMA.

2.6 Channel Estimation

CSI is obtained through channel estimation. Four different types of channel estimators are commonly used: the data-aided (DA), the non-data-aided (NDA) (or blind), the semi-blind and the code-aided (CA) estimators. The sections that follow will elucidate on these technique.

2.6.1 Pilot-assisted

Pilot-assisted channel estimation technique can also be referred to as training-based channel estimation. The channel state information (CSI) for communication systems is estimated or identified using training-based data. Pilot-assisted estimation uses pilot symbols which are known data at the receiver. Prior to transmission, at some known position, the pilot symbols are multiplexed with the transmitted information symbols. At the receiver side of the system, pilot symbols are used to estimate the CSI corresponding to the known position. According to [64] two different pilot positions were analysed in the time-domain and frequency domain. The first is referred to as block-type position where the pilot symbols are assigned to just one OFDM block and sent periodically in the time-domain. This is suitable for slow-fading radio channels. It does not require channel interpolation in the frequency domain thus relatively insensitive to frequency selectivity. The second type is comb-type pilot position; the pilot symbols are distributed within each OFDM block uniformly. The properties of the two are summarized in table 2.2.

Block-Pilot	Comb-Pilot
Better performance in Slow Fading channels	Better resistance to fast-fading channels
Insensitive to frequency selectivity	Sensitive to frequency selectivity
Available for only one OFDM block	Uniformly distributed within each OFDM block
Lower transmission rate	Higher transmission rate
It does not require channel interpolation.	It requires channel interpolation

Table 2-2: Comparison of Block and Comb pilot assisted channel estimation.

In wireless communication systems, pilot symbol assisted modulation (PSAM) is a special form of pilot assisted channel estimation techniques, used to mitigate the effect of fading. A close-form expression can be obtained and which is an advantage of pilot-training based channel estimators.

Disadvantages of pilot-assisted channel estimation

Pilot-Training based channel estimators requires the transmission of pilot symbols which occupies the spectrum and leads to a loss in terms of spectral and power efficiency. Channel estimation also depends strictly on the pilot symbol information alone. The estimate is used for extracting information when the payload arrives and there is introduction of unresolved error due to interpolation techniques deployed.

2.6.2 Blind channel estimation techniques

These estimators are referred to as blind channel estimators because the inherent information in the received signals are employed in the estimation instead of using the pilot symbol that consume and wastes the valuable channel bandwidth. It requires a large number of received symbols between hundreds to thousands to be processed before it can converge to a reliable channel estimate. Blind techniques make use of the structural and statistical properties of the data transmitted. They exploit all the received observations and assume hypotheses on data. There are two types of blind channel estimation techniques which are the statistical and deterministic channel estimation techniques. In the statistical method, the channel is estimated using the cyclic statistical properties of the received signals. It processes the pre-DFT received data and use the cyclostationarity induced data by the Cyclic Prefix (CP). Deterministic type use the received signals and the channel coefficients instead of the statistical properties of the received signals [65]. It processes the post DFT signal and exploits the finite-alphabet property of the information bearing symbols. This means that the deterministic method is applied on the received OFDM symbols after DFT demodulation. Comprehensive classification of blind channel estimation methods and their relative advantages, disadvantages and performance are established in reference [66]. They are summarised below as follows:

2.6.2.1 Deterministic methods

Deterministic methods consider that both the unknown input symbols and the channel coefficients are deterministic quantities. These methods leave ambiguity in the channel [66]. They have lower performance than second-order statistical (SOS) methods and higher-order statistical (HOS) methods. According to [66] these methods possess the remarkable property of providing, in the noiseless case and with a finite amount of data, the exact channel.

2.6.2.2 Statistical methods

The Statistical approach can be divided into second-order statistical (SOS) methods and higher-order statistical (HOS) methods. Second-order statistical (SOS) methods only exploit the first- and second-order moments (i.e. mean and covariance matrix) of the received signal. Frequency domain and time domain are the two categories of statistical method. Higher-order statistical (HOS) methods exploit not only the first and second moments of the received signal but also higher-order moments. Consequently, they perform better than SOS methods, but have a higher complexity and higher variance. In various works published in [67], the higher-order statistics of the received signals are exploited for channel estimation but there is high complexity in the computation process due to large number of data samples that are used. SOS mitigate the problem of HOS by exploring second-order cyclic statistics of the over-sampled channel output [68].

The table below highlights the difference between the deterministic and the statistical methods.

	Deterministic Methods	Statistical Methods
Convergence	Faster (high speed of convergence)	Slow
Complexity	Very High [34]	High
Finite Data Effect	Finite	Suffer from finite data effect when dealing with an extremely short sample sequence [69]
Performance	Lower performance	Better performance
Ambiguity in Channel	yes	yes

Table 2-3: Difference between Deterministic methods and Statistical Methods.

Advantages of Blind Estimation

- There is no use of pilot symbol (training-data), hence no wastage of bandwidth and therefore high spectral efficiency is achieved [70].

Disadvantages of blind estimation

- They have a slow convergence rate
- A slowly time-varying channel is required and this is impractical for wireless communication systems where the channel may be fast time-varying.
- It presents a high sensitivity to channel order over estimation**.**

- There is higher computational complexity.
- There are ambiguities in channel estimation in the case of multi-users.
- The method is not suitable for fast-time varying channel like mobile wireless communication systems.

2.6.3 Decision –Directed Channel Estimation techniques (DDCE)

In DDCE techniques, the channel is estimated to obtain the CSI of the system using all the pilot symbols and the re-modulated detected data symbols are considered as pilot symbols and employed for channel estimation [71]. Though DDCE can be referred to as a special type of pilot assisted estimation, it provides a better performance than pilot assisted estimation. Comparing DDCE with pilot assisted estimation, DDCE uses almost all its pilot information symbols while its counterpart pilot assisted estimation use few pilot symbols for the same estimation. An accurate channel transfer function will be obtained when there is no transmission error (symbol error) and channel fluctuation rate is minimal. Similarly, few known pilot symbols are used, therefore it is bandwidth efficient. Literature review on different works on DDCE was carried out in [72]. Assumption of prior knowledge of the noise and channel covariance was employed by Van de Beek et al [73] but this assumption does not hold in real world. A singular value decomposition was used in [74] as well. The method required the knowledge of the channel frequency correlation which might not be available thus constitute a problem. The ideal algorithm proposed in [75] for both OFDM and MC-CDMA is not feasible in realistic channel conditions and it has a major drawback. The method used in DDCE can either be hard or soft iterative in nature. A soft output from the equalizer are fed back to the channel estimator and treats it as known bits, which then produces a better estimate. The iterative soft feedback is preferred to the hard feedback because it avoids propagation error by allowing to properly weight reliable and unreliable symbols [76]. This means that DDCE exploits the turbo principle to produce accurate channel estimate.

2.6.4 Semi-blind channel estimation techniques

This technique combines the effort of both pilot-assisted estimation technique and blind estimation techniques, where the intrinsic information in the unknown data symbols and the known pilot information are used for channel estimation. Using the same number of pilot symbols, semi-blind estimation techniques perform better than pilot based techniques. Semi-blind techniques solve the uncertainty problem associated with blind estimation using few pilot symbols. Some literatures have investigated semi-blind estimation using the subspace method. According to [77, 78], there is high computational complexity associated with the subspace method. Also, in the study in [79], linear prediction was used to estimate the blind constraint while the matrix A was estimated using the least square (LS) algorithm. The use of semi-blind channel estimation in single input multiple output (SIMO) systems achieved good performance using the subspace method and it has a simple structure [80] but its application in multiple input multiple output (MIMO) system is not too successful because it can only estimate channel subject to a polynomial matrix ambiguity [81]. Some other literature state that estimations are based on second-ordered statistics of a long vector, therefore there is need for a large number of OFDM symbols to estimate the correlation matrix and this is not suitable for fast time-varying channels. According to [81] the subspace method is not practical for general MIMO channel estimation. The subspace algorithm in [79] is limited to MIMO-OFDM systems. Semi-blind estimation was reviewed in [72]. It analyses the various semi-blind channel estimation including the estimation on MIMO systems. Parallel data and training signal algorithm is developed in [82] and [83], block pre-coded space time OFDM transmission is presented. The least square estimator based on known pilot sequence is analysed in [84] and the statistical structure of the observation is used in the estimation. The first and second order statistic are used in [85] to estimate the channel.

	Pilot-assisted	Blind	Semi-blind	DDC
Bandwidth Wastage	Yes	Spectrally efficient	No	Less Efficient
Ambiguity	None	Yes	Solves the Ambiguity of Blind systems	None
Performance	Good	Good	Better	Best
Known Pilot data	large	None	Few	Very few
Convergence	Fast	Slow	Fast	Fast

Table 2-4: Different channel estimation comparisions.

Table 2.4 gives a general overview of the performance of the different channel estimation methods and figure 2.11 shows their classification.

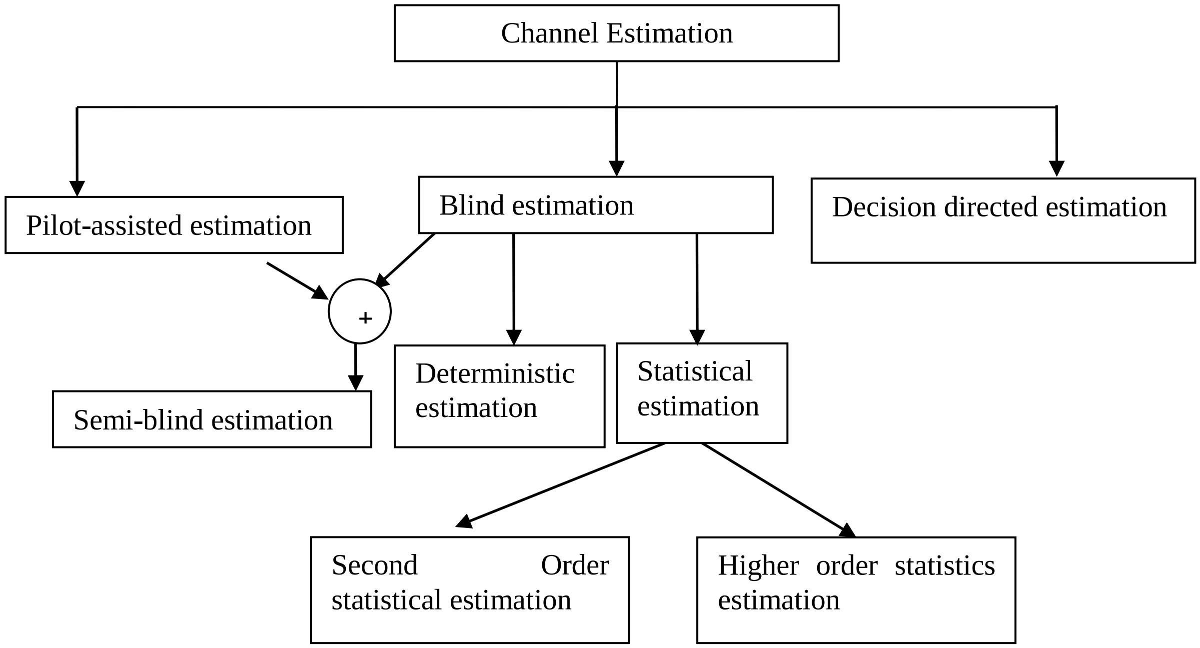


Figure 2.11: Classification of channel estimation techniques.

2.7 Chapter Summary

This chapter started with the introduction and discussion of the various MAS techniques. It elucidate how different users share the available spectrum. OFDM was introduced as a multicarrier modulation scheme. The various operations of the scheme were discussed and compared with FDM and concluded that OFDM saves 50% of the spectrum. IDMA was introduced and the principles of operation of IDMA scheme were established. In a similar manner, with iterative turbo-type multi-user detector IDMA overcomes the MAI problem efficiently. The detailed functions of ESE APP decoder at the receiver end were analysed. It showed the various detection steps involved in the coarse CBC to roughly remove from the system the existing interference among users, until the last iteration when the a posteriori probability (APP) based decoders produced the hard decision on the transmitted chips [60]. Likewise, four different MA schemes were compared based on parameters such as: Separation of users, elimination of inter and intra-cell interference, synchronization requirement etc. Amid the comparison discussed, IDMA features demonstrate that it is suitable to support media services in broadband network for 4G wireless communication. The chapter concludes with discussion on the different type of channel estimation techniques. Their differences were discussed based on the different short comings associated with each of the estimation techniques. A reasonable conclusion on the recommended channel estimation method implemented in this thesis was made.

CHAPTER 3

OFDM-IDMA CHANNEL MODEL AND ESTIMATION

3.1 Channel Modelling

The transfer characteristic of the physical medium which are derived from the observed characteristic of the received signal and the sum of the multipath components impinging at the receiver can be represented mathematically and is referred to as the channel model. Figure 3.1 represent the physical representation of the channel.

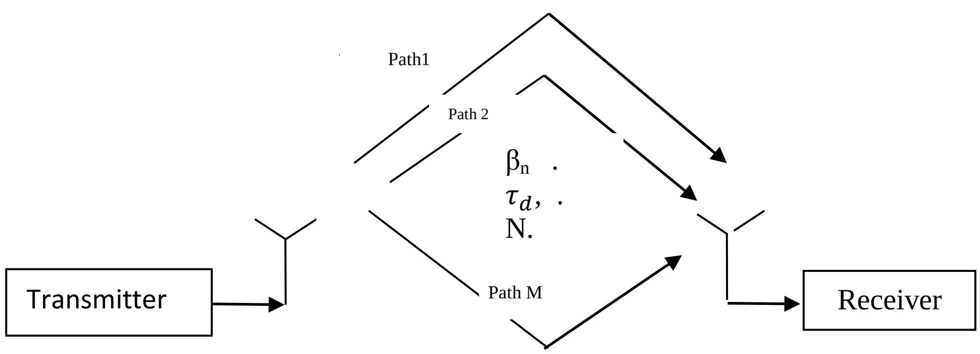


Figure 3.1: Multipath fading channel.

A channel can either be modelled as time variant or invariant. It can be modelled in time domain or frequency domain. The channel model used in this thesis is a time variant model characterised by multipath fading where the transmitted signal s(t) and its equivalent complex signal x(t) is given as

$$s(t) = Re\{x(t)e^{jw_c t}\}, \tag{3.1}$$

where, the real part is denoted by Re{}, the carrier frequency is $w_c = 2\pi f_c$. Additive white Gaussian noise with zero mean and variance corrupt the baseband form of the received signal z(t) which is represented by [86]

$$z(t) = \int_{-\infty}^{\infty} h(t;\, \tau\,)x(t\, -\, \tau\,)d\tau\, +\, n(t), \tag{3.2}$$

where, h(t;τ), is the the time-varying channel response (CIR) at an instant time t, and $n(t)$ is the additive white Gaussian noise. In mobile wireless communication, the used CIR model is given as [87]

$$h(t;\ \tau\ ) = \sum_{n=0}^{N-1} \beta_n(t) e^{-jw_c \tau_d} \delta(\tau - \tau_n), \tag{3.3}$$

where, β_n is the propagation paths gain, τ_d is the delay associated with the nth path and N is the number of paths in the channel, as shown in figure 2.1.

Channel Transfer Function (CTF) is obtained using the Fourier transform of h(t; τ) given by [86]

$$H(\,f\ ;\ \tau\ ) = \int_{-\infty}^{\infty} h(t;\ \tau\ )\, e^{-j2\pi f \tau} d\tau\ . \tag{3.4}$$

3.2 The OFDM transmitter model

The summation of the prototype burst which changes in time and frequency domain and multiplied by the data symbols is the mathematical expression of the OFDM signal. The OFDM k_{th} symbol is written in continuous-time notation.

$$S_{R,k}(t - kT) = \begin{cases} Re\left\{ w(t-kT) \displaystyle\sum_{i=-N/2}^{N/2-1} x_{i,k} e^{j2\pi\left(f_c + \frac{i}{T_{FFT}}\right)(t-kT)} \right\} \\ for\ \ kT - T_{win} - \ T_{guard}\ \le t\ \le kT + T_{FFT} + T_{win} \\ 0 \qquad\quad otherwise \end{cases} \tag{3.5}$$

The complete list of symbols is given here as:

T: The time interval between two successive OFDM symbols; symbols period.

T_{FFT}: The functional part of the OFDM symbol; FFT time.

T_{guard}: Cyclic prefix length.

T_{win}: Spectral shaping windowed prefix or postfix length; Window interval.

f_c: Centre frequency of the occupied frequency spectrum.

F= 1/ TFFT: Adjoin SCs with their frequency spacing.

N: Number of point simulated by FFT points; FFT period.

k : Transmitted symbol Index.

i: Index on SC; $i \in \{-N/2, -N/2+1, \ldots -1, 0, 1, \ldots N/2-1\}$.

$x_{i,k}$**:** Constellation signal point; the modulated complex data symbol on the i_{th} SC of the k_{th} OFDM symbol.

w(t) **:** Pulse shaped transmitted waveform is given as:

$$w(t) = \begin{cases} \frac{1}{2}\left[1 - cos\pi\left(t + T_{win} + T_{guard}\right)/T_{win}\right], & -T_{win} - T_{guard} \le t \le -T_{guard} \\ 1, & -T_{guard} \le t \le -T_{FFT} \\ \frac{1}{2}\left[1 - cos\pi\left(t - T_{FFT}\right)/T_{win}\right., & -T_{FFT} < t \le T_{FFT} + T_{win} \end{cases} \quad (3.6)$$

The sequence of the transmitted symbols in its continuous form is given as

$$S_R(t) = \sum_{k=-\infty}^{\infty} S_{R,k}(t - kT). \quad (3.7)$$

The equivalent complex low-pass transmitted signal can be derived from equation (3.5) to (3.7). The complex envelope of the OFDM signal is given as

$$s(t) = \sum_{k=-\infty}^{\infty} S_k(t - kT), \quad (3.8)$$

where,

$$s_k(t - kT) = \begin{cases} w(t - kT) \sum_{i=-N/2}^{N/2-1} x_{i,k} e^{j2\pi\left(\frac{i}{T_{FFT}}\right)(t-kT)} \\ for \;\; T - T_{win} - T_{guard} \le t \le kT + T_{FFT} + T_{win} \\ 0 \qquad otherwise \end{cases} . \quad (3.9)$$

and equation (3.10) is the comparison of this expression with Fourier series.

$$v(t) = \sum_{n=-\infty}^{\infty} c(nf_0) e^{j2\pi n f_0 t}, \quad (3.10)$$

It sums the entire sine and cosine waves of amplitudes $c(nf_0)$ stored in constellation signal points X[k] array is given by

$$v(t) = \sum_{k=0}^{N-1} c(nf_0)\left\{\cos\left(\frac{2\pi nk}{N}\right) + j\sin\left(\frac{2\pi nk}{N}\right)\right\}. \quad (3.11)$$

The complex valued constellation signal points $x_{i,k,}$ are denoted by complex-valued Fourier coefficients $c(nf_0)$ and the frequencies nf_0 are equivalent to the SC frequency spacing $1/T_{FFT}$.

In a digital system, IDFT is used to generate the modulated waveform. The IFFT is used to compute IDFT due to its computational efficiency. From equation (3.11), IDFT modulate a series of complex exponential carriers with a different symbol from the information array $c(nf_0)$, to produce N samples of a time domain signal as shown in Figure 3.2. The input to this IFFT are the data constellations $x_{i,k}$ and the time domain OFDM symbol v(t) is its output.

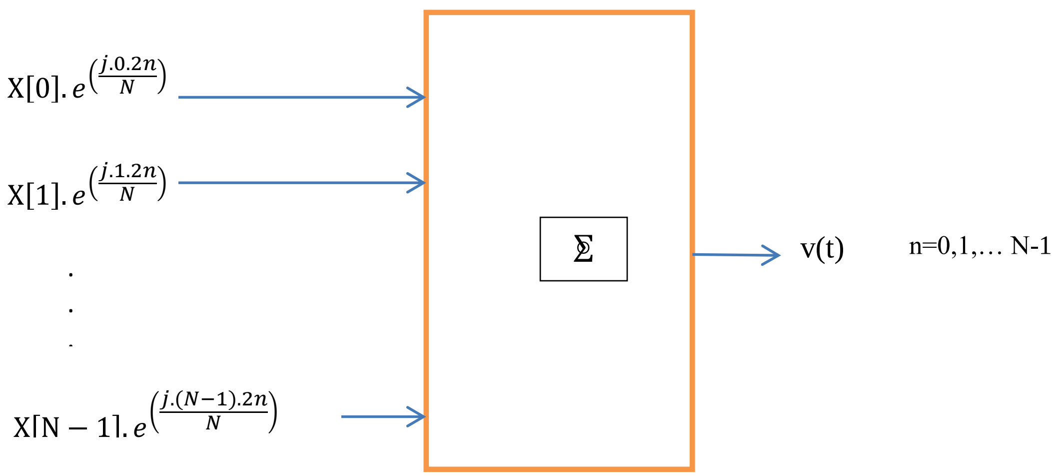


Figure 3.2: IDFT formation using a summation.

3.3 OFDM receiver model

OFDM signal is demodulated using a bank of filters, which are "matched" to the functional part $[kT, kT+T_{FFT}]$ of the OFDM symbol. The inverse of the Fourier operation which is the opposite operation to (3.10), occur at the receiver end. The time domain signal v(t) is converted back to its equivalent frequency domain. The removal of the Fourier coefficients $c(nf_0)= x_{ik}$ from the time domain signal v(t)X= (r(t) + n(t)), exactly formulates such a bank of matched filters is given

$$c(nf_0) = \frac{1}{T_0}\int_{T_0} v(t)e^{j2\pi n f_0 t}dt \tag{3.12}$$

where, T_0 is equal to T_{FFT} and it represents the integration period. These filters are realized using DFT or FFT in a digital systems. The amplitudes and phases of sine and cosine waves are determine by the FFT demodulator using N times domain transmitted samples and forming the received signal, according to the equation

$$c(nf_0) = \sum_{k=0}^{N-1} v(t)\left\{\cos\left(\frac{2\pi nk}{N}\right) + j\sin\left(\frac{2\pi nk}{N}\right)\right\}. \tag{3.13}$$

Suppose one has an idea of the precise timing kT at which OFDM symbols begin, we can obtain the transmitted signal constellations $x_{i,k}$ from the received signal r(t) and derive the receiver signal constellations which are denoted by $y_{i,k}$.

$$y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} r(t)e^{-j2\pi i(t-kT)/T_{FFT}}dt\ ,$$

$$y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} \left[\int_{\tau=0}^{\tau_{max}} h_k(\tau)s(t-\tau)d\tau + n(t)\right] e^{-j2\pi i(t-kT)/T_{FFT}}dt\ . \tag{3.14}$$

Due to the area of variation between the upper and lower integral of equation (3.14) and the fact that T_{max} is smaller than T_{guard}, there is no impact on the adjacent OFDM transmitted symbols, and s(t) can be substituted with $s_k(t)$ [see (3.9)]

$$y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} \left[\int_{\tau=0}^{\tau_{max}} h_k(\tau) \sum_{i=-N/2}^{N/2-1} x_{i',k} e^{j2\pi\left(\frac{i'}{T_{FFT}}\right)(t-kT-\tau)} d\tau\right] e^{-j2\pi i(t-kT)/T_{FFT}}dt + \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} n(t)e^{-\frac{j2\pi i(t-kT)}{T_{FFT}}}dt\ . \tag{3.15}$$

It should be noted that in the rage of integration $w(t - KT)$ =1, consequently, in this equation the window was excluded. The second integral in (3.15) gives rise to n_{ik} as samples of independent additive noise because the complex exponential terms are orthogonal functions. If integration and summation order are interchanged and notation is simplified by substituting $u = t - KT$, the derived equation (3.15) becomes;

$$y_{i,k} = \sum_{i'=-N/2}^{N/2-1} x_{i',k} \frac{1}{T_{FFT}} \int_{u=0}^{T_{FFT}} \left[\int_{\tau=0}^{\tau_{max}} h_k(\tau) e^{-j2\pi i'(u-\tau)/T_{FFT}} d\tau \right] e^{-j2\pi i u/T_{FFT}} du + n_{i,k}$$

$$= \sum_{i'=-N/2}^{N/2-1} x_{i',k} \frac{1}{T_{FFT}} \int_{u=0}^{T_{FFT}} \left[\int_{\tau=0}^{\tau_{max}} h_k(\tau) e^{-j2\pi i'\tau/T_{FFT}} d\tau \right] e^{-j2\pi(i-i')u/T_{FFT}} du$$

$$+ n_{i,k} \ . \qquad (3.16)$$

From the second expression of equation (3.16), the inner integral denotes the Fourier transform of h_k (τ) at instantaneous frequency $i'/ T_{FFT}= i'F$. It is the sampled channel Transfer Function at time kT. The channel coefficient can be expressed as

$$h_{i'k} = FT\{h_k(\tau)\} = \int_{\tau}^{\tau_{max}} h_k(\tau) e^{-j2\pi i'\tau/T_{FFT}} d\tau = H(i'F, kT) \ , \qquad (3.17)$$

The filter bank receiver output simplifies to

$$y_{i,k} = \sum_{i'=-N/2}^{N/2-1} x_{i',k} h_{i',k} \frac{1}{T_{FFT}} \int_{u=0}^{T_{FFT}} e^{-j2\pi(i-i')u/T_{FFT}} du + n_{i,k} \ . \qquad (3.18)$$

In equation 3.16, the integral value can be equated to 1 only if $i = i'$. At all other instance of i, the integral values becomes zero that is $i \neq i'$. Therefore, the received signal is finally given as

$$y_{i,k} = x_{i,k} h_{i,k} + n_{i,k} \ . \qquad (3.19)$$

From the received signal output given by equation (3.19), it is clear that a perfect synchronized OFDM system is perceived as a set of parallel Gaussian channels. Also, the equation shows that y_{ik} is a mixture of attenuation/amplification and phase rotation according to the channel coefficients h_{ik} which were introduced by the multipath channel. In order to recover the data integrated in these signal constellations, channel estimation is required.

3.4 Introduction to MC-IDMA

According to various literatures on the subject, MC-CDMA is robust to multipath propagation but as the number of users increases the system performance deteriorates rapidly because of the multiple access interference in the system. Different multiuser detection techniques have been presented to combat MAI. However, their implementation is practically difficult because of their high computational complexity [21]. The same applies to the case of MC-CDMA systems in which the OFDM transmission scheme is combined with the CDMA multiple access scheme and MC-IDMA in which the OFDM transmission scheme and IDMA multiple access technique are combined [12] and [16]. Each MAS such as OFDM-CDMA, CDMA, OFDMA, and IDMA has its various merits and demerits. MC-IDMA inherits the majority of the merits of each MAS and avoids their individual demerits [16]. The transmitted sequences from all the available users are almost un-correlated in MC-IDMA. Therefore, the matrix operations which are used in MC-CDMA systems are not required in MC-IDMA because the MUD structure is simple, thus attains optimal performance, and the number of users in the system does not affect the system complexity [16]. MC-IDMA is an efficient system to mitigate cross-cell interference, MAI and combats ISI over a multipath channel. It has low complexity because CBC is involved, reaches a high spectral efficiency, and supports high data rate transmission [16]. The multicarrier scheme ensures an enhanced cellular performance, higher diversity order, and spectral efficiency with associated low cost MUD [12]. Below is the development MC-IDMA system model.

3.4.1 MC-IDMA system model

Consider an MC-IDMA system for uplink transceiver structure with U simultaneous users as shown in figure 3.3. The active user in the system is denoted by u. The data is subjected to forward error correction producing the encoded data sequence C_u. Each data sequence is then interleaved by a specific interleaver to generate the sequence X_u. Interleavers disperse the coded sequence so that adjacent chip sequences are uncorrelated.

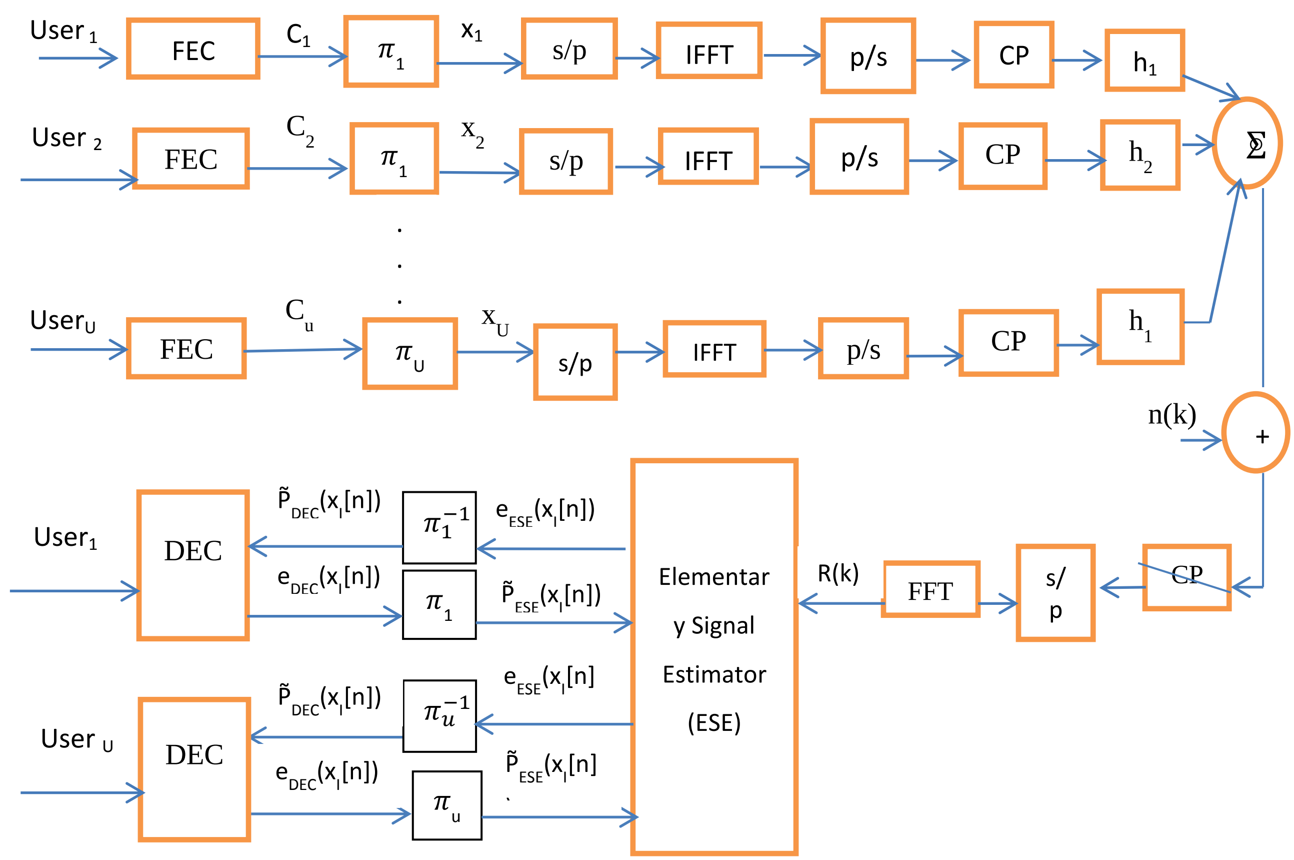


Figure 3.3: MC-IDMA transciever system.

$$y = \sum_{U} y_u + n = \sum_{U} h_u * x_u + n \, , \qquad (3.20)$$

where, * denotes the convolution and n denote the additive white Gaussian noise. At the receiver side, the various users are distinguished by their specific interleaver, the received signal is sampled, removal of cyclic prefixes take place, serial to parallel conversion takes place, then Fast Fourier Transform (FFT) coherently demodulates the received signal. The demodulated signal is

$$R(k) = \sum_{u} H_u(k) x_u(k) + N(k) \, , \qquad (3.21)$$

N(k) is the additive white Gaussian noise (AWGN) with zero mean and variance σ^2, and $H_u(k)$ is referred to as the channel gain of the uth subcarrier for user u. Let the uth user be our desired user, we can rewrite (3.21) as:

$$R(k) = \sum_{u=1}^{U} x_u\,(k) H_u(k) + ב_u(k), \qquad k = 1,2,\ldots,N, \tag{3.22}$$

For a particular user-u , the eqn (3.22) becomes

$$R(k) = x_u(k) H_u(k) + ב_u(k), \tag{3.23}$$

where, symbol $ב_u(k)$ represents distortion which is a combined effect of additive noise z(k) and interference due to other users with respect to user u and can be expressed as

$$ב_u(k) = R(k) - x_u(k) H_u(k)\,, \tag{3.24}$$

$$ב_k(n) = \sum_{u \neq u'} x_{u'}\,(k) H_u{}'(k) + N(k). \tag{3.25}$$

Equations (3.22 – 3.25) are the same as equations (2.5 – 2.8) of chapter two. These similarities indicate that the receiver principles outlined for IDMA systems earlier can be directly applied to MC-IDMA systems. Also, at MC-IDMA receiver, a multiuser elementary signal detector is used to process the interference rejection and U^{th} decoder block for decoding. The MC-IDMA ESE and APP decoder operation is the same as for IDMA operation.

3.5 Channel Estimation MC-IDMA system

The high spectral efficiency, the high capacity and robustness to frequency selective fading possessed by MC-IDMA has evoked a great deal of recognition and those features present MC-IDMA as among efficient technologies. The nature of a radio channel is usually frequency selective and time variant. The channel transfer function for an MC-IDMA mobile communication system, at different subcarriers appears uneven in both frequency and time domains. Hence, the need for the channel to be dynamically estimated is necessary. Its substantial iterative multiuser detector is used to subdue the MAI at the receiver and it assumed that Channel State Information (CSI) is known [12, 16]. In the area of channel estimation for MC-IDMA system, not much work has been done. Among the various channel estimation algorithm developed, according to literature, the LMS algorithm is used in [18, 19]. Excessive mean square error is a limiting factor of this algorithm. Slow and data dependence convergences are other limiting factors. Other MC-IDMA channel estimation algorithms are given in [18-21]. In mobile communication, channel estimation is a burning issue. It is relatively simple for the case of downlink than for the case of uplink because all the signals from the base station pass through the same channel to the users. Channel estimation at the uplink systems is more general and challenging. All that is subsequent to multiple users who transmit their signals through different channels before arriving at the base station. The channel estimation algorithm proposed in [22] is adopted and modified in this thesis, also, an efficient semi-blind channel estimation algorithm for MC-IDMA for uplink systems is developed. The algorithm completely identifies the channel correlation matrix with few orthogonal pilot symbols and estimates the channel instead of using pilot symbols alone. Least Square Error (LSE) is employed for initial estimation of the system at pilot position and MMSE for the estimation of the system. From this algorithm, modified MMSE algorithms are developed. Bits error rate (BER) versus signal to noise (SNR) ratio simulation results show that the adopted proposed algorithm performs better when adapted to MC-IDMA uplink system and the modified algorithms outperforms the conventional adopted MMSE algorithm.

3.5.1 The Least Square Estimation

The initial channel estimation employed in this thesis is the Least Square (LS) estimator. At this initial state, estimation of the channel impulse response h_{LS} is done employing the pilot symbols. The LS channel estimator minimizing $h = arg_h \min\|y - Mh\|^2$ which reduces to equation (3.26) following [88] is given as

$$\mathrm{h_{LS}[k] = x_p^H y[k]}\,, \tag{3.26}$$

where, $\mathrm{x_p[k]}$ represents pilot training symbols used for initialization of the semi-blind channel estimation and subscripts ^{H} represent the Hermitian transpose. Equation (3.26) can be written as

$$\mathrm{h_{LS}[k]} = \frac{\mathrm{y}[k]}{\mathrm{x_p[k]}}. \tag{3.27}$$

Recall, y[k] is the time domain received signal, equation (3.27) can be written as

$$\mathrm{h_{LS}[k] = h_u[k]} \times \left(\frac{\mathrm{x_u[k]}}{\mathrm{x_P[k]}}\right) + \frac{\mathrm{z[k]}}{\mathrm{x_P[k]}} \quad . \tag{3.28}$$

If it is assumed that the estimation is error free, we then have $x_u[k] = x_p[k]$ and (3.28) becomes:

$$\mathrm{h_{LS}[k] = h_u[k]} + \frac{\mathrm{z[k]}}{\mathrm{x_P[k]}} \tag{3.29}$$

with the aid of FFT operation on $h_{LS}[k]$, the initial estimate $H_{pLS}[k]$ is obtained as:

$$H_{pLS}[k] = \sum_{m=0}^{N-1} h_{PLS}[k,m] e^{-j2\Pi km/N} \tag{3.30}$$

3.5.2 MMSE-Based Channel Estimation

In this section, the estimation of channels for all the users is done together taking into consideration inter-user correlation. According to [89, 90] the Minimum Mean Square Error (MMSE) approach is based on the knowledge of the auto-correlation matrix of the channel frequency response H. It estimate it as AH_{LS}, where the mean-square estimation error is minimized using A as the weighting matrix and $\mathrm{H_{pLS}}$ as the orthogonal pilot sequence. The estimation is given by

$$h = \min_{A} E\left\{\left\|H - AH_{pLS}\right\|^2\right\} \tag{3.31}$$

$E\{\cdot\}$ corresponding to the expected value. The solution of A is:

$$A = R_{HH}(R_{HH} + \delta_n^2(XX^H)^{-1})^{-1},$$

where, $R_{HH} = E\ (HH^H)$ represent the correlation matrix of the channel, and XX^Hrepresent matrix with diagonal entries.

The estimation from MMSE of H is thus given as:

$$H_{\mathrm{MMSE}} = R_{HH}(R_{HH} + \delta_n^2(XX^H)^{-1})^{-1}\mathrm{H_{P,LS}} \tag{3.32}$$

Since $X^H X = I$, then (3.32) becomes

$$H_{MMSE} = R_{HH}(R_{HH} + \delta_n^2 I)^{-1}\mathrm{H_{P,LS}} \tag{3.33}$$

3.5.3 Semi-blind estimation (MMVE Algorithm)

The novel MMVE algorithm was derived in [22] and it was adopted to an OFDM system. In this thesis, MMVE algorithm is adopted to a Multicarrier-Interleave Division Multiple Access (MC-IDMA) system. A semi-blind channel estimation approach is obtain using the combined information and fact from the above $\mathrm{H_{MMSE}}$ and eigenvalue decomposition (EVD) of the channel autocorrelation matrix. The semi-blind channel estimation algorithm is termed Minimized Mean Value Estimator (MMVE).

The algorithm is summarized as follows:

The channel autocorreltaion matrix is given by

$$R_{HH} = R_{YY} - R_w\ , \tag{3.34}$$

where, R_w is the noise covariance matrix given by $R_w = \delta_n^2 I$, and R_{yy} is the output autocorrelation matrix. The EVD technique is applied on the autocorrelation matrix and the corresponding eigenvalue and eigenvectors were obtained as:

$$R_{yy} = \quad U \wedge U^H \, , \tag{3.35}$$

$$R_{yy} = \; R_{HH} + \delta_n^2 I \, , \tag{3.36}$$

$$R_{yy} \; = \; R_{HH} + \delta_n^2 U \, U^H$$

where, U is a unitary matrix and $\wedge = diag\{\lambda_1, \lambda_2, \dots, \lambda_i, \dots, \lambda_k\}$
From equation (3.36),

$$R_{HH} = R_{YY} - \delta_n^2 I \, . \tag{3.37}$$

Recall MMSE estimator of equation (3.33)

$$H_{MMSE} = R_{HH}(R_{HH} + \delta_n^2 I)^{-1} \mathrm{H}_{\mathrm{PLS}} \tag{3.33}$$

Substituting (3.37) into (3.33), the semi-blind minimized mean value estimator (MMVE) is built.

$$H_{MMVE} = \; (R_{YY} - \delta_n^2 I) \, [(R_{YY} - \delta_n^2 I) + \delta_n^2 I \,]^{-1} \tag{3.34}$$

$$H_{MMVE} = \; (R_{YY} - \delta_n^2 I) \, [R_{YY} - \delta_n^2 I + \delta_n^2 I \,]^{-1} \tag{3.34b}$$

$$H_{MMVE} = \; (R_{YY} - \delta_n^2 I) \, [R_{YY} \,]^{-1} \tag{3.34c}$$

Substituting the decomposed autocorrelation matrix $R_{yy} = \quad U \wedge U^H$, it gives

$$H_{MMVE} = \; U(\wedge - \delta_n^2 I) U^H (U \wedge U^H)^{-1} \mathrm{H}_{\mathrm{p,LS}} \tag{3.35}$$

Since $U^H U = I$

$$H_{MMVE} = U diag \left(\frac{\lambda_i - \delta_n^2 I}{\lambda_i} \right) U^H \mathrm{H}_{\mathrm{p,LS}} \tag{3.36}$$

From (3.36), represent $\frac{\beta}{SNR} = \delta_n^2 I$

$$H_{MMVE} = U diag \left(\frac{\lambda_i - \beta / SNR}{\lambda_i} \right) U^H \mathrm{H}_{\mathrm{P,LS}} \tag{3.37}$$

SNR is obtained from:

$$\delta^2 = N_0/2 \quad (3.38)$$

$$N_0/2 = 2R_c M(SNR) \quad (3.39)$$

Therefore,

$$SNR = (N_0 R_C M)^{-1} \quad (3.40)$$

N_0 = Noise power spectral density,

M = Number of bit per symbols, M = 2 for QPSK ,

R_C = the coding rate of channel encoding, and

SNR = Signal to noise ratio.

$$H_{MMVE} = Udiag\left(\frac{\lambda_i - 1/SNR}{\lambda_i}\right)U^H \mathrm{H}_{\mathrm{PLS}} \quad (3.41)$$

3.6 Simulation Result and Discussion

In this section, a comparative computer simulation result of the proposed MMVE algorithm-based semi-blind channel estimator are presented to verify the scheme effectiveness and compare it with the Least mean square estimator and known channel which serves as the benchmark. The Bit Error rate (BER) versus Signal to Noise ratio (SNR) results, as well as the MSE versus SNR simulated results describe and analyse the performance of the system. It employs QPSK modulation technique with 2 GHz operating carrier frequency. The MC-IDMA system model used 32 input data length, 8 spreading length for different instances of users. The number of sub-carriers is N = 64, and for the convenience, all the users in the simulations are assumed to have equal number of sub-carriers. The channel was subject to 16-paths i.e M = 16. The normalized Doppler frequencies of the Rayleigh fading channel is fDn = 0.0045, and fDn = 0.1085 with mobile speeds equivalent of 5km/h, and 120km/h, respectively. The general performance of the system is investigated for a constant slow fading scenario with mobile speed of 5km/h and then at increasing mobile speeds of 120km/h. The number of users used in the simulation also serves as a means through which the systems are compared.

Fig. 3.4 illustrates the system performance of the semi-blind scheme on an OFDM-IDMA system. The performance is illustrated in terms of its Bit Error Rate (BER) for a very fast mobile speed of 120km/h. The proposed scheme is compared to that of the LMS scheme using the impact of BER on SNR. The proposed and the LMS schemes have the same number of users and the known channel is used as the bench mark. It can be seen from this that the BER of the proposed MMVE scheme outperforms the LMS scheme, but underperforms the known channel which is the perfect system. It shows the performance of the proposed scheme with increased number of users compared to the LMS algorithm. There is significant degradation experienced as the number of users' increases from 4 to 8. Despite this degradation, the 8-users of the proposed scheme perform better compared to the 4-users in the LMS algorithm. Fig 3.5 shows the performance of the proposed scheme with increased number of users for fast fading. The performance of the proposed scheme for different number of users is compared to the performance of the known channel which serves as the bench mark. It is noted, that when the known channel is compared to the proposed scheme, the BER degrades as the number of users in the proposed scheme increases. Fig 3.6, illustrates the channel estimation mean square error (MSE) obtained for 4-users for the proposed scheme for fast fading and slow fading channels with mobile speed of 120km/h and 5km/h, respectively, compared with a fast fading channel at a mobile speed of 120km/h for LMS algorithm. Fig 3.7 illustrates the achievable MSE for all the different users that exist for the proposed scheme under fast fading condition.

Fig 3.8 demonstrates the equivalent BER performance for slow fading channel when mobile speed is 5km/h. The propose estimation scheme has different number of users and are compared to the known channel as a bench mark. The BER performance degrade as the number of users increase, but has a better BER performance when compared to fast fading of fig 3.4 and 3.5.

Fig. 3.9 shows the equivalent slow fading MSE versus SNR for the proposed scheme. The obtainable MSE versus SNR of the proposed scheme is illustrated in Fig 3.10. It combines the performance of the new scheme for fast fading and slow fading channel. The figure shows the difference between the MSE of fast fading and slow fading and it is observed that, when there is higher number of users in the system, the MSE difference between fast and slow fading is relatively small. Effectively, at a very high speed, with higher number of users, the proposed scheme is capable of estimating the channel accurately.

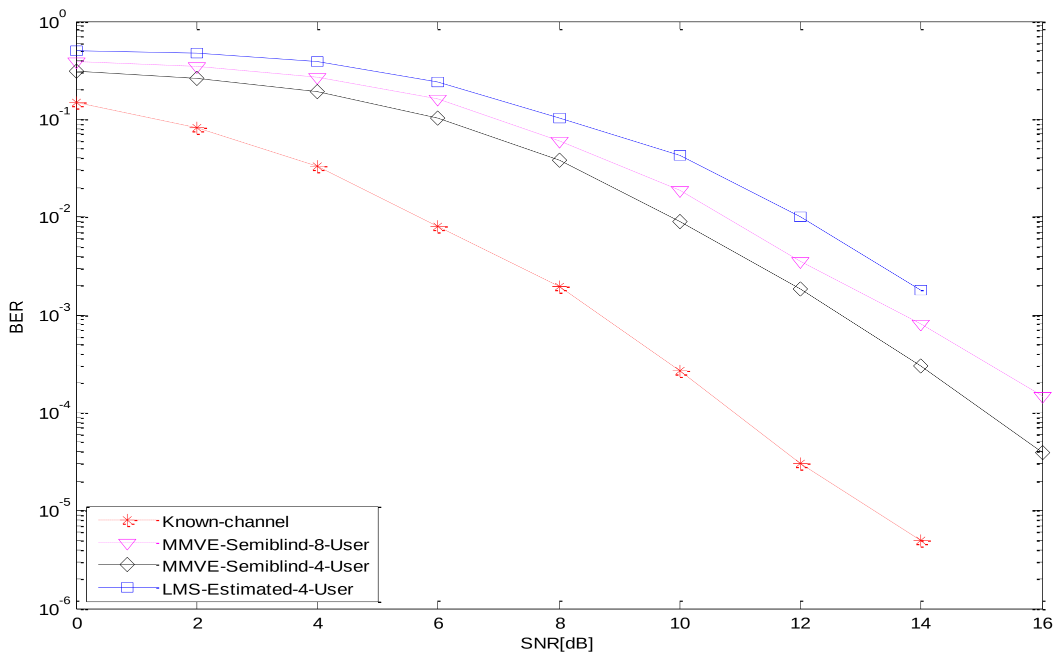


Figure 3.4: Performance of an MMVE semi-blind compared to LMS channel estimation algorithm in an OFDM- IDMA system over a fast fading channel.

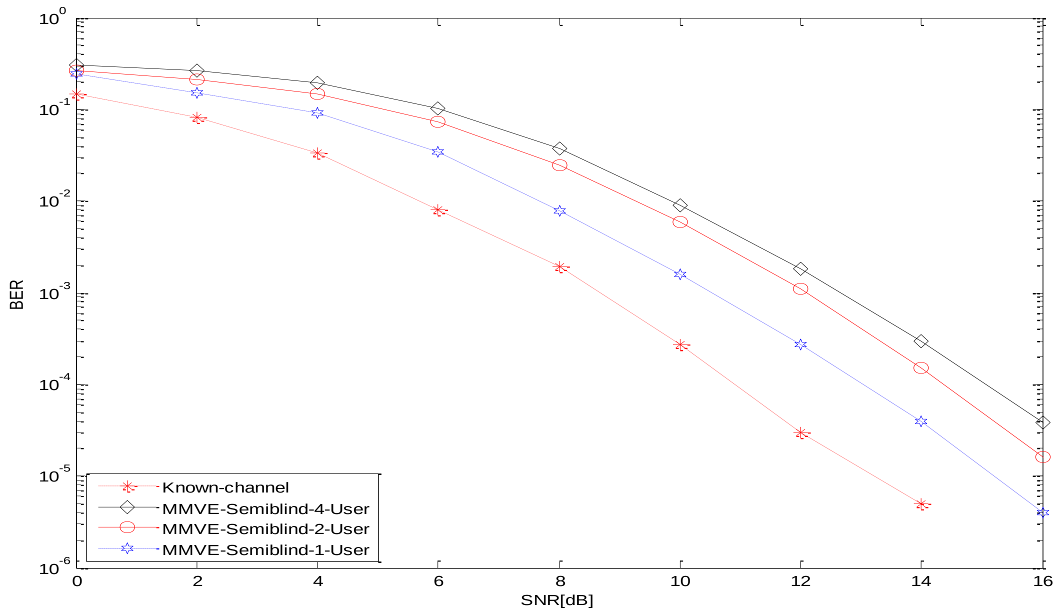


Figure 3.5: Performance of an MMVE semi-blind channel estimation algorithm in an OFDM- IDMA system over a fast fading channel; impact of number of users on BER Vs SNR

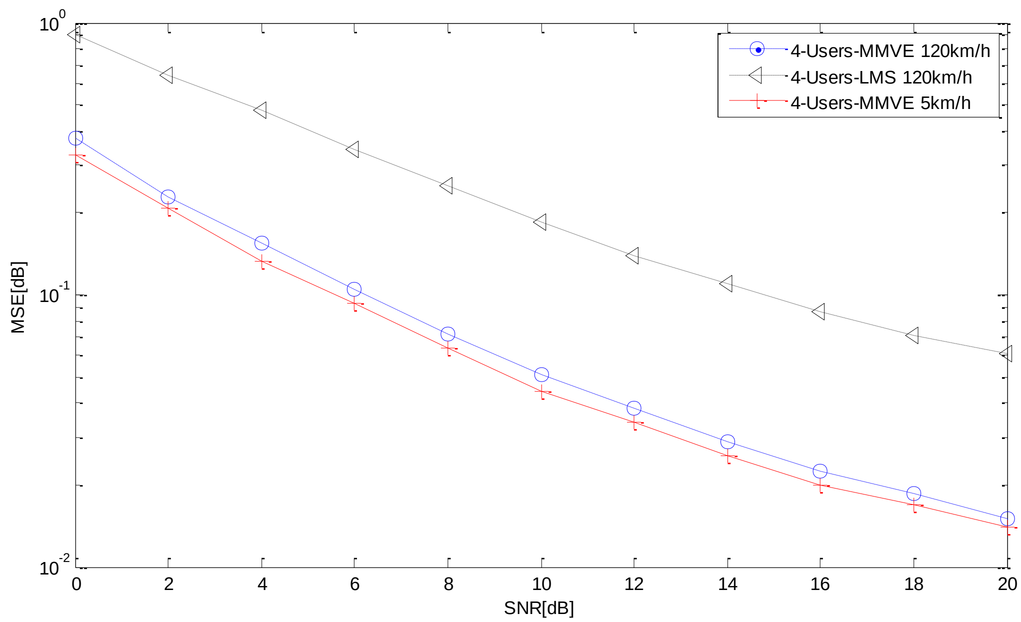


Figure 3.6: Performance of an MMVE semi-blind channel estimation algorithm compared to LMS in an OFDM- IDMA system over a fast fading channel and slow fading; impact of SNR on MSE

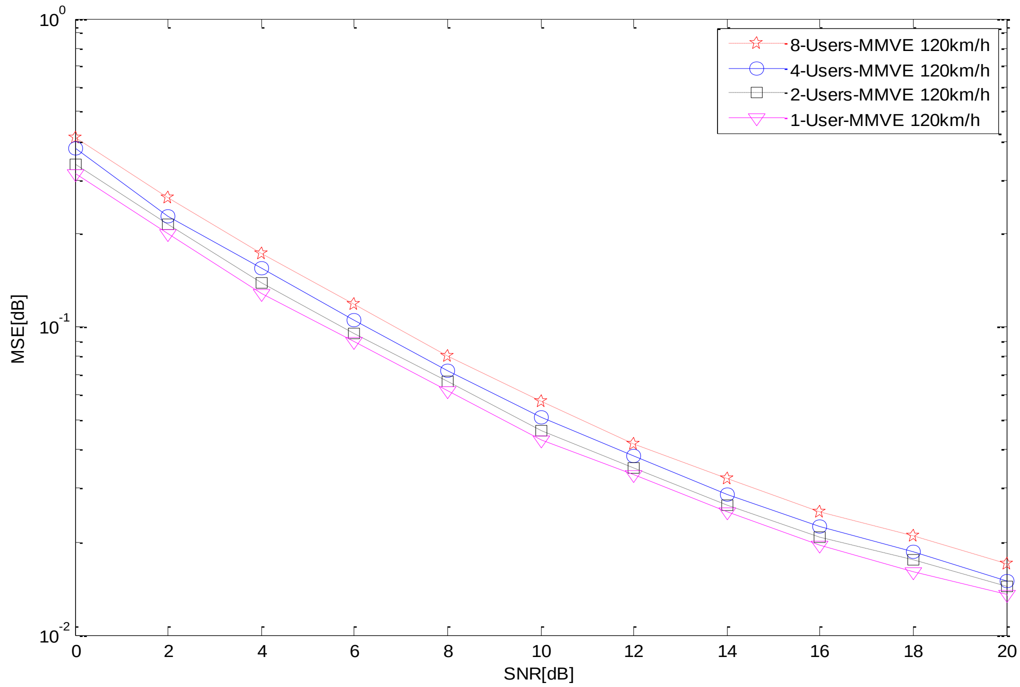


Figure 3.7: Performance of an MMVE semi-blind channel estimation algorithm over a fast fading channel; impact of the number of users on MSE Vs SNR.

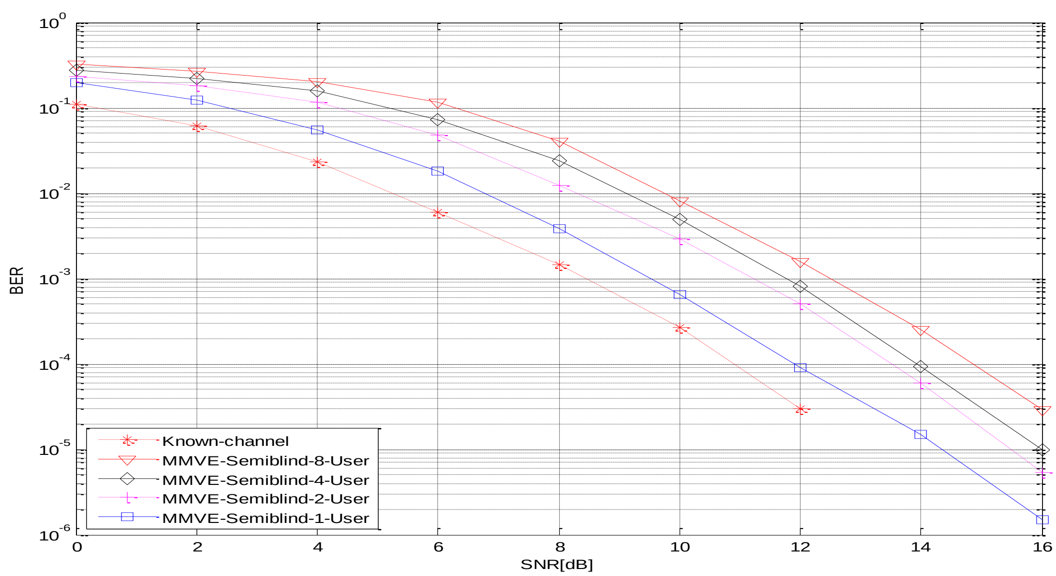


Figure 3.8: Performance of an MMVE semi-blind channel estimation algorithm in an OFDM- IDMA system over a slow fading channel; impact of number of users on BER Vs SNR

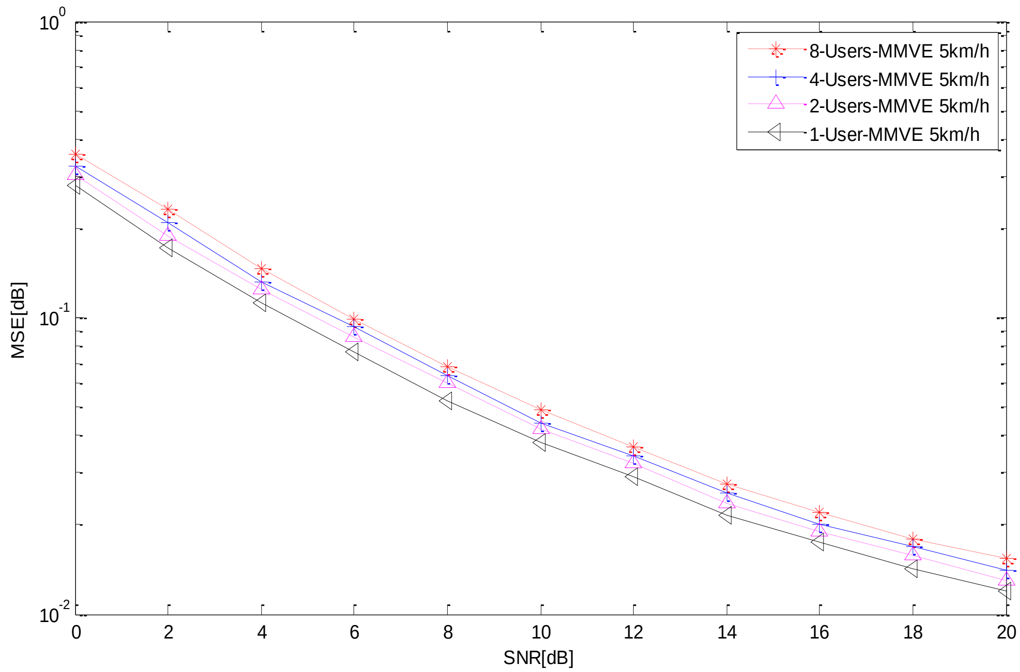


Figure 3.9: Performance of an MMVE semi-blind channel estimation algorithm over a slow fading channel; impact of the number of users on MSE Vs SNR.

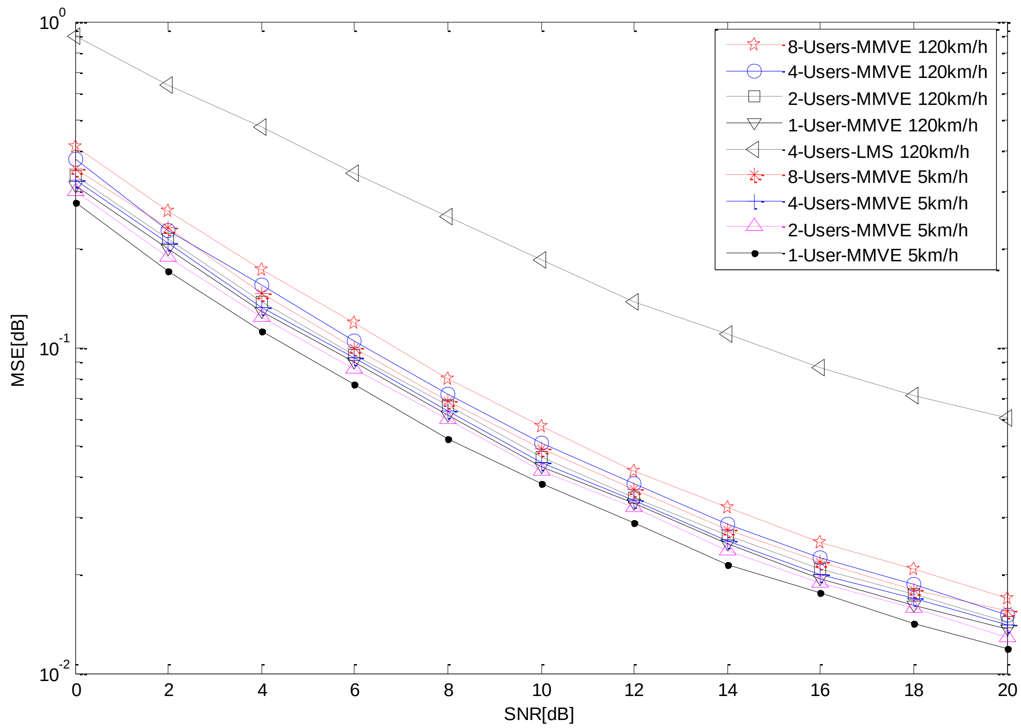


Figure 3.10: Performance of an MMVE semi-blind channel estimation algorithm over a fast and slow fading channel; impact of number of users on MSE Vs SNR.

3.7 Summary

The chapter analysed the channel model used in this thesis mathematically. MC-IDMA was introduced. The individual advantages of both OFDM and IDMA were combined together in the MC-IDMA scheme. The result gives an efficient resolution to ISI and MAI limitations of the multicarrier CDMA scheme. With sufficient cyclic prefix, ISI can be completely removed by OFDM, and the iterative CBC removes the MAI as well. It chapter concluded that synchronization and channel estimation are vital in an OFDM system. However, a perfectly synchronised system is assumed in this thesis. The assumption led to the introduction of channel estimation in an OFDM-IDMA system. The channel estimation adopting the principle in [22] to an OFDM-IDMA system. It can be deduce from simulated result that the algorithm performs well in an OFDM-IDMA system. These findings led to a reasonable conclusion that using the OFDM principle in MC-IDMA system, it is possible to transmit data symbols over multipath radio channels without influence from others in the multipath radio channel.

CHAPTER FOUR

THE IMPROVED MMSE ALGORITHMS

4.1 Introduction

In this section, an efficient estimation algorithm is presented and implemented. Two different improved version of MMSE algorithm are used. The first improved version of MMSE channel estimator is a modification to MMSE based on the assumption of a finite length impulse response and the second algorithm employs a structured correlation estimator on the OFDM-IDMA system. The advantage of the structured correlation estimator over the conventional MMSE is also analysed. To the best of our knowledge, this method has not been utilized for channel estimation in MC-IDMA systems, previously.

4.2 Modified Minimum Mean Value Estimator (MMMVE)

The MMSE estimator is modified using the assumption of a finite length impulse response by reducing the estimator matrix. The system use is given as

$$y(k) = h_u x_u + n_u \qquad u = 0\ .\ .\ .N-1, \tag{4.1}$$

where, $h_{(u)}$ denote the channel impulse response which is presented as $h = \left[h_{0,}\ h_1\ .\ .\ .h_{N-1}\right]^T = DFT_N(g)$ and $n = [n_0\ n_1\ .\ .\ .n_{N-1}]^T = DFT_N(\tilde{n})$ is the zero mean Gaussian noise vector. Equation (4.23) becomes

$$y = XFg + n, \tag{4.2}$$

where, g is the channel vector, and the matrix X has its element on its diagonal and

$$F = \begin{bmatrix} W_N^{00} & \cdots & W_N^{0(N-1)} \\ \vdots & \ddots & \vdots \\ W_N^{(N-1)0} & \cdots & W_N^{(N-1)(N-1)} \end{bmatrix}, \tag{4.3}$$

F = is a large DFT matrix for k= 0 to N-1. where $W_N = \frac{1}{\sqrt{N}} e^{-j2\pi nk/N}$.

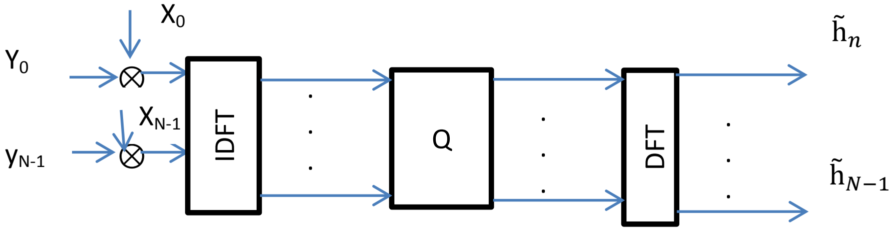


Figure 4.1: Estimator structure.

The transmitted symbol is denoted by X_0 in the Fig. 4.1, the IDFT and other OFDM operation take place. Q is a large matrix of twiddle factors and $\tilde{h}_n$is the estimated impulse response.

4.2.1 MMSE Estimator

The Gaussian channel vector g is uncorrelated with the channel noice n and its MMSE estimate of g is given by [91]

$$\tilde{g}_{MMSE} = R_{gy} R_{yy}^{-1} y \,, \tag{4.4}$$

where, g is the channel vector.

$$R_{gy} = E\{gy^H\} = R_{gg} F^H X^H \quad , \tag{4.5}$$

$$R_{yy} = E\{yy^H\} = XFR_{gg}F^H X^H + \sigma^2 I \quad ,$$
$$XX^H = I \;,$$

$$R_{yy} = R_{gg} + \sigma^2 I \,, \tag{4.6}$$

where, R_{gy} is the cross covariance between g and y, auto-covariance matrix of y and R_{gg} is R_{yy}, denoted the auto-covariance matrix of g.

$$\tilde{h}_{MMSE} = F\tilde{g}_{MMSE} = FQ_{MMSE} F^H X^H y \quad , \tag{4.7}$$

where,

$$Q_{MMSE} = R_{gg}\left[(F^H X^H XF)^{-1}\sigma^2 I + R_{gg}\right]^{-1}(F^H X^H XF)^{-1} \,,$$
$$Q_{MMSE} = R_{gg}\left[\sigma^2 I + R_{gg}\right]^{-1}. \tag{4.8}$$

Therefore equation 4.7 becomes:

$$\tilde{h}_{MMSE} = R_{gg}\left(\sigma^2 I + R_{gg}^{-1}\right) * X^H y \,. \tag{4.9}$$

Q_{MMSE} is obtained using a large N by N matrix. Reducing the matrix to L by L therefore simplifies the Q_{MMSE} to $\tilde{Q}_{MMSE}$.

4.2.2 Modified MMSE

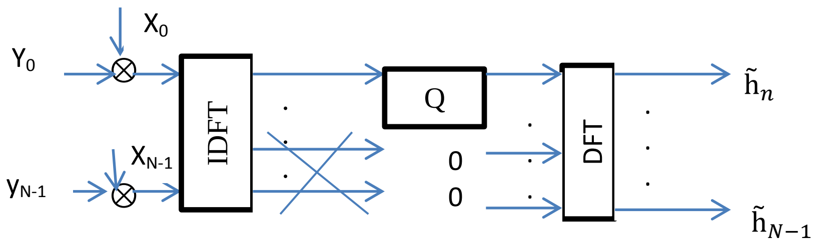


Figure 4.2: Modified estimator structure.

At the modification stage, consideration is given to the taps with significant energy and the rest taps with low energy are approximated to zero as shown in fig. 4.2. The new Q-matrix is of size L by L where,

$$L = \left[\frac{T_G}{T_S}\right] \ taps \ ,$$

with T_G being the cyclic extension of time length, and T_S sampling interval. T is the first L columns of the DFT matrix F (a small fraction of F), which is the part considered to be with the most significant energy.

$$T = \begin{bmatrix} \begin{bmatrix} W_L^{00} & W_L^{0(L-1)} \\ W_L^{(L-0)0} & W_L^{(L-1)(L-1)} \end{bmatrix} & \cdots & W_N^{0(N-1)} \\ \vdots & & \vdots \\ W_N^{(N-1)0} & \cdots & W_N^{(N-1)(N-1)} \end{bmatrix}$$

$$W_L^{IK} = \frac{1}{\sqrt{L}} e^{-j2\pi Ik/L} \quad .$$

The auto covariance matrix of g is given as R_{gg}, $\tilde{R}_{gg}$ denote the upper side L by L of R_{gg}. R_{gg} is an N by N matrix and $\tilde{R}_{gg}$ is derived from it by considering the first L taps of g, also setting R_{gg} (r,s) = 0 for [r,s] outside of [0, L-1]. This can be expressed as

$$\tilde{R} = \begin{bmatrix} \begin{bmatrix} \tilde{R}_{gg}^{11} & \tilde{R}_{gg}^{1L} \\ \tilde{R}_{gg}^{L1} & \tilde{R}_{gg}^{LL} \end{bmatrix} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & 0 \end{bmatrix} \quad .$$

Therefore equation (4.6) is modified as

$$\tilde{R}_{yy} = \tilde{R}_{gg} + \sigma^2 I \ . \tag{4.10}$$

$$\tilde{R}_{gg} = \tilde{R}_{yy} - \sigma^2 I \ . \tag{4.11}$$

$$\tilde{Q}_{MMSE} = \tilde{R}_{gg}\left[(T^H X^H XT)^{-1}\sigma^2 I + \tilde{R}_{gg}\right]^{-1}(T^H X^H XT)^{-1} \ .$$

$$\tilde{Q}_{MMSE} = \tilde{R}_{gg}\left[\sigma^2 I + \tilde{R}_{gg}\right]^{-1} \ . \tag{4.12}$$

$$h_{new} = T\tilde{Q}_{MMSE}T^H X^H y \ . \tag{4.13}$$

$$X^H y = h_{LS}$$

Therefore equation (4.13) become

$$h_{new} = \tilde{Q}_{MMSE}h_{LS} \ . \tag{4.14}$$

subtituting (4.12) into (4.14) gives

$$h_{new} = \tilde{R}_{gg}\left[\sigma^2 I + \tilde{R}_{gg}\right]^{-1}h_{LS} \ . \tag{4.15}$$

Substituting for $\tilde{R}_{gg}$ with equation (4.11) gives

$$h_{mmmve} = \left(\tilde{R}_{yy} - \sigma^2 I\right)\left[\sigma^2 I + \tilde{R}_{yy} - \sigma^2 I\right]^{-1} h_{LS}.$$

$$h_{mmmve} = \left(\tilde{R}_{yy} - \sigma^2 I\right)\tilde{R}_{yy}^{-1}h_{LS} \tag{4.16}$$

Semi-blind estimation

The reduced auto-covariance matrix of y is decomposed

$$\tilde{R}_{yy} = \tilde{U} \wedge \tilde{U}^H \tag{4.17}$$

$$h_{mmmve} = \tilde{U}(\wedge - \sigma^2 I)\tilde{U}^H\left(\tilde{U} \wedge \tilde{U}^H\right)^{-1}h_{LS} \tag{4.18}$$

$$h_{mmmve} = U\left(\frac{\wedge - \sigma^2 I}{\wedge}\right)U^H h_{LS}$$

$$h_{mmmve} = \tilde{U}diag\left(\frac{\wedge - \sigma^2 I}{\wedge}\right)\tilde{U}^H \mathrm{h_{LS}} \tag{4.19}$$

4.3 Simulations and discussion of MMMVE algorithm.

The computer simulations shows the performance of the modified MMSE algorithm using the finite length impulse response. The modified algorithm is termed Modified Minimum Mean Value Estimator (MMMVE). This section presents simulation results on performance of the OFDM-IDMA system based on BER and MSE against SNR when the MMMVE channel estimatior is used. The

OFDM-IDMA model employs QPSK modulation, a Rayleigh fading multipath channel with 16 paths (M = 16) and operating at 2 GHz carrier frequency. The MC-IDMA system model used input data 32 bits long. The spreading length for different instance of users is 8. The number of sub-carriers is 64, and for convenience, all the users in the simulations have equal number of sub-carriers. The simulation results presented (BER and MSE) are documented, operate with a normalised Doppler frequencies of fDn = 0.0045 and fDn = 0.1085 is used in a Rayleigh channel with mobile speeds of 5km/h, and 120km/h, respectively.

Fig 4.3 shows the performance of the modified algorithm (MMMVE) in comparison to the known channel as a bench mark, for a fast fading Rayleigh channel. Increase in the users' number in MMMVE leads to degradation of the system BER performance and as the SNR increases, the BER performance becomes better. Fig 4.4 gives the mean square error achievable by different users in MMMVE for fast fading channel at a mobile speed 120km/h.

Fig. 4.5 gives the performance of MC-IDMA system when the MMMVE estimator is used for slow fading channel with a mobile speed of 5km/h for different users in the system. Fig 4.6 illustrates the various users MSE versus SNR for the modified MMMVE algorithm at slow fading speed of 5km/h. Fig 4.7 shows the MSE for the combined fast fading and slow fading channel with relative speeds of 120km/h and 5km/h respectively.

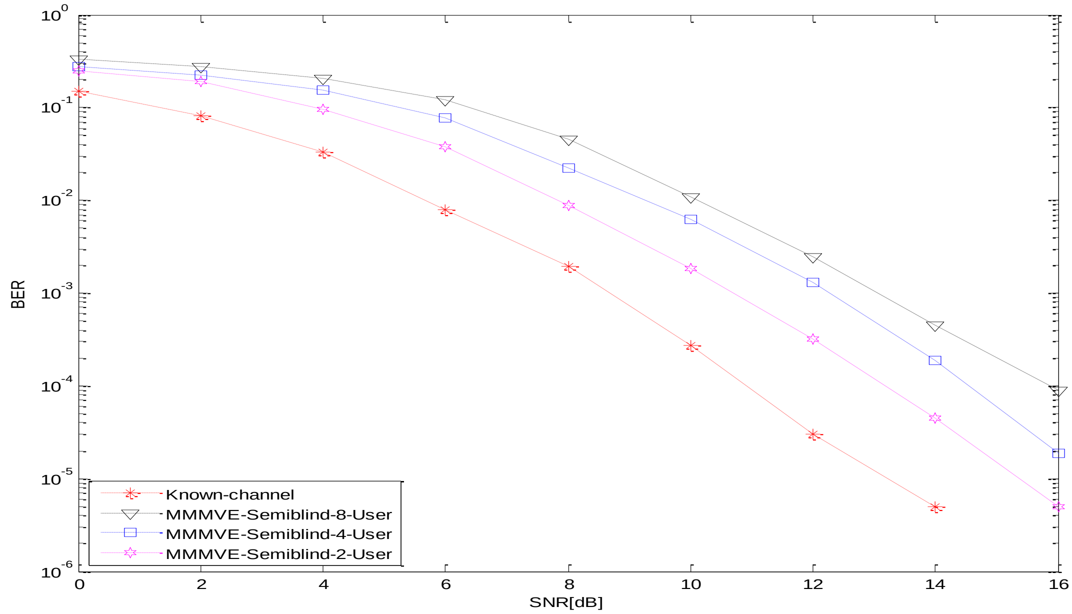


Figure 4.3: Performance of an MMMVE semi-blind channel estimation algorithm in an OFDM-IDMA over a fast fading channel; impact of number of users on BER Vs SNR.

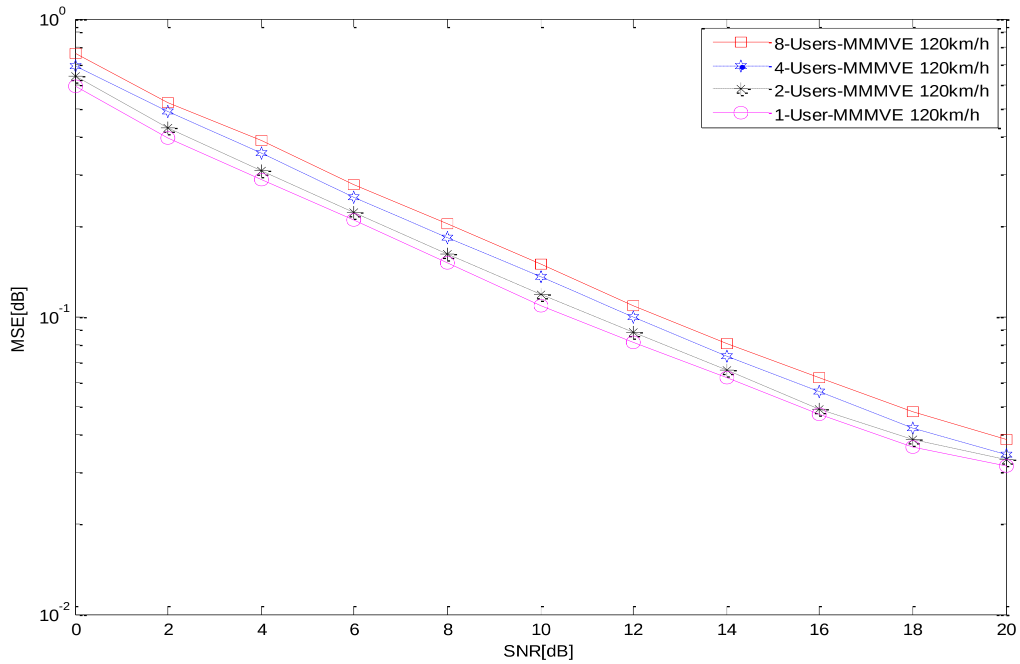


Figure 4.4: Performance of an MMMVE semi-blind channel estimation algorithm in an OFDM-IDMA over a fast fading channel; impact of number of users on MSE Vs SNR.

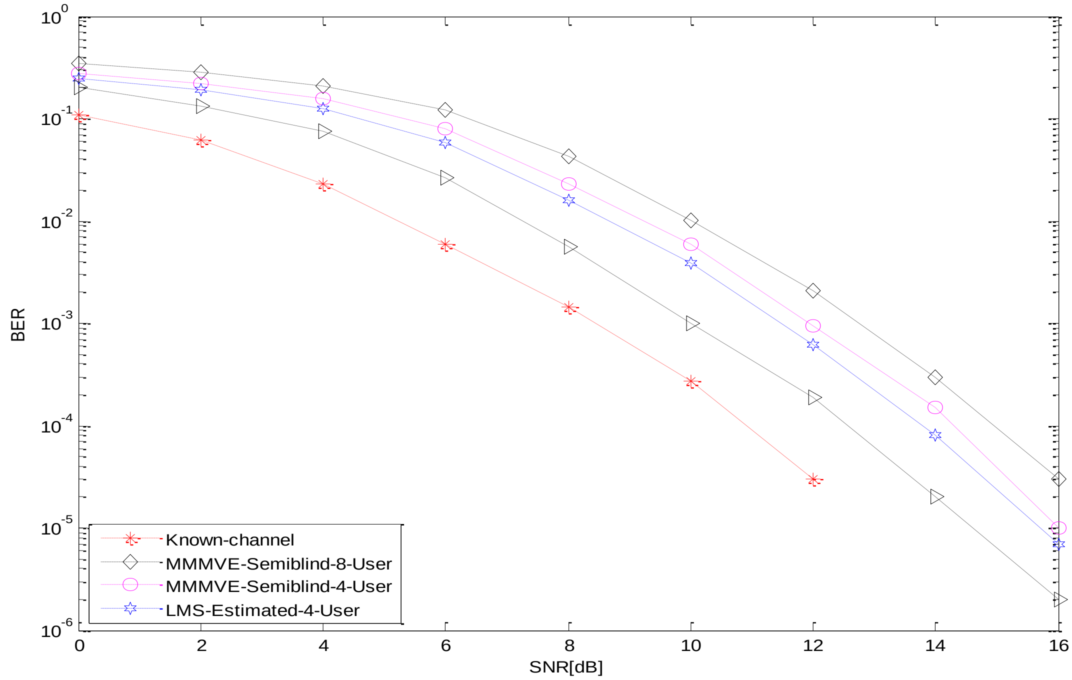


Figure 4.5: Performance of an MMMVE semi-blind channel estimation algorithm in an OFDM-IDMA over a slow fading channel; impact of number of users on BER Vs SNR.

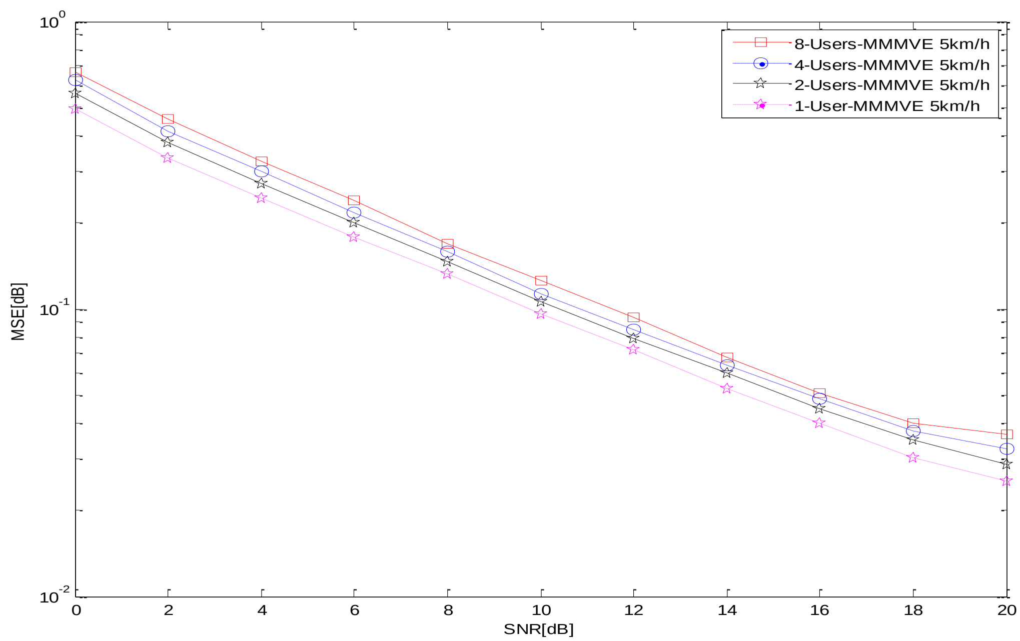


Figure 4.6: Performance of an MMMVE semi-blind channel estimation algorithm in an OFDM-IDMA over a slow fading channel; impact of number of users on MSE Vs SNR

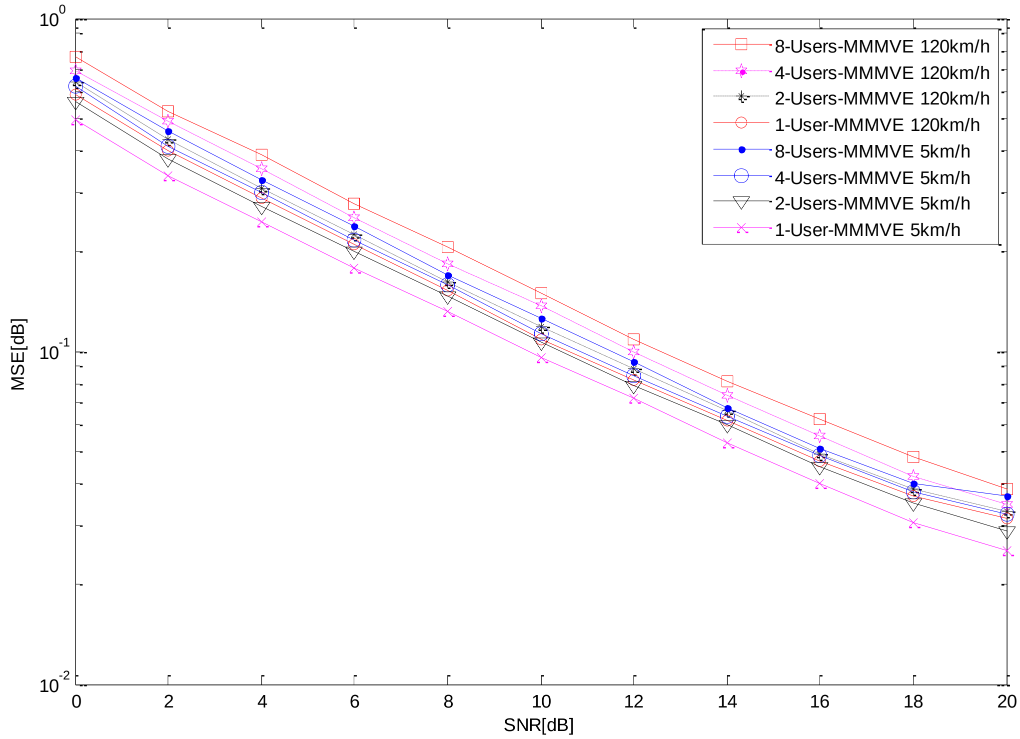


Figure 4.7: Combined MSE versus SNR for both fast and Slow fading; impact of number of users.

4.4 Simplified Mean Value Estimator (SMVE)

A semiblind channel estimation using an algorithm that combines MMSE and eigen decomposition is proposed. In this section, the received baseband signal is given as:

$$\mathrm{Y(k)} = SH(k) + W(k)\,, \tag{4.20}$$

where, the symbol S is an L by N matrix, the Gaussian channel is represented by H[k] which is N by N matrix, W(k) is an L by N Gaussian noise matrix and the received output is denoted by Y(k) is the L by N matrix. The symbol matrix S has a full rank so that $(S^H S)^{-1}$ exists. The correlation of channel matrix elements is given as $R_h = E\{h[k]h^H[k]\}$. The MMSE channel estimation is given by [90]

$$\tilde{\mathrm{h}}[k] = R_{yh}^H R_y^{-1} y[k] \quad , \tag{4.21}$$

Where, $R_{yh} = E\{y[k]h^H[k]\}$ and $R_y = E\{y[k]y^H[k]\}$,

$$R_{yh} = SR_h\,, \qquad and \tag{4.22}$$
$$R_y = SR_H S^H + \sigma^2 I\,. \tag{4.23}$$
Inserting (4.23) and (4.22) into (4.21) yields:
$$\tilde{h}[k] = R_h S^H (SR_h S^H + \sigma^2 I)^{-1} y \quad . \tag{4.24}$$
From 4.23
$$SR_h S^H = R_y - \sigma^2 I \quad \equiv \; R_h = \left(R_y - \sigma^2 I\right)(SS^H)^{-1}\,. \tag{4.25}$$
Inserting (4.25) into (4.3) gives
$$R_{yh} = SR_h = \left(R_y - \sigma^2 I\right)S(SS^H)^{-1}\,. \tag{4.26}$$
Substituting (4.26) in (4.21) yields:
$$\tilde{\mathrm{h}}[n] = R_{yh}^H R_y^{-1} y = \left(\left(R_y - \sigma^2 I\right)S(SS^H)^{-1}\right)^H R_y^{-1} \quad y\,.$$
Note: $S^{\#} = S(SS^H)^{-1}$

$$\tilde{\mathrm{h}}[n] \; = \left(\left(R_y^{-1}R_y - \sigma^2 I R_y^{-1}\right)S^{\#}\right)^H y \; = \left(S^{\#}\left(I - \sigma^2 R_y^{-1}\right)\right)^H \tag{4.27}$$

Based on structural correlation learning, the noise variance is given by [92]
$$\tilde{\sigma}^2 = \frac{1}{N-M}\; tr\left\{P^{\perp}\tilde{\mathrm{R}}_y P^{\perp}\right\}, \tag{4.28}$$
$$\tilde{\sigma}^2 = \frac{1-\lambda}{N-M}\sum_{i=1}^{k} \lambda^{k-i} \|P^{\perp} y(i)\|^2\,. \tag{4.29}$$

The channel correlation estimate $\tilde{\mathrm{R}}_h$ is
$$\tilde{\mathrm{R}}_h = S^{\#}\tilde{\mathrm{R}}_y S^{\#H} - \tilde{\sigma}^2 (S^H S)^{-1}\,, \tag{4.30}$$
$$\tilde{\mathrm{R}}_h = S^{\#}\left(\tilde{\mathrm{R}}_y - \tilde{\sigma}^2 I\right)S^{\#H} \quad . \tag{4.31}$$

The structured correlation estimate of $R_y$is given as [92]
$$\tilde{\mathrm{R}}_y^{(s)} = S\tilde{\mathrm{R}}_h S^H + \tilde{\sigma}^2 I = P\tilde{\mathrm{R}}_y P + \tilde{\sigma}^2 P^{\perp}\,. \tag{4.32}$$

Insert the noise variance estimate $\tilde{\sigma}^2$, structured correlation estimate $\tilde{\mathrm{R}}_y$ and the channel correlation $\tilde{\mathrm{R}}_h$ into equation (4.24)
$$\tilde{\mathrm{h}} = R_h S^H (SR_h S^H + \sigma^2 I)^{-1} y = \tilde{\mathrm{R}}_h S^H \left(S \; \tilde{\mathrm{R}}_h S^H + \tilde{\sigma}^2 I\right)^{-1} y. \tag{4.14}$$
Note that: $\tilde{\mathrm{h}}_{LS} = S^{\#}y$= preliminary channel estimation based on Least Square Algorithm.
(4.33) can be written as:
$$\tilde{\mathrm{h}} = \tilde{\mathrm{R}}_h S^H \left(S\tilde{\mathrm{R}}_h S^H + \tilde{\sigma}^2 I\right)^{-1} \tilde{h}_{Ls}\;,$$
$$\tilde{\mathrm{h}} = \tilde{\mathrm{R}}_h S^H (SS^H)^{-1}\left(\tilde{\mathrm{R}}_h + \tilde{\sigma}^2 (SS^H)^{-1}\right)^{-1} \frac{\tilde{\mathrm{h}}_{Ls}}{S^{\#}}\,,$$
$$\tilde{\mathrm{h}} = \tilde{\mathrm{R}}_h S^{\#}\left(\tilde{\mathrm{R}}_h + \tilde{\sigma}^2 (SS^H)^{-1}\right)^{-1} \frac{\tilde{\mathrm{h}}_{Ls}}{S^{\#}}\;,$$

$$\tilde{h} = \tilde{R}_h\left(\tilde{R}_h + \tilde{\sigma}^2(SS^H)^{-1}\right)^{-1}\tilde{h}_{LS},$$

$$\tilde{h} = \tilde{R}_h\left(\tilde{R}_h + \tilde{\sigma}^2 I\right)^{-1}\tilde{h}_{LS}. \tag{4.15}$$

From equation (4.32):

$$\tilde{R}_h = S^{\#}\left(\tilde{R}_y - \tilde{\sigma}^2 I\right)S^{\#H} \quad = \left(\tilde{R}_y - \tilde{\sigma}^2 I\right)S^{\#}S^{\#H},$$

$$\tilde{R}_h = \tilde{R}_y - \tilde{\sigma}^2 I \quad ,$$

$$\tilde{R}_h = \tilde{R}_y - \tilde{R}_w , \tag{4.16}$$

where,

$$\tilde{R}_w = \tilde{\sigma}^2 I,$$

from equation (4.32)

$$\tilde{R}_y^{(s)} = S\tilde{R}_h S^H + \tilde{\sigma}^2 I\,. \tag{4.17}$$

$\tilde{R}_h$is obtained from the preliminary channel estimate

$$\tilde{h}_{prel} = \tilde{H}_{LS}$$

$$\tilde{R}_h = \frac{1}{N}\sum_{n=1}^{N}\tilde{h}_{prel}\tilde{h}_{prel}^H = \frac{1}{N}\sum_{n=1}^{N}h_{LS}h_{LS}^H = E\{h_{LS}h_{LS}^H\}\,.$$

Making use of the approximation that $\tilde{R}_h = h_{LS}h_{LS}^H = \tilde{h}\tilde{h}^1$
equation (4.36) becomes:

$$\tilde{R}_y^{(s)} = \tilde{R}_h(SS^H) + \tilde{\sigma}^2 I = \tilde{h}\tilde{h}^H(I) + \tilde{\sigma}^2 I = \tilde{h}\tilde{h}^H + \tilde{\sigma}^2 I\,. \tag{4.37}$$

From the decomposition of $\tilde{R}_y^{(s)}$, we obtain the eigen value and vector

$$\tilde{R}_y^{(s)} = \tilde{U} \wedge \tilde{U}^H = \tilde{R}_h - \tilde{\sigma}^2 I = \tilde{R}_h + \tilde{\sigma}^2\, \tilde{U}\tilde{U}^H \;, \tag{4.38}$$

$$\tilde{H}_{new} = \left(\tilde{R}_y^{(s)} - \tilde{\sigma}^2 I\right)\left[\left(\tilde{\sigma}^2 I + \tilde{R}_y^{(s)} - \tilde{\sigma}^2 I\right)\right]^{-1}\tilde{h}_{,LS},$$

$$\tilde{H}_{new} = \left(\tilde{R}_y^{(s)} - \tilde{\sigma}^2 I\right)\left(\tilde{R}_y^{(s)}\right)^{-1}\tilde{h}_{,LS} \;,$$

$$\tilde{H}_{new} = \left(\tilde{R}_y^{(s)} - \tilde{\sigma}^2 I\right)\left[\left(\tilde{R}_y^{(s)}\right)^{-1}\right]\tilde{h}_{,LS}\;. \tag{4.39}$$

Recall equation (4.38) $\tilde{R}_y^{(s)} = \tilde{R}_h + \tilde{\sigma}^2\, \tilde{U}\tilde{U}^H$

Therefore,

$$\tilde{H}_{SMVE} = \tilde{U}\left(\wedge - \tilde{\sigma}^2 I\right)\tilde{U}^H\left(\tilde{U} \wedge \tilde{U}^H\right)^{-1}\tilde{h}_{,LS}\;, \tag{4.40}$$

$$\tilde{H}_{SMVE} = \tilde{U}\left(\frac{\wedge - \tilde{\sigma}^2 I}{\wedge}\right)\tilde{U}^H\tilde{h}_{,LS}\;,$$

$$\tilde{\mathrm{H}}_{SMVE} = \tilde{\mathrm{U}} diag\left(\frac{\wedge\ - \tilde{\sigma}^2 I}{\wedge}\right)\tilde{\mathrm{U}}^H \tilde{\mathrm{h}}_{,\mathrm{LS}}\ . \qquad (4.41)$$

4.5 Simulations and discussion of the SMVE estimation algorithm.

The system model used for the the simplified MMSE algorithm remain the same as the previous system. The computer simulations show the performance of the simplified MMSE algorithm using the structured correlation. The simplified algorithm is termed the Simplified Mean Value (SMVE) Estimator. The MC-IDMA system model used 32 input data length, 8 bit spreading length for different instances of users. The number of sub-carriers is 64, and for the convenience, all the users in the simulations are assumed to have equal number of sub-carriers. The channel was subject to 16-paths i.e M = 16. The normalized Doppler frequencies of the Rayleigh fading channel is fDn = 0.0045, and fDn = 0.1085 with mobile speeds equivalent of 5km/h, and 120km/h, respectively. The performance of the system is measured based on BER and MSE against SNR of the computer simulations.

Fig 4.8 shows the performance of the simplified algorithm SMVE on an OFDM-IDMA system in comparison to the known channel bench mark for a fast fading speed of 120km/h. BER performance of the system degrades as the number of users' increases, and BER becomes better as the SNR increases. Fig 4.9 illustrates the mean square error of the channel estimation achievable by different users using SMVE in a fast fading channel for a mobile speed 120km/h.

Fig. 4.10 gives the performance of an OFDM-IDMA system using the SMVE algorithm in a slow fading channel for a mobile speed of 5km/h. Fig 4.11 shows the impact of MSE versus SNR on the number of users in SMVE at slow fading speed of 5km/h. Fig 4.12 shows the MSE for the combined fast fading and slow fading channel with relative speeds of 120km/h and 5km/h respectively. The difference between the fast fading and slow fading MSE is observed here. Fig4.12 shows that there is relatively small difference between the fast and slow fading MSE.

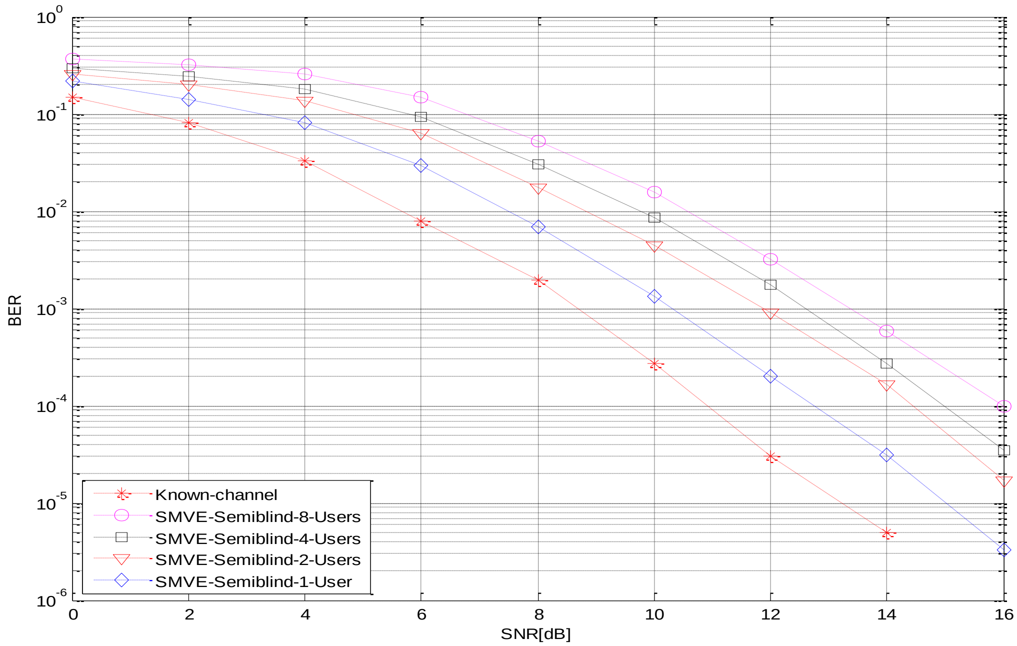


Figure 4.8:Performance of SMVE semi-blind channel estimation algorithm in an OFDM-IDMA system over a fast fading channel; impact of number of users on BER Vs SNR.

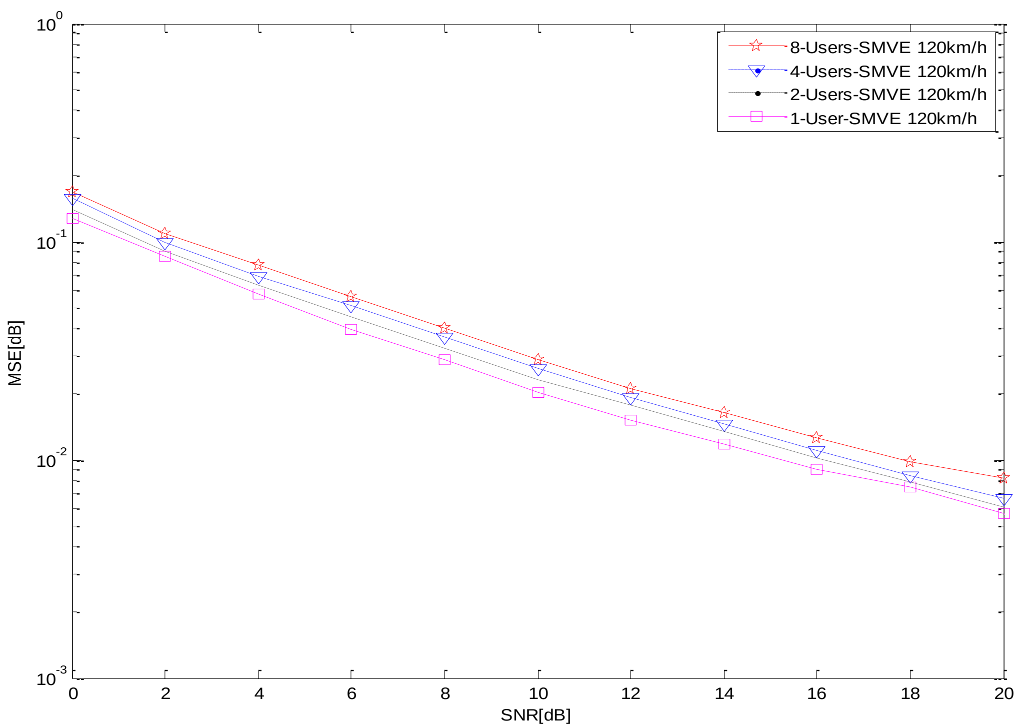


Figure 4.9: Performance of SMVE semi-blind channel estimation algorithm in an OFDM-IDMA system over a fast fading channel; impact of number of users on MSE Vs SNR.

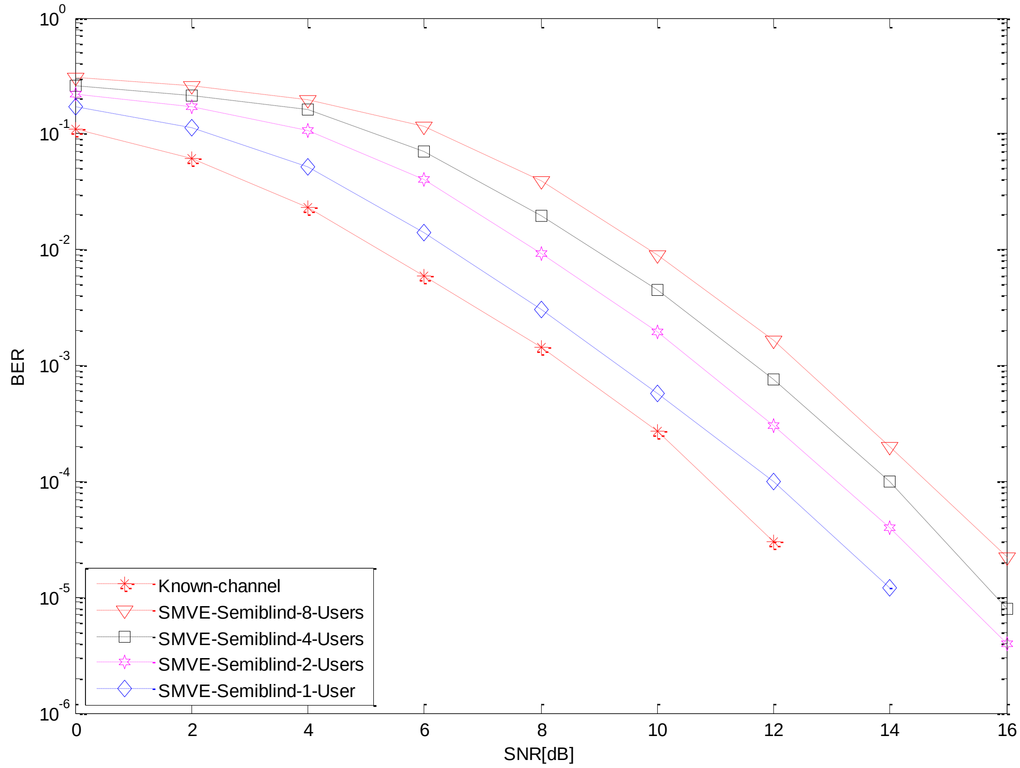


Figure 4.10: Performance of SMVE semi-blind channel estimation algorithm in an OFDM-IDMA system over a slow fading channel; impact of number of users on BER Vs SNR.

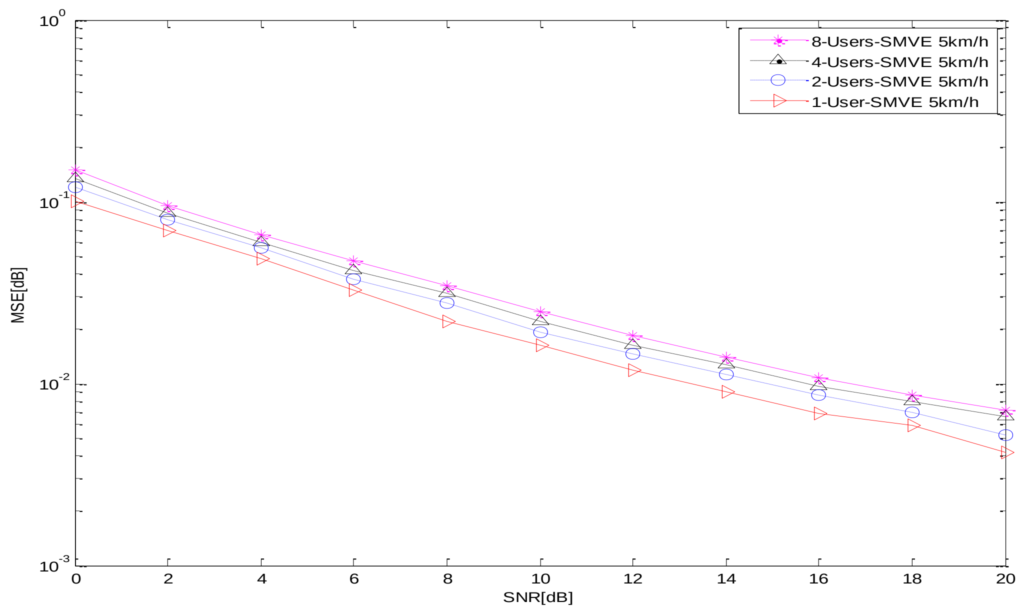


Figure 4.11: Performance of SMVE semi-blind channel estimation algorithm in an OFDM-IDMA system over a slow fading channel; impact of number of users on MSE Vs SNR.

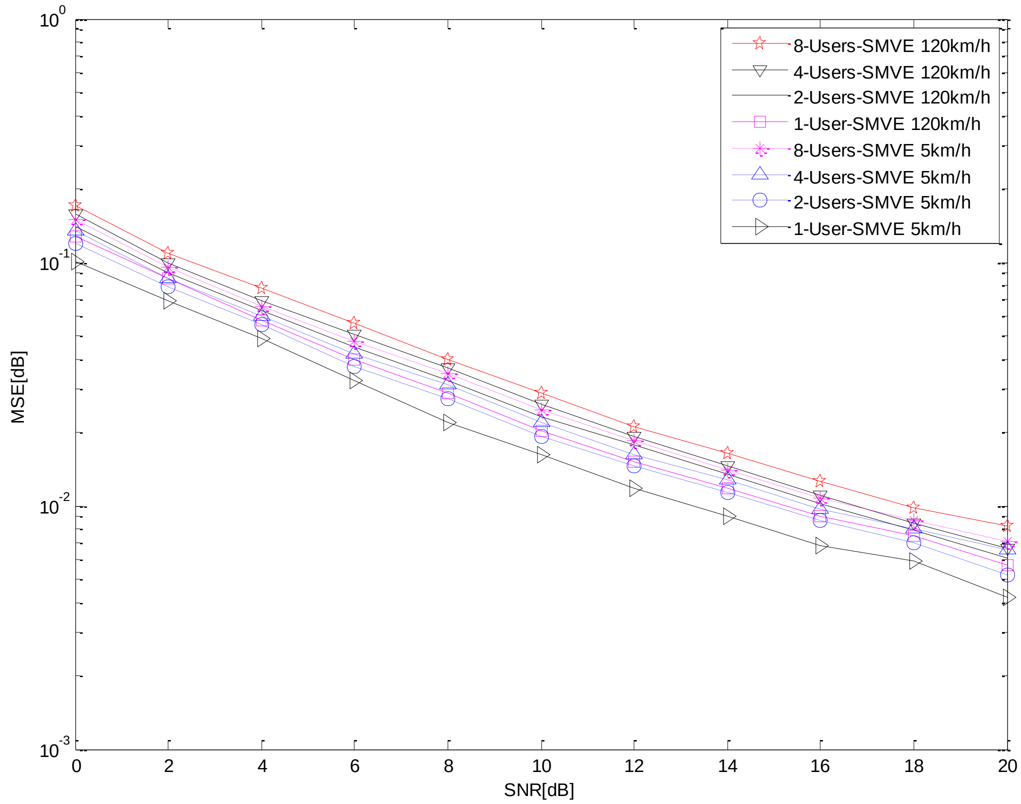


Figure 4.12: Combined MSE versus SNR for both fast and Slow fading.

4.6 Chapter Summary

Two different versions of improved MMSE algorithms are presented. The algorithms were improved using the eigen value decomposition for semi-blind channel estimation and adopted to OFDM-IDMA system. The chapter shows the performance of these algorithms using the MSE and BER versus SNR with respect to increase in number of users. The difference between the fast fading and slow fading MSE is observed in Fig. 12, which shows that there is relatively small difference between the fast and slow fading MSE. It can therefore be concluded that the proposed scheme is capable of estimating the channel effectively at a very high speed with higher number of users.

CHAPTER 5

COMPARISON OF THE ESTIMATION METHODS

In this section, we present the comparison of the simulation results for the 3 different estimation algorithm proposed for OFDM-IDMA systems. The Mean Square Error (MSE) of each algorithm was investigated to determine the best performance algorithm. The computer simulations show the performances of the MMVE, the modified MMSE algorithm using the finite length impulse response and the SMVE using the structured correlation. The MMSE algorithm is referred to as Minimum Mean Value Estimator (MMVE), the modified algorithm is referred to as Modified Minimum Mean Value Estimator (MMMVE) and the last which makes use of structured correlation is referred to as Simplified Mean Value Estimator (SMVE). This section present the performance of the OFDM-IDMA system model in terms of its MSE versus SNR performances for different number of users. The system model remains as OFDM-IDMA model employing QPSK modulation technique and in a Rayleigh fading multipath channel for 16 paths and operates at carrier frequency 2 GHz. The MC-IDMA system model used 32 input data bits, spreading code of length 8 for different users. The number of sub-carriers is 64, and for the convenience, all the users in the simulations are assumed to have equal number of sub-carriers. The simulation results presented (BER and MSE) are documented for a Rayleigh channel with normalised frequencies of fDn = 0.0045, fDn = 0.1085 and the corresponding mobile speeds of 5km/h and 120km/h, respectively.

5.1 Simulations and discussion

Fig 5.1 shows the comparison of the MMSE algorithm, the MMVE of fig 3.8 and the modified algorithm MMMVE of fig 4.2. It can be seen from MSE versus SNR in Fig 5.1 that the performance of the MMVE is better than that of the MMMVE under a fast fading channel for a mobile speed of 120km/h mobile speed; Fig. 5.2 displays the comparison between MMVE and MMMVE for slow fading condition at a mobile speed of 5km/h. This is equivalent to comparing the result of Fig. 3.10 and Fig. 4.4. The performance degradation of the modified MMSE i.e the modified MMMVE is due to the fact that some of the channel statistics are ignored.

Also, in Fig 5.3, the performance of MMVE is compared with SMVE algorithm. The MMVE is chosen, as we have shown that it outperforms the MMMVE algorithm. The SMVE algorithm also outperforms the MMVE algorithm.

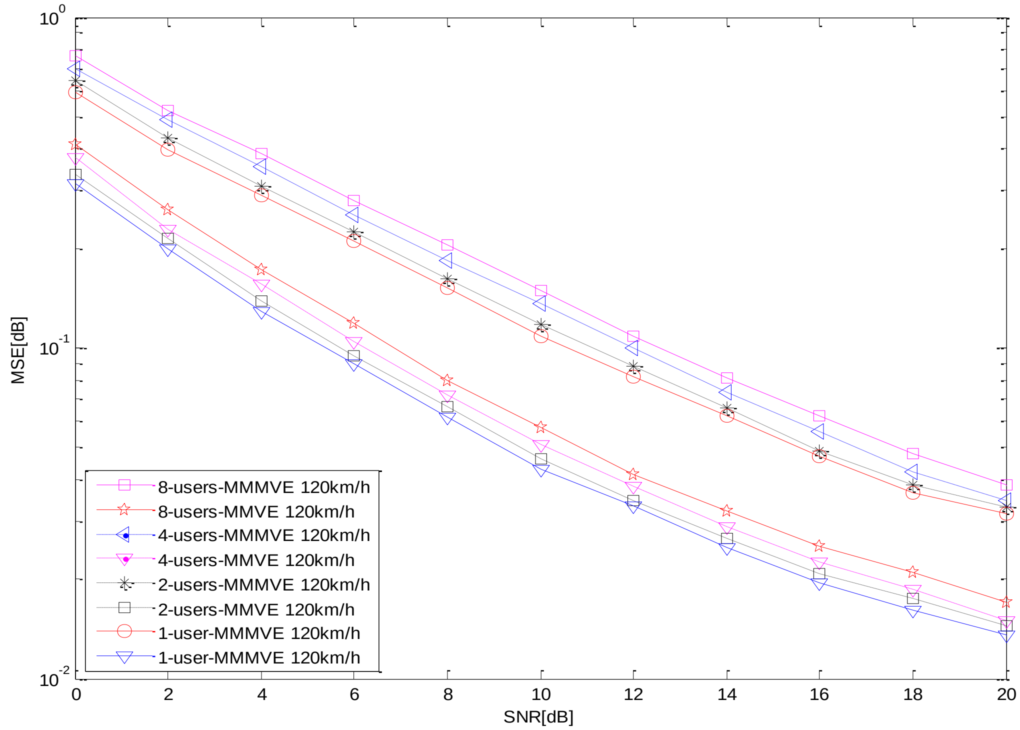


Figure 5.1: MSE comparison for MMVE and MMMVE for a fast fading channel.

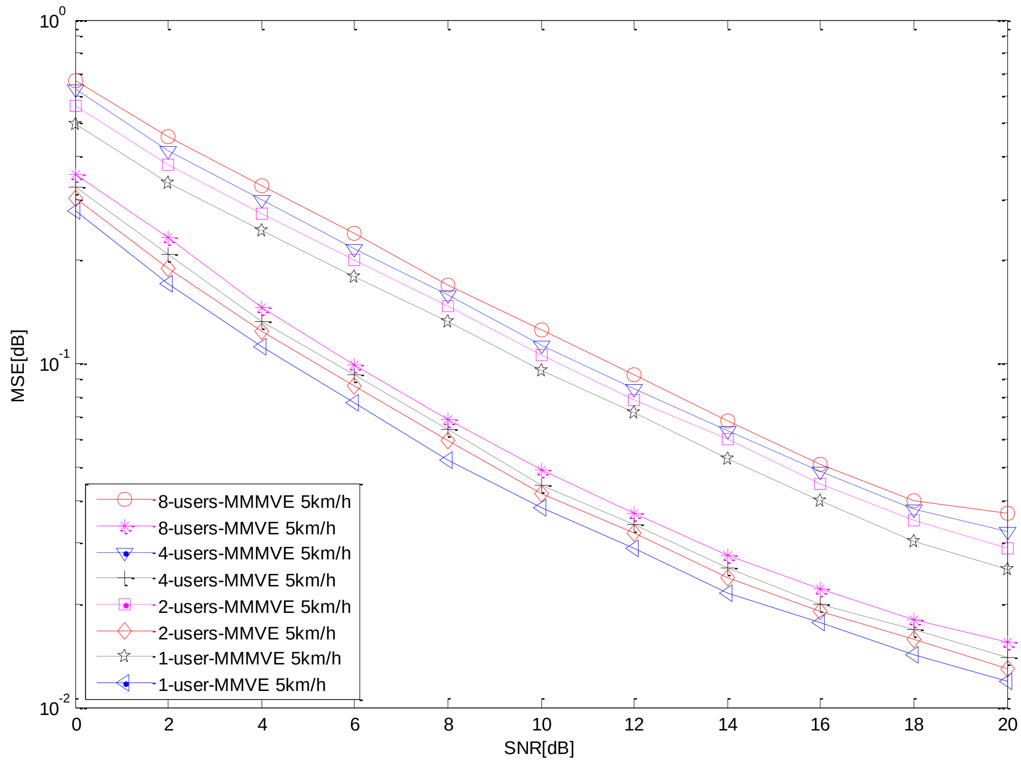


Figure 5.2: MSE comparison for MMVE and MMMVE for a slow fading channel.

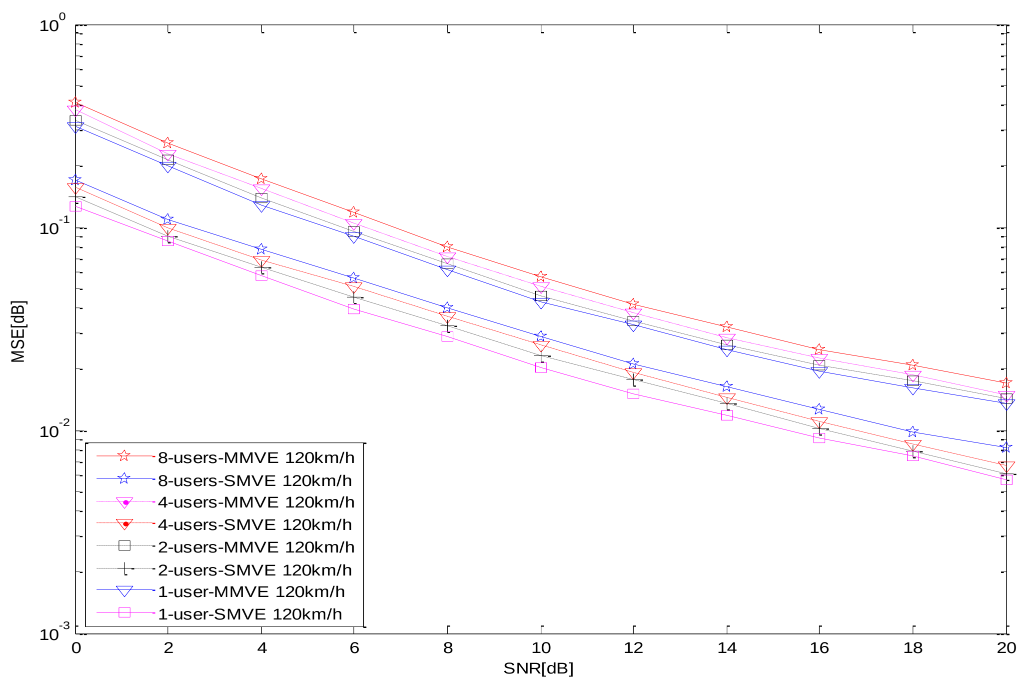


Figure 5.3: MSE comparison for MMVE and SMVE for a fast fading channel.

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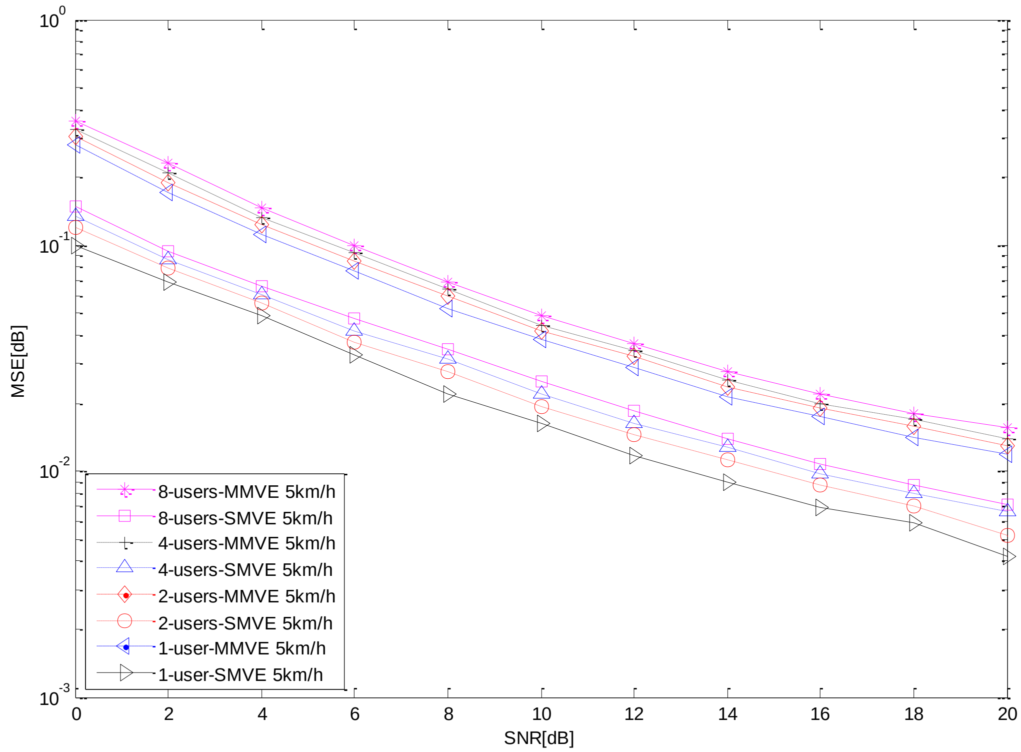

Figure 5.4: MSE comparison for MMVE and SMVE for a slow fading channel.

5.2 Computational complexity

The three algorithms used in this thesis required M^2 operation for the Least Square (LS) based initialization estimator. After the initialisation by LS, The H_{MMVE} based estimator will require $[M^2 + M(2K +1)]$ multiplication/division operations and $2M(K+1)$ addition/subtraction operations. The modified algorithm H_{MMMVE} requires $[M^2+M(K +1)]$ multiplication/division and $M(K+2)$ addition/subtraction operations. The last algorithm H_{SMVE} require $M^2 + M(3K+1)$ multiplication/division and $M(3k+2)$ addition/subtraction operations. These complexities are compared with the Least Mean Square because the LMS was compared with the first algorithm in chapter 3. LMS complexity after LS initialisation is $(M^2 + 3MF + M)$ multiplication/division and $M(F+2)$ addition / subtraction operations. For $K = 64$, $M = 16$, and $F = 10$, where K is the number of OFDM subcarriers, M is the length of the channel and F is the adaptive filter length for the LMS.

Estimator	$\times / \div$	$+/-$
LS based initialization estimator	M^2	----------
LMS-based estimator	$(M^2 + 3MF + M)$	M(F+2)
H_{MMVE} based estimator	$M^2 + M(2K + 1)$	2M(K+1)
H_{MMMVE} based estimator	$M^2 + M(K + 1)$	M(K+2)
H_{SMVE} based estimator	$M^2 + M(3K + 1)$	M (3k +2)

Table: 5-1: Computational complexity analysis.

These parameters remain the same for all the algorithms. The comprehensive computational complexity for the first algorithm H_{MMVE} is 4400, also the overall computational complexity for H_{MMMVE} is 2352 and the last H_{SMVE} has a computation complexity of 6464. Likewise, the overall computational complexity of LMS-based estimator is 944. From these analyses, LMS-based estimator has the lowest computational complexity while the third algorithm H_{SMVE} based estimator exhibits the highest computational complexity.

5.3 Summary

From the MSE versus SNR comparison, the MMSE algorithm MMVE performs better than the modified version but has a higher complexity because it makes use of a large Q martrix. Modified algorithm MMMVE makes use of taps with significant energy, though the complexity is reduced but with reduced performance because some parts of the channel statistics are not taken to consideration. The best performance is due to the simplified algorithm SMVE. It performs better than the MMSE and its modified version.

CHAPTER 6

CONCLUSION AND FUTURE WORK

6.1 Conclusion

In mobile communication systems, there is a big increase in the demand for spectrum usage. An efficient usage of the scarce spectrum is obtained using Multicarrier IDMA systems. This thesis is devoted to the investigations of channel estimation in an OFDM-IDMA scheme on an uplink multipath fading channel environment.

In this thesis a multicarrier multiple access communication technique namely, MC-IDMA is developed. The advantages exhibited by this system (MC-IDMA) are inherited from other multiple access techniques with additional benefits. Interleavers are the only means of users separation at the receiver and the entire bandwidth expansion is devoted to forward error correcting codes. The system used is a combination of OFDM and IDMA. The system used is designed in a way that the CP length of the OFDM component used is longer compared to the length of the delay spread. Thus Inter-symbol Interference (ISI) is removed and its iterative IDMA with its CBC detection algorithm overcomes multiple access interference (MAI) efficiently in the system.

Comprehensive work is carried out on MC-IDMA system. The system is assumed to be perfectly synchronised. However, due to the frequency selective and time-varying nature of the radio channel, channel estimation be done. An efficient channel estimation algorithm is proposed. The channel estimation using a linear minimum mean square error (MMSE) based algorithm is used for the estimation of the system. The performance and complexity of this algorithm on MC-IDMA is investigated in a Rayleigh fading multipath channel.

The bit error rate is evaluated to predict the performance of channel estimation of MC-IDMA system using the MMSE algorithm. The performance is also evaluated in terms of the number of users in the system. Simulation results in terms of BER and MSE show that the MMSE based algorithm performs better irrespective of the number of users in the system compared to LMS estimation.

Also, an improved version of the MMSE algorithm is used on this MC-IDMA system. The performance of the improved MMSE algorithm, the MMMVE, where modification of the finite length impulse response is used to modify the MMSE algorithm is degraded but it has small complexity.

Likewise, in the third algorithm SMVE, the structured correlation is used to modify and improve the MMSE algorithm. Simulation results show that the proposed algorithm outperforms the linear MMSE and second improved MMSE algorithm i.e the MMMVE has reduced channel estimation errors and estimation complexity.

This algorithm SMVE excels in performance compared to the other implemented algorithms, as seen from the simulation results. The implemented algorithms were compared for increasing number of users in the system. The BER performance of the implemented algorithms is presented concurrently. The application of the algorithms resulted in significant improvement of the overall output of the multicarrier system. The algorithms also achieve great performances even in a fast fading multipath scenario. However, it is noteworthy that the finite length impulse response modified algorithm offers a better, effective, and more efficient system performance than both its structured correlation based MMSE algorithm and the linear MMSE-based algorithm.

6.2 Future work

In this thesis, all simulations are carried out with the assumption of perfect time and frequency synchronization. This assumption is not true in practical scenarios. It would be of great advantage to integrate the channel estimation algorithm with Channel Frequency Offset (CFO) compensation algorithm. Hence, there is need to develop an algorithm that combines the estimation and CFO correction at the same time.

There is also a need to investigate an effective and efficient way to minimise peak-to-average power ratio problem that exist in practical scenario.

Adoption of these algorithms to MIMO techniques is an area that can be investigated.

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