ADDIS ABABA UNIVERSITY
ADDIS ABABA INSTITUTE OF TECHNOLOGY (AAiT)
ELECTRICAL AND COMPUTER ENDINEERING DEPARTMENT

# TIME DOMAIN EQUALIZATION FOR OFDM SYSTEMS IN TIME VARYING CHANNELS

**BY**
**AWOKE ATENA**

*a thesis submitted to Addis Ababa Institute of Technology in partial fulfillment of the requirement  for the degree of Master of Science in Electrical Engineering*

ADVISOR
Dr.-Ing. Hailu Ayele
July 2011
ADDIS ABABA, ETHIOPIA

Declaration

I, the undersigned, declare that this thesis work is my original work, has not been accepted for a degree and is not concurrently submitted in candidature of any other degree in this or any other universities, and all sources of materials used for the thesis work have been fully acknowledged.

Awoke Atena ___________

Name Signature

Addis Ababa, Ethiopia June 2011

Place Date

This thesis has been submitted for examination with my approval as the university advisor.

Dr.-Ing. Hailu Ayele ___________

Advisor's Name Signature

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# ABSTRACT

Orthogonal Frequency Division Multiplexing (OFDM) is an emerging multi-carrier modulation scheme, which has been adopted for several wireless standards such as IEEE 802.11a and HiperLAN2. OFDM as a transmission technique has been known to have a lot of strengths compared to any other transmission technique due to its high spectral efficiency, robustness to the channel fading, immunity to impulse interference and ability to handle very strong echoes. The efficacy of OFDM implementation in many areas such as DAB (Digital audio Broadcasting), DVB (Digital Video Broadcasting) and Wireless LAN has gained its popularity.

A well-known problem of OFDM is its sensitivity to frequency offset between the transmitted and received carrier frequencies. This frequency offset introduces inter-carrier interference (ICI) in the OFDM symbol. This thesis investigates time domain equalization technique by using well designed windows for combating the effects of ICI entitled equalization with time domain windowing. This method is compared with other methods like existing frequency domain correlative coding and self cancellation methods in terms of bit error rate performance, carrier to interference ratio and bandwidth efficiency. Effects of different orders of windowing on the carrier to interference power ratio are investigated for various normalized frequency offset values. Through simulations, it is shown that the Time domain windowing technique is effective in mitigating the effects of ICI. It shows a better performance in terms of BER and CIR compared to the existing frequency domain correlative coding and self cancellation techniques. The designed window of leading coefficient and correlation order value of 1 gives an optimum design based on maximizing CIR.

Over10 dB performance gain has been obtained with employment of windowing compared to the standard OFDM system without any equalization techniques employed at the normalized frequency offset of $\varepsilon = 0.15$ and BER of $10^{-3}$. The time domain windowing scheme shows better tolerance to frequency offset by considerable reduction of the sensitivity to frequency errors.

## ACKNOWLEDGEMENT

Words are inadequate to express the overwhelming sense of gratitude and humble regards to my supervisor and advisor Dr.-Ing Hailu Ayele, for his constant motivation, support, expert guidance, constant supervision and constructive suggestion up to the submission of my thesis work.

This thesis work would have been impossible if not for the perpetual moral support from my family members, and my friends. I would like to thank them all.

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## LIST OF SYMBOLS

| | |
|---|---|
| N | Number of OFDM subcarriers / IFFT bin size |
| k, p | Discrete frequency indices (from 0 to N-1) |
| n | Time sample |
| $S_{k,p}$ | Frequency spectrum of subcarrier k (for p from 0 to N-1) |
| X(k),Y(k) | Frequency domain symbol at subcarrier k |
| x(n), y(n) | Time domain signals at time sample n |
| h(n) | Discrete channel impulse response |
| H(k) | frequency domain channel response |
| w(n) | Additive White Gaussian Noise |
| $\otimes$ | Convolution operator |
| C(k) | Carrier signal at the $k^{th}$ subcarrier |
| I(k) | Interference signal to the $k^{th}$ subcarrier |
| ε | Normalized frequency offset |

## LIST OF ABBREVIATIONS

AWGN - Additive White Gaussian Noise

BER - Bit Error Rate

BPSK - Binary Phase Shift Keying

CFO – Carrier Frequency Offset

CIR - Carrier to Interference Ratio

CP – Cyclic Prefix

DAB - Digital Audio Broadcasting

(I)DFT – (Inverse) Discrete Fourier Transform

DMT - Discrete Multi Tone

DVB - Digital Video Broadcasting

(I)FFT – (Inverse) Fast Fourier Transform

FEC - Forward Error Correction

ICI - Inter Carrier Interference

ISI - Inter Symbol Interference

LAN - Local Area Network

MCM - Multi Carrier Modulation

MF – Maximally Flat

MLSD – Maximum Likelihood Sequence Detection

OFDM - Orthogonal Frequency Division Multiplexing

QAM - Quadrature Amplitude modulations

QPSK - Quadrature Phase Shift Keying

PAPR - Peak to Average Power Ration

SNR - Signal to Noise Ratio

# CHAPTER ONE

# INTRODUCTION

This chapter consists of four parts: introduction to OFDM history, research background and motivation, literature survey, and organization of the thesis. We begin by introducing a brief history of OFDM.

## 1.1 History of OFDM

Orthogonal frequency division multiplexing (OFDM) is one of the multi-carrier modulation (MCM) techniques that transmit signals through multiple carriers. These carriers (subcarriers) have different frequencies and they are orthogonal to each other.

Although OFDM currently attracts much concern of researchers, it is an old concept. The original OFDM principle was first introduced by Chang in 1966 [1], based on the multicarrier modulation technique used in the high frequency military radio. However, the modulation, synchronization, and coherent demodulation induced were complicated for more subcarriers requiring additional hardware cost. In 1971, Weinstein and Ebert proposed a modified OFDM system [2] in which the discrete Fourier Transform (DFT) was applied to generate the orthogonal subcarrier waveforms. Their scheme reduced the implementation complexity significantly, by making use of the IDFT modules and the digital-to-analog converters. Cyclic prefix (CP) or cyclic extension was first introduced by Peled and Ruiz in 1980 [3] for OFDM systems. In their scheme, conventional null guard interval is substituted by cyclic extension for fully-loaded OFDM modulation. As a result, the orthogonality among the subcarriers was guaranteed. In 1985, Cimini introduced a pilot-based method to reduce the interference emanating from the multipaths and co-channels [4]. In the 1990s, OFDM systems have been exploited for high data rate communications. In the IEEE 802.11 standard, the carrier frequency can go up as high as 2.4 GHz or 5 GHz. Researchers tend to pursue OFDM operating at even much higher frequencies nowadays. For example, the IEEE 802.16 standard proposes yet higher carrier frequencies ranging from 10 GHz to 60 GHz [5].

High capacity and variable bit rate information transmission with high bandwidth efficiency are just some of the requirements that the modern transceivers have to meet in order for a variety of new high quality services to be delivered to the customers.

Because in the wireless environment signals are usually impaired by fading and multipath delay spread phenomenon, traditional single carrier mobile communication systems do not perform well. In such channels, extreme fading of the signal amplitude occurs and Inter Symbol Interference (ISI) due to the frequency selectivity of the channel appears at the receiver side. This leads to a high probability of errors and the system‘s overall performance becomes very poor. Orthogonal Frequency Division Multiplexing (OFDM) thus has attracted much interest for advantages it provides.

## 1.2 Research Background and Motivation

OFDM systems have recently gained increased interest. One of the main reasons to use OFDM is to increase the robustness against frequency selective fading and narrow band interference. Data bearing symbol stream is split into several lower rate streams which are to be transmitted on different carriers; this increases the symbol period by the number of non-overlapping carriers (sub-carriers) and then multipath will affect only a small portion of the neighboring symbols. This provides resistance against frequency-selectivity of the channel for wideband data transmission. The remaining ISI can be removed by cyclically extending the OFDM symbol. The length of the cyclic extension should be at least as long as the maximum excess delay of the channel.

In OFDM, the sub-carriers are totally independent and orthogonal to each other. The sub-carriers are placed exactly at the nulls in the modulation spectrum of one another. At the peak point of one sub-carrier waveform, the sample values of other sub-carriers at the nulls are zeros and thus contribute no ICI to the sampled sub-carrier. This is where the high spectral efficiency of OFDM comes from. It can be shown that keeping the orthogonality of the sub-carriers is very critical for an OFDM system to be free from inter-carrier interference.

Orthogonal frequency division multiplexing (OFDM) is the projected modulation of choice for fourth-generation broadband multimedia wireless systems [6]. However, it suffers from inter-carrier interference (ICI), and some inter-symbol interference. ICI results when the orthogonality of the carriers can no longer be maintained due to vulnerability to frequency offset errors caused by oscillator inaccuracies and Doppler shift [7, 8]. Depending on the Doppler spread in the channel and the block length chosen for transmission, ICI can potentially cause severe deterioration of quality of service in OFDM systems.

While OFDM solves the ISI problem by using cyclic prefix, it has another self-interference problem: Inter-carrier Interference (ICI), or the crosstalk among different sub-carriers, caused by the loss of orthogonality due to frequency instabilities. ISI and ICI are dual of each other occurring in different domains; one in time domain and the other in frequency-domain. ICI is a major problem in multi-carrier systems and needs to be taken into account when designing systems. Therefore, efficient cancellation of ICI is very crucial.

To mitigate the effect of intercarrier interference different techniques have been proposed by researchers [9, 10, 11]. The Cyclic prefix (the copy of a certain part of the symbol at the back and add at the front) is a crucial feature of OFDM used to combat inter-symbol interference (ISI) and inter-carrier interference (ICI) introduced by the multipath channel through which the signal propagates. Although ISI can be removed by cyclically extending the OFDM symbol, an equalizer is needed in an OFDM system to mitigate ICI and increase the transmission efficiency.

Combined with the facts that the spectrum is a scarce resource and propagation conditions are hostile due to time varying fading and interference from other subcarriers of OFDM system, this work proposes to investigate equalizer techniques as means to make the use of OFDM spectrally efficient through an equalizer based on time domain windowing.

### 1.3 Literature Survey

Equalization is a technique used to help accomplish recovering the original signal with the best possible signal to noise ratio (SNR) [12]. In OFDM ICI reduction is currently the main area of research in mobile wireless communication. This section presents a brief survey of basic literature survey. But more of other relevant literatures are reviewed in Chapter three.

Since equalization embodies a sophisticated set of signal processing techniques, making it possible to compensate for channel induced interference, it is an important area of many researches. A block minimum mean squared error (MMSE) equalizer for orthogonal frequency-division multiplexing (OFDM) systems over time-varying multipath channels is presented in [13]. The complexity of their method increases linearly with the number of subcarriers.

Seyedi and Saulnier emphasize in [14] that self cancellation technique is a better ICI reduction technique than the frequency offset estimation techniques for considerable computational complexity minimization. They investigate coarse frequency-offset estimation (which estimates CFO multiple of the subcarrier spacing) and fine frequency-offset estimation (which estimates CFO <half the subcarrier spacing) together with self cancellation technique. They showed that although there are many methods that can estimate and remove the frequency offset quite accurately, they often have considerable computational complexity. Thus it is better to use signal processing and/or coding to reduce the sensitivity of the OFDM system to the frequency offset. These methods can either be used as low complexity alternatives to frequency-offset estimation techniques or they can be used together with a somewhat accurate oscillator. Their proposed self cancellation scheme is also very effective in reducing the ICI when the OFDM system operates over a fast fading channel. They also propose that windowing is one technique to reduce the ICI created as a result of frequency offset. The performance of the self cancellation technique is demonstrated in this thesis for comparison with the time domain windowing equalization technique.

A study of ICI self-cancellation scheme for combating the effects of ICI due to the frequency offset between the transmitted and received carrier frequencies was presented in [14, 15, 16] with Zhoa & Haggman *et al.* [16] considered to be the pioneers. The technique is based on modulating one signal onto a group of subcarriers with specially defined weighting coefficients. At the receiver side, the signals received within a group are linearly combined with the same weighting coefficients. All the papers agree that this technique achieves the ICI reduction and BER performance improvement at the cost of lowering the transmission rate and reducing the bandwidth efficiency.

Cheng, Jiao, and Lee, *et al.* [8] developed a dual-window technique to reduce the sensitivity to carrier frequency offset (CFO). They define two windows and use them alternately onto the adjacent subcarriers for pulse shaping at the transmitter. The receiver selects one of the two windows to maximize the output of the desired subcarriers and suppresses the others by window- matching and anti-matching functions respectively. By introducing two pulse-shaping windows one for even and the other for odd number of the total subcarriers at the transmitter and the receiver, the proposed technique shows the efficiency in reducing the ICI. The simulation results confirm this approach by the obvious improvements on the BER performance. They stated that more complexity in numerical calculations is generated by using optimum coefficients for further improved performance.

Kumar, Malarvizhi and Jayashri in [6] proposed a time domain equalization technique based on the window function which creates a correlation between two adjacent subcarriers and gives a higher signal-to-ICI ratio than standard OFDM. In their paper the 1-D correlative polynomial is used in the frequency domain to suppress the ICI. They propose a window function in equivalent to the Correlative polynomial used in the frequency domain. Time domain windowing technique proposed in their paper offers better BER performance compared to the correlative coding method. Their paper does not show the effects of increased window orders and appropriate demodulation techniques for better BER performance.

A time-domain windowing of OFDM signal to reduce the sensitively to carrier frequency offset was also described in [17], where the Nyquist window was used to suppress the side lobes of subcarriers. Nyquist-type time windows offer reduced side lobes when compared to original OFDM. The MMSE optimized solutions have significantly reduced side lobes directly adjacent to the main lobe. But MMSE Optimization of Window Shape is computationally intensive.

## 1.4 Objective

Several methods have been presented to reduce inter carrier interference (ICI); including frequency offset estimation and correction techniques [18], frequency domain equalization [19],[20], [21], ICI self-cancellation scheme [14], [15].

The main objective of this thesis is to investigate time domain equalization based on the optimized window function for ICI mitigation. Performances of different window correlation orders are compared and the windowing technique is compared with frequency domain correlative coding and self-cancellation techniques. The performance is measured in terms of the spectral efficiency, reduction in the BER and performance of the Carrier to Interference Ratio (CIR) in terms of the frequency offset.

**General objective**: Investigate the effectiveness of time domain windowing for mitigation of ICI in OFDM systems. Carry out computer simulations to evaluate the performance in ICI suppression. Carrier-to-interference ratio (CIR) and bit-error rate (BER) improvements will be studied compared to the standard OFDM systems will be studied.

**Specific objectives:**

- Implement time domain equalization technique using well designed window to suppress ICI and optimize the parameters through theoretical analysis
- Investigate the effects on CIR and BER under AWGN and time varying fading channel conditions
- Develop a method that can potentially prevent error propagation and evaluate its performance in ICI suppression.
- Compare performance with the existing frequency domain correlative coding, self cancellation and standard OFDM systems.
- Study the effect of the proposed window function on the inband and outband power spectrum.

### 1.5 Organization of Thesis

We begin by introducing the preliminary concepts about OFDM in Chapter two. This chapter is written based on study of the literature reading, covering the explanation of basic theory of OFDM transceiver and radio mobile channel. It focuses on the OFDM system operation, pros and cons of OFDM and an overview of the mobile wireless channel.

Chapter three discusses the problems of OFDM system and presents an analysis of intercarrier interference. It briefly introduces the multipath channel with carrier frequency offset and investigates the existing equalization techniques for ICI reduction.

The fruit of this research, analysis of time domain equalization for ICI reduction, is presented in Chapter four. This chapter presents the type of the window used with its special property and its difference from other existing pulse shaping and filtering windows. The carrier to interference signal power ratio and appropriate demodulation technique for the proposed window are also provided here.

Chapter five is where the computer simulation results and discussions are presented. We present the parameters that are used in this simulation and the graphical and numerical results that illustrate the main issues dealt within this work.

Finally, we make a few concluding remarks in Chapter six. It concludes the analysis of the results and proposes some future works that can be done in order to extend the current research further.

# CHAPTER TWO
# BASICS OF OFDM THEORY

## 2.1 Basics of OFDM

Most of the transmission systems experience degradations, such as large attenuation, noise, multipath and interferences. High capacity and variable bit rate information transmission with high bandwidth efficiency are just some of the requirements that the modern transceivers have to meet in order for a variety of new high quality services to be delivered to the customers [22]. Because in the wireless environment signals are usually impaired by fading and multipath delay spread phenomenon, traditional single carrier mobile communication systems do not perform well. In such channels, extreme fading of the signal amplitude occurs and ISI due to the frequency selectivity of the channel appears at the receiver side. This leads to a high probability of errors and the system‘s overall performance becomes very poor. One physical-layer technique that has recently gained much popularity due to its robustness in dealing with these impairments is multi-carrier modulation. Multi-carrier modulation is the concept of splitting a high data stream into a number of low rate streams modulating separate orthogonal frequencies and combining the data received on the multiple channels at the receiver. OFDM is an example of multi carrier system.

OFDM is simply defined as a form of multi-carrier modulation where the carrier spacing is carefully selected so that each sub carrier is orthogonal to the other sub carriers. Since the demand for efficient use of bandwidth and high data rate services has been increasing very rapidly, OFDM is emerging as the preferred modulation scheme in modern high data rate wireless communication systems. In OFDM high rate bit-stream is split into a number of (say N) parallel bit-streams of lower rate and each of these are modulated using one of N orthogonal sub-carriers. A large number of orthogonal, overlapping, narrow band sub-carriers are transmitted in parallel. The separation of the sub-carriers is such that there is a very compact spectral utilization. The basic idea of OFDM is to divide the available spectrum into several orthogonal sub channels so that each narrowband sub channels experiences almost flat fading. Thus, OFDM can provide large data rates with sufficient robustness against radio channel impairments.

OFDM is emerging as the preferred modulation scheme in modern high data rate wireless communication systems. OFDM has been adopted in the European digital audio and video broadcast radio system and is being investigated for broadband indoor wireless communications. Standards such as HIPERLAN2 (High Performance Local Area Network) and IEEE 802.11a and IEEE 802.11g have emerged to support IP-based services [2].

### 2.2 Basic OFDM system model

The basic model of OFDM system with major components is shown in Figure 2.1. The basic components of the system model are also described in this section.

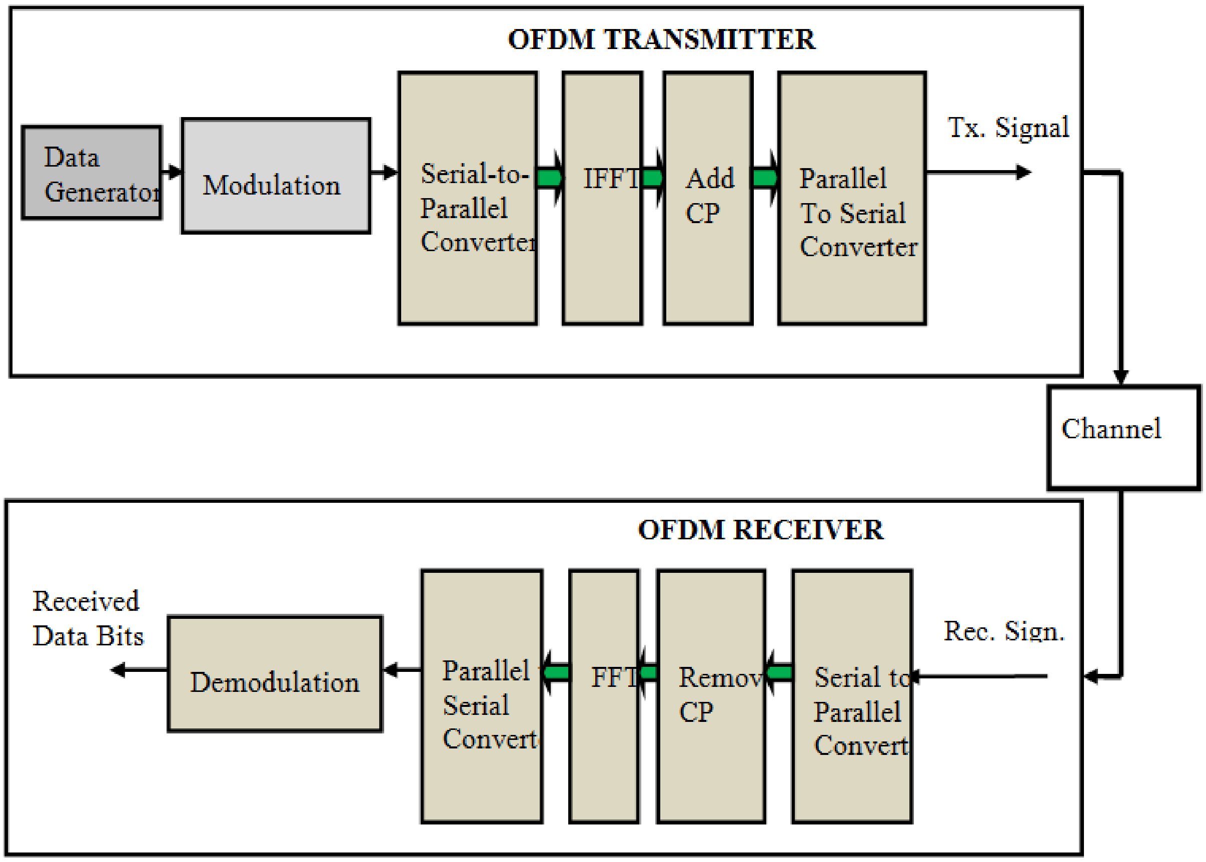



Figure 2.1- Basic OFDM system model

**Serial to Parallel Conversion**: In an OFDM system, each channel can be broken into various sub-carriers. The use of sub-carriers makes optimal use of the frequency spectrum but also requires additional processing by the transmitter and receiver. This additional processing is necessary to convert a serial bit stream into several parallel bit streams to be divided among the individual carriers. Once the bit stream has been divided among the individual sub-carriers, each sub-carrier is modulated as if it was an individual channel before all channels are combined back together and transmitted as a whole.

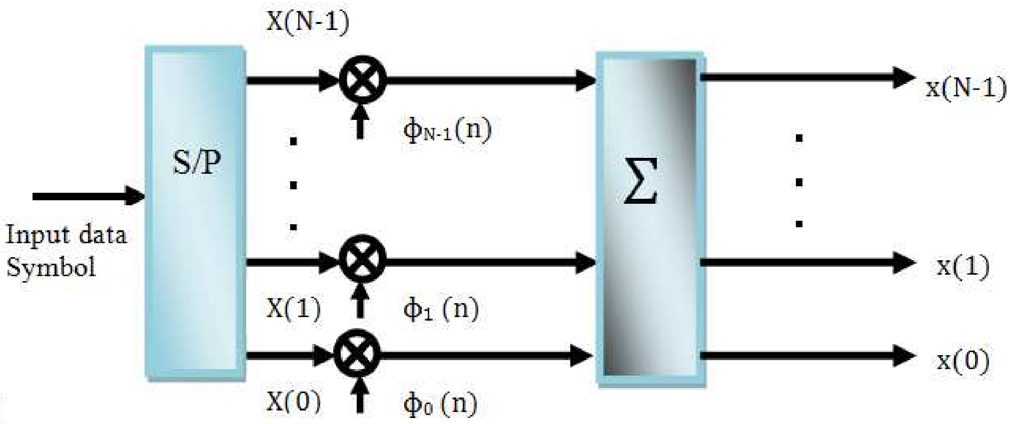
The receiver performs the reverse process to divide the incoming signal into appropriate sub-carriers and then demodulating these individually before reconstructing the original bit stream [10].

**Modulation with the Inverse FFT**: OFDM transmits a large number of narrowband carriers, closely spaced in the frequency domain. In order to avoid a large number of modulators and filters at the transmitter and complementary filters and demodulators at the receiver, it is desirable to be able to use modern digital signal processing techniques, such as fast Fourier transform (FFT).

Inverse FFT at the transmitter and FFT at the receiver are key components in the OFDM performing linear mappings between N complex data symbols and N complex OFDM symbols result in robustness against fading multipath channel. This is achieved by transforming the high data rate stream into N low data rate streams, each experiencing flat fading during transmission over a wireless channel.

At the transmitter, the signal is defined in the frequency domain, X(k). The role of the IFFT is to modulate each sub-channel onto the appropriate carrier. In fact, the modulation scheme can be chosen completely independently of the specific channel being used and can be chosen based on the channel requirements. It is possible for each individual sub-carrier to use a different modulation scheme. The modulation of data into a time domain complex waveform occurs at the Inverse Fast Fourier Transform (IFFT) stage. The total number of subcarriers translates into the number of points of the IFFT/FFT. The amplitudes and phases of the carriers depend on the data to be transmitted. The data transitions are synchronized at the carriers, and can be processed together, symbol by symbol.

Consider Figure 2.2 below to visualize how IFFT block replaces subcarrier modulation techniques.




Where, $\phi_k(\mathrm{n}) = \frac{1}{\sqrt{N}} e^{\frac{j2\pi kn}{N}}, n = 0,1,\dots,N$

Figure 2.2- Discrete time OFDM system with N-subcarriers

Note:

- Inputs to IFFT are parallel frequency domain data streams each controls signal at one frequency
- Outputs of IFFT are discrete time samples of modulated and multiplexed signals

Suppose the data set to be transmitted is X(0),X(1), ...,X(N-1), where N is the total number of sub-carriers. The discrete-time representation of the signal after IFFT is:

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k).e^{j2\pi k\frac{n}{N}}, \quad n = 0,1,\dots,N-1$$

The signal x(n) can be determined as:

$$x(n) = \left\{ \begin{matrix} \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi\frac{k}{N}n}, 0 \leq n \leq N-1 \\ 0, \qquad otherwise \end{matrix} = IFFT\{X(k)\} \right\}$$

At the receiver side, the data is recovered by performing FFT on the received signal,

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x(n).e^{-j2\pi k\frac{n}{N}}, \quad k = 0,1,\dots,N-1 \tag{2.1}$$

Consider the OFDM signal x(n) passes through an ideal channel to visualize the IFFT/FFT implementation, the discrete time representation at the receiver will be:

$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x(n) e^{-j2\pi \frac{m}{N} n} = FFT\{x(n)\} = \sum_{m=0}^{N-1} X(m)\delta(m-k) = X(k)$ , where $X(k)$ is the original input data symbol fed to the serial to parallel converter at the transmitter at subcarrier k. For the channel with channel impulse response of h(n) and additive white Gaussian noise w(n), the received signal from the channel is:

$$r(n) = s(n) \otimes \mathrm{h(n)} + \mathrm{w(n)} \tag{2.2}$$

$$\begin{aligned} Y(k) = DFT\big(r(n)\big) &= \mathrm{DFT(IDFT(X(k))} \otimes \mathrm{h(n) + w(n))} \\ &= X(k).\,\mathrm{DFT(h(n)) + DFT(w(n))} \\ &= X(k).\,\mathrm{H(k) + W(k)} \ , 0 \le \mathrm{n} \le \mathrm{N} - 1 \end{aligned}$$

Where, $\otimes$ denotes circular convolution and $\mathrm{W}(k) = \mathrm{DFT(w(n))}$.

**Guard Interval and Cyclic Prefix:**

There are several options for GI. The GI is introduced initially to eliminate the inter block interference (IBI). Since one block of input data symbols is associated with a single transmitted waveform in an OFDM system, most people refer IBI as ISI. Without GI, many time-delayed visions of the transmitted waveform would interfere with each other. Nevertheless, in those cases where the GI is employed, the portions of waveforms received in the GI duration would be totally discarded. Thus, the ISI could be completely eliminated accordingly. It is noted that the GI duration must be larger than the maximum channel delay time. Otherwise, it could not entirely remove the ISI.

One choice of GI is zero padding. In this scheme, no waveform is transmitted in the GI duration. However, the zero-padded waveform would destroy the orthogonality of subcarriers and results in intercarrier interference (ICI). The cyclic prefix (CP) is a good substitute of the zero-padding GI. Cyclic prefix is a crucial feature of OFDM to combat the effect of multipath.

The expressions of the subcarrier waveforms after the addition of cyclic prefix are now given by

$$s(n) = \begin{cases} x(n+N) & -M \le n \le 0 \\ x(n) = \dfrac{1}{\sqrt{N}} \displaystyle\sum_{k=0}^{N-1} X(k).e^{j2\pi k \frac{n}{N}} & 0 \le n \le N-1 \end{cases}$$

As depicted in Figure 2.3, an end-portion of waveform is copied and inserted prior to the beginning of waveform. Given that transmission of cyclic prefix reduces the data rate, the system designers will want to minimize the cyclic prefix duration. Typically, cyclic prefix duration is determined by the expected duration of the multipath channel in the operating environment.

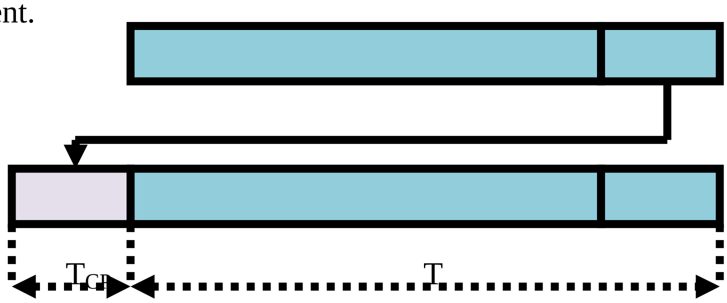



Figure 2.3- Generation of Cyclic Prefix

The time duration of CP, denoted as $T_{CP}$ in the Figure 2.3, is often chosen according to the following relation: $T_{CP} = \frac{T}{2^m}$ [23] where, m is an integer and T is the duration of OFDM symbol. For example, in the IEEE 802.11a standard, $m$ = 2 is chosen; in Europe DVB (Digital Video Broadcasting) standards, $m$ = 1, 2, …, 6 can be employed. As aforementioned, $m$ should also depend on the maximum delay time of the channel.

One of the main advantages of OFDM is its effectiveness against the multi-path delay spread which frequently encountered in Mobile communication channels. The reduction of the symbol rate by N times, results in a proportional reduction of the relative multi-path delay spread, relative to the symbol time. To completely eliminate even the very small ISI that results, a guard time is introduced for each OFDM symbol. The guard time must be chosen to be larger than the expected delay spread, such that multi-path components from one symbol cannot interfere with the next symbol. This Guard time acts as a buffer region where delayed information from the previous symbols can get stored. The receiver has to exclude samples from the cyclic prefix which got corrupted by the previous symbol when choosing the samples for an OFDM symbol.

Another advantage with the cyclic prefix is that it serves as a guard between consecutive OFDM frames. This is similar to adding guard bits, which means that the problem with inter frame interference also will disappear.

**Parallel to Serial Conversion**: Once the cyclic prefix has been added to the sub-carrier channels, they must be transmitted as one signal. Thus, the parallel to serial conversion stage is the process of summing all sub carriers and combining them into one signal. As a result, all sub-carriers are generated perfectly simultaneously.

### 2.3 OFDM subcarrier orthogonality and spectral analysis

**Orthogonality of Sub-Channel Carriers**: Orthogonality can be achieved by carefully selecting carrier spacing, such as letting the carrier spacing be equal to the reciprocal of the useful symbol period. As the sub carriers are orthogonal, the spectrum of each carrier has a null at the center frequency of each of the other carriers in the system. This results in no interference between the carriers, allowing them to be spaced as close as theoretically possible.

Two signals are orthogonal if their dot product is zero. That is, if one takes two signals, multiply them together and if their integral over an interval is zero, then two signals are orthogonal in that interval. For symbol duration T, number of subcarriers N, and subcarrier frequency spacing of Δf, Set of subcarriers:

$f_n(t) = e^{\mathrm{j}2\pi \mathrm{n}\Delta \mathrm{f}\mathrm{t}}, n = 0,1,2 \dots \mathrm{N} - 1 \; and \; 0 \leq t \leq T$ are orthogonal if:

$$\frac{1}{N}\int_0^T f_n(t) f_m^*(t)dt = \begin{Bmatrix} 0, & n \neq m \\ 1, & n = m \end{Bmatrix}$$

Figure 2.4 illustrates the frequency domain of 3 OFDM subcarriers graphically such that each sub-carrier is represented by a different peak. In addition, the peak of each sub-carrier corresponds directly with the zero crossing of all the other sub channels [25].

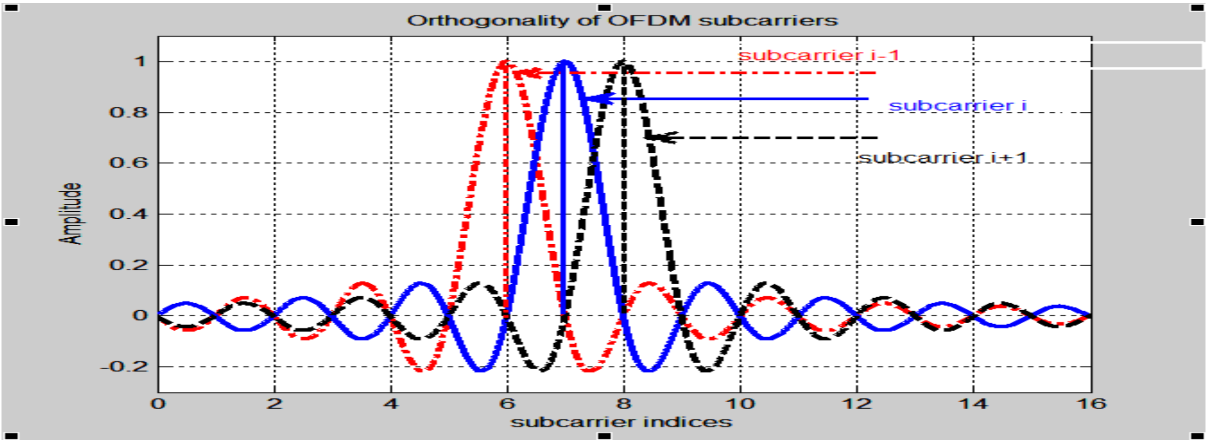



Figure 2.4- Orthogonality principle of OFDM

The common understanding about subcarrier spectrum of Orthogonal Frequency Division Multiplexing (OFDM) systems is a sinc(x) function [24]. In the presence of ICI the OFDM subcarrier spectrum of subcarrier k can be obtained from a common sinc(x) function with a certain phase rotation, which is different from common knowledge of sinc(x) function OFDM spectrum.

When the channel frequency errors do not exist, subcarrier spectrums give the same values as sinc(x) function at each subcarrier. However, for the system with frequency offsets, the ICI signals caused by the spectrum shift are not the same. The subcarrier spectrum shape determines the ICI signal values.

In OFDM a single channel utilizes multiple sub-carriers on adjacent frequencies. The sub-carriers in an OFDM system are overlapping to maximize spectral efficiency. Ordinarily, overlapping adjacent channels can interfere with one another. Figure 2.5 shows the frequency domain representation of a sample of six sub-carriers for set of 16 subcarriers in individual channel. Because the symbol rate increases as the channel bandwidth increases, OFDM implementation allows for greater data throughput.

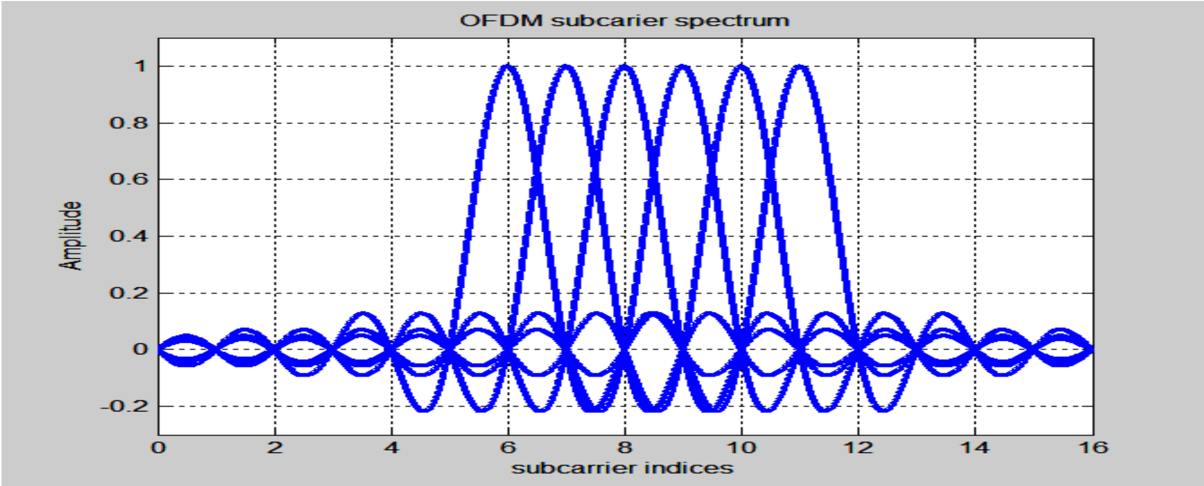



Figure 2.5- frequency domain representation of an OFDM system

In a normal OFDM signal, the complex values modulating the subcarriers in each symbol period are statistically independent of each other [4]. They are also independent of the values modulating any subcarrier in any previous or subsequent symbol period. As a result the power spectrum of the overall signal can be found by summing the power spectra of all individual subcarriers for any symbol period. It is known that power spectrum of individual subcarriers is obtained from the Fourier transform of the time domain signals.

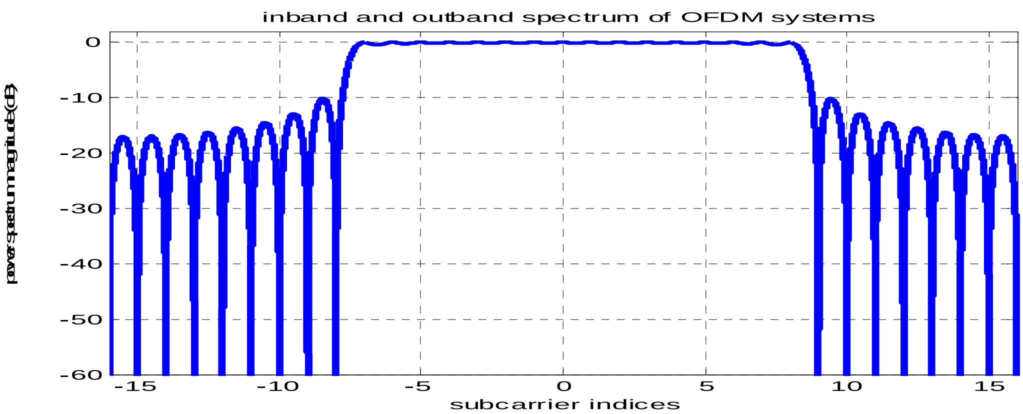



Figure 2.6- Power spectrum of the transmitted OFDM signal

The spectrum of each subcarrier decreases according to a sinc function. The sinc functions have side lobes that are relatively large and do not decay quickly with frequency. As a result, the spectral rolloff of OFDM signals is slow. Thus as we can see in figure 2.6 above, the outband power is not low enough for many OFDM application systems.

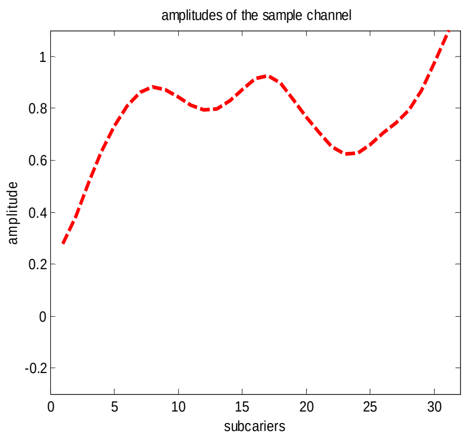
The out of band power should be minimized to avoid interference between adjacent broadcast channels [25]. It is demonstrated in chapter five of this paper that the time domain equalization using windowing reduce the out of band power in addition to bring better BER performance. This eliminates the sharp transitions at symbol boundaries in the time domain signal and results in more rapid spectral rolloff.

## 2.4 Strength and Weakness of OFDM

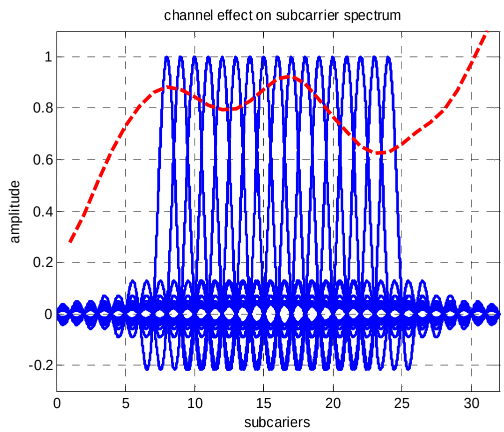
OFDM has several advantages over single carrier modulation systems and these make it a viable alternative for CDMA in future wireless networks. The major advantages of OFDM are its ability to convert a frequency selective fading channel into several nearly flat fading channels and high spectral efficiency.

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

In a single carrier system the entire signal is lost during the fading intervals. But as in case of OFDM the signal consists of many sub-carriers, so only few sub-carriers are affected during the fading intervals & hence a very small percentage of the signal is lost which can be easily recovered.




(a) channel frequency response




(b) Signal frequency spectrum

Figure 2.7- Robustness of OFDM to Frequency Selective fading channel

The robustness of OFDM to frequency selective fading channels is illustrated graphically in Figure 2.7 where the channel is divided into narrowband flat fading sub channels.

Although OFDM is more advantageous than single carrier systems, it has also disadvantages. One of the main disadvantages of OFDM is its sensitivity against carrier frequency offset which causes attenuation and rotation of subcarriers, and Inter carrier interference (ICI).

In general, here are some of the advantages and disadvantages of OFDM system.

OFDM advantages:

- Spectral efficiency: the orthogonal sub channels are closely spaced and overlap in frequency
- Ability to combat ISI caused by multipath delay spread
- Robust against frequency selective fading channels
- Efficient implementation by IFFT/FFT
- Low sensitivity to time synchronization errors
- Simple channel equalization instead of complex adaptive channel equalization

OFDM disadvantages:

- Sensitivity to frequency offset (due to Doppler shift and frequency synchronization problems): results in ICI
- High Peak to average power ratio (High power transmitter amplifiers need linearization, Low noise receiver amplifiers need large dynamic range)
- Capacity and power loss due to guard interval (Bandwidth and power loss due to the guard interval can be significant)

### 2.5 Overview of Mobile Wireless channel

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

AWGN and fading are the two common channel models.

### 2.5.1 AWGN Channel

The most common channel model is the Gaussian channel, which is generally called the additive white Gaussian noise (AWGN) channel. When signal is transmitted through the channel, it is corrupted by the statistically independent Gaussian noise. This channel model assumes that the only disturber is the thermal noise at the front end of the received symbol. Typically thermal noise has a flat power spectral density over the signal bandwidth.

The AWGN channel is simple and usually it is considered as the starting point to develop the basic system performance results.

### 2.5.2 Fading Channel

Fading is a random fluctuation of the transmitted signal experienced differences in attenuation, delay and phase shift while travelling from the source to the receiver. The effects of multipath include constructive and destructive interference and phase shifting of the signal. This causes Rayleigh Fading named after Lord Rayleigh [26]. Rayleigh fading with a strong line of sight is said to have a Rician distribution or Rician fading.

The most common types of fading are known as "slow fading" and "fast fading" as they apply to a mobile radio environment (classified based on Doppler spread or Signal time spread versus channel time variance).

**Slow fading:** refers to the time variation of the received signal power caused by changes in the transmission medium or path. It has a low Doppler spread (the Doppler spread of the channel is much less than the bandwidth of the baseband signals). The coherence time is greater than the symbol period and the channel variations are slower than the baseband signal variation.

**Fast fading:** also known as Multipath fading or small scale fading is a kind of fading occurring with small movements of a mobile. When there is relative motion between the transmitter and the receiver, Doppler spread is introduced in the received signal spectrum, causing frequency dispersion. If the Doppler spread is significant relative to the bandwidth of the transmitted signal, the received signal is said to undergo *fast fading*. This form of fading typically occurs for very low data rates. It has a high Doppler spread and the coherence time is less than the symbol time and the channel variations are faster than baseband signal variation.

Another effect that affects digital transmission is that the signal coming from different paths has different time delays depending on the length of path (delay spread). There are two types of fading based on multipath time delay spread:

**Flat fading:** A received signal is said to undergo *flat fading*, if the mobile radio channel has a constant gain and a linear phase response over a bandwidth larger than the bandwidth of the transmitted signal. Under these conditions, the received signal has amplitude fluctuations due to the variations in the channel gain over time caused by multipath. However, the spectral characteristics of the transmitted signal remain intact at the receiver.

**Frequency selective fading:** One problem of multipath is that the resultant of waves from different paths can be constructive or destructive depending on the position, so signal change over time when moving. Characteristics may change fast when moving. That‘s why the channel is also time varying. If the mobile radio channel has a constant gain and linear phase response over a bandwidth smaller than that of the transmitted signal, the transmitted signal is said to undergo *frequency selective fading*. This is when the bandwidth of the signal is greater than the coherence bandwidth of the channel or the delay spread greater than the symbol period.

In this chapter the basics of OFDM system including its operation, system model, its strength and weakness together with a brief introduction to the mobile wireless channels have been discussed. The next chapter, chapter 3, deals with the ICI problem of OFDM systems in detail and the solutions proposed by many researchers.

# CHAPTER THREE

# INTERCARRIER INTERFERENCE AND EQUALIZATION TECHNIQUES IN OFDM

## 3.1 Problems in OFDM

The two main OFDM problems are **high peak-to-average power ratio** and **intercarrier interference.** In high peak-to-average power ratio peak signals power is much greater than average signal power which needs linear amplifiers with large dynamic range. Intercarrier interference, the point of concern in this thesis, needs tight specifications for local oscillators and Doppler limitations.

### 3.1.1 High Peak-to-Average Power Ratio (PAPR)

One of the major drawbacks of OFDM is the high peak-to-average power ratio (PAPR) which results in high out-of-band radiation, and bit error rate performance degradation, mainly due to the nonlinearity of the high power amplifier as reported in [27].

The PAPR of the transmit signal x(n), in discrete time, is defined as the ratio of the maximum instantaneous power and the average power.

$$PAPR = \frac{max\ \ |x(n)|^2}{E\{|x(n)|^2\}}, 0 \leq n \leq N-1 \tag{3.1}$$

PAPR of OFDM increases exponentially with the number of subcarriers. When N sinusoids add, the peak magnitude would have a value of N, where the average might be quite low due to the destructive interference between the sinusoids. High PAPR signals are usually undesirable for it usually strains the analog circuitry. High PAPR signals would require a large range of dynamic linearity from the analog circuits which usually results in expensive devices and high power consumption with lower efficiency (for e.g. power amplifier has to operate with larger back-off to maintain linearity).

OFDM transmitters therefore require power amplifiers with large linear range of operation which are expensive and inefficient. Any amplifier non-linearity causes signal distortion and inter-modulation products resulting in unwanted out-of-band power and higher BER [28]. If power amplifiers are not operated with large power back-offs, it is impossible to keep the out-of-band power below the specified limits. This situation leads to very inefficient amplification and expensive transmitter, so it is highly desirable to reduce the PAPR.

There are different techniques proposed by researchers to reduce the PAPR. Some of these techniques as described in [27] and [28] are:

- Clipping and Filtering
- Selective mapping
- Partial Transmit Sequence

**3.1.2 Intercarrier interference**

Frequency offset is a critical factor in OFDM system design. It results in inter-carrier interference (ICI) and degrades the orthogonality of sub-carriers. Frequency errors will tend to occur from two main sources: synchronization errors and Doppler shift [29].

It may be assumed that most of the wireless receivers cannot make perfect frequency synchronization. In fact, practical local oscillators are usually unstable, which introduce frequency offset. Although this small offset is negligible in traditional communication systems, it becomes a severe problem in the OFDM systems. In most situations, the oscillator frequency offset varies from 20 ppm (parts per million) to 100 ppm [20]. Given that an OFDM system operates at 5 GHz, the maximum offset would be 100 KHz to 500 KHz (20-100 ppm.). Hence, this much frequency offset could not be ignored.

The relative motion between receiver and transmitter, or a mobile medium among them, would result frequency shift in narrow-band communications which is known to be Doppler's effect. For example, the Doppler effect would influence the quality of a cell phone conversation in a moving car.

Intercarrier interference leads to degradation of the SNR. The amount of degradation is proportional to the normalized (fractional) frequency offset which is equal to the ratio of frequency offset to the carrier spacing. In this thesis, we consider the CFO due to the Doppler shift under both pedestrian and vehicular conditions.

### 3.2 Analysis of Intercarrier interference

ICI is a special problem in the OFDM system. ICI problem would become more complicated when the multipath fading is present [30]. Before we look at the different equalization techniques used for reduction of ICI, we will discuss the effect of multipath channel with carrier frequency offset.

#### 3.2.1 Multipath channel with carrier frequency offset

In this research, we consider both AWGN and fading channel models with carrier frequency offset. The multipath fading does not cause ICI, but it will make the ICI problem worse. Since ICI cannot be neglected in practice, the impact of multipath fading should be discussed. It is recognized that the cyclic prefix has been used to eliminate ISI entirely and therefore only ICI needs to be concerned. Because there are many time-delayed versions of received signals with different gains and different phase offsets, the ICI is more complicated to calculate.

A typical mobile radio channel is one suffering from severe multipath fading and the carrier frequency offsets lead to severe degradation of the bit error rate (BER) performance. In an OFDM system, as the signals in each channel are of low flow rate (or have larger symbol duration) channel can be considered as frequency flat. This is true when the coherence bandwidth of the channel is greater than the symbol rate. But, the real channel is frequency selective and the OFDM system exhibits diversity effect in a frequency selective channel. This is the major benefit in using OFDM system in a multipath fading channel.

To show the channel model with carrier frequency offset, let’s consider a general multipath fading channel with carrier frequency offset.

In mobile radio environment, the time variant impulse response model of the multipath channel for one data block using discrete time domain index *n* is defined [31] **as**

$$h(n) = \sum_{m=0}^{M-1} h_m \mathrm{e}^{\frac{\mathrm{j}2\pi\varepsilon_{\mathrm{m}}(n-n_m)}{N}} \tag{3.2}$$

*N* = the number of subcarriers

$M$= number of paths

$n_m$ = the delay samples number of the m[th] path. For each path, the amplitude of $h_m$ is Rayleigh distributed.

$\varepsilon_m$ = is the normalized frequency offset of the m[th] path determined from the vehicular Doppler bandwidth which is

$$\varepsilon_m = \frac{f_{Dm}}{\Delta f} = \frac{f_c V}{c\Delta f} \tag{3.3}$$

$f_c$= carrier frequency
***V*** = speed of the vehicle
c = speed of light in the air
$f_{Dm}$ = the Doppler frequency of m[th] path
$\Delta$f = subcarrier frequency spacing

$\varepsilon_m$, the normalized frequency offset of m[th] path, is a more efficient parameter when analyzing frequency offset impact in OFDM systems.

Consider one typical path with amplitude of $h_m$, delay chip number $n_m$ and the normalized frequency offset $\varepsilon_m$, the channel impulse response of the m[th] path can then be expressed by

$$h_m(n) = h_m e^{\frac{j2\pi\varepsilon_m(n-n_m)}{N}} \tag{3.4}$$

Therefore, the corresponding frequency domain response can be obtained by FFT, which gives

$$H_m(k) = \frac{1}{N}\sum_{n=0}^{N-1} h_m\left(e^{\frac{j2\pi\varepsilon_m(n-n_m)}{N}}\right)e^{\frac{-j2\pi nk}{N}}$$

$$H_m(k) = \frac{h_m}{N}e^{\frac{-j2\pi\varepsilon_m n_m}{N}}\sum_{n=0}^{N-1} e^{\frac{-j2\pi n(k-\varepsilon_m)}{N}}$$

By using the mathematical relation, $\sum_{n=0}^{N-1} a^n = \frac{1-a^N}{1-a}\ for\ |a| < 1$ where $a = e^{\frac{-j2\pi(k-\varepsilon_m)}{N}}$, which is the property of finite geometric series; this can alternatively be expressed as:

$$\sum_{n=0}^{N-1} e^{\frac{-j2\pi n(k-\varepsilon_m)}{N}} = \frac{1-e^{-j2\pi(k-\varepsilon_m)}}{1-e^{\frac{-j2\pi(k-\varepsilon_m)}{N}}} = \frac{e^{-j\pi(k-\varepsilon_m)}\left[e^{j\pi(k-\varepsilon_m)} - e^{-j\pi(k-\varepsilon_m)}\right]}{e^{\frac{-j\pi(k-\varepsilon_m)}{N}}\left[e^{\frac{j\pi(k-\varepsilon_m)}{N}} - e^{\frac{-j\pi(k-\varepsilon_m)}{N}}\right]}$$

After some mathematical simplifications,

$$\sum_{n=0}^{N-1} e^{\frac{-\mathrm{j}2\pi \mathrm{n}(k-\varepsilon_{\mathrm{m}})}{N}} = \frac{e^{-\mathrm{j}\pi(k-\varepsilon_{\mathrm{m}})}}{e^{\frac{-\mathrm{j}\pi(k-\varepsilon_{\mathrm{m}})}{N}}} * \frac{\sin\left(\pi(k-\varepsilon_{\mathrm{m}})\right)}{sin(\frac{\pi(k-\varepsilon_{\mathrm{m}})}{N})}$$

$$\sum_{n=0}^{N-1} e^{\frac{-\mathrm{j}2\pi \mathrm{n}(k-\varepsilon_{\mathrm{m}})}{N}} = \frac{\sin\,(\pi(k-\varepsilon_{\mathrm{m}}))}{Nsin(\frac{\pi(k-\varepsilon_{\mathrm{m}})}{N})}\ e^{(-j(1-\frac{1}{N})\pi(k-\varepsilon_{\mathrm{m}}))}$$

Finally,

$$H_m(k)\ = h_m e^{\frac{-\mathrm{j}2\pi\varepsilon_{\mathrm{m}} n_m}{N}} * \frac{\sin\,(\pi(k-\varepsilon_{\mathrm{m}}))}{Nsin(\frac{\pi(k-\varepsilon_{\mathrm{m}})}{N})}\ e^{(-j(1-\frac{1}{N})\pi(k-\varepsilon_{\mathrm{m}}))} \tag{3.5}$$

Equation (3.5) is a function of $n_m$, $\varepsilon_{\mathrm{m}}$**,** and $k$ for a fixed $N$. The first exponential component $e^{\frac{-\mathrm{j}2\pi\varepsilon_{\mathrm{m}} n_m}{N}}$ is independent of $k$**,** which means that the received signals on any of the subcarriers are rotated by the phase angle of $\frac{-\mathrm{j}2\pi\varepsilon_{\mathrm{m}} n_m}{N}$.

When frequency offset $\varepsilon_{\mathrm{m}} \neq 0$ , Equation (3.5) represents the frequency domain response of the time varying channel with carrier frequency offset for a single path.

If all the paths are considered, the channel frequency domain response $H(k)$ becomes

$$H(k) = \sum_{m=0}^{M-1} h_m e^{\frac{-\mathrm{j}2\pi\varepsilon_{\mathrm{m}} n_m}{N}} * \frac{\sin\,(\pi(k-\varepsilon_{\mathrm{m}}))}{Nsin(\frac{\pi(k-\varepsilon_{\mathrm{m}})}{N})}\ e^{(-j(1-\frac{1}{N})\pi(k-\varepsilon_{\mathrm{m}}))} \tag{3.6}$$

Equation (3.6) is called *subcarrier frequency offset response* (SFO response). This function expresses the **ICI** property with respect to the system frequency offset in the time variant multipath radio channel. It is a basic function for analyzing frequency domain performance of OFDM systems.

The ICI analysis using the above channel in frequency domain gives the more general model and do not clearly reveal the structure on which ICI cancellation depends. To understand the different ICI cancellation methods let's consider the impairments due to carrier frequency offset.

The samples at the receiver before DFT due to the carrier frequency offset of an OFDM symbol are given by

$$y(n) = r(n) * e^{(\frac{j2n\pi\varepsilon}{N})} \tag{3.7}$$

Where r(n) represents the received time domain samples which is given by

$$r(n) = \sum_{k=0}^{N-1} X(k) * e^{(\frac{j2\pi \mathrm{kn}}{N})} \tag{3.8}$$

For OFDM systems, the received signal at subcarrier *l* can be expressed from the DFT as

$$Y(l) = \frac{1}{N}\sum_{n=0}^{N-1} y(n) * e^{\frac{-\mathrm{j}2\pi\mathrm{nl}}{N}}$$

$$Y(l) = \frac{1}{N}\sum_{n=0}^{N-1}\sum_{k=0}^{N-1} X(k) * e^{(\frac{\mathrm{j}2\pi\mathrm{kn}}{N})} * e^{(\frac{j2\pi\varepsilon\mathrm{n}}{N})} * e^{\frac{-\mathrm{j}2\pi\mathrm{nl}}{N}}$$

$$Y(l) = \frac{1}{N}\sum_{k=0}^{N-1} X(k) \sum_{n=0}^{N-1} e^{(\frac{j2\pi n(k-l+\varepsilon)}{N})} \tag{3.9}$$

By using finite geometric series relationship:

$$\sum_{n=0}^{N-1} e^{(\frac{j2\pi n(k-l+\varepsilon)}{N})} = \frac{\sin\,(\pi(l-k-\varepsilon))}{N sin(\frac{\pi(l-k-\varepsilon)}{N})}\; e^{(j(1-\frac{1}{N})\pi(l-k-\varepsilon))}$$

Thus,

$$Y(l) = \frac{1}{N}\sum_{k=0}^{N-1} X(k)\frac{\sin\,(\pi(l-k-\varepsilon))}{N sin(\frac{\pi(l-k-\varepsilon)}{N})}\; e^{(j(1-\frac{1}{N})\pi(l-k-\varepsilon))} \tag{3.10}$$

$$Y(l) = X(l)S(-\varepsilon) + \sum_{k=0,k\neq l}^{N-1} X(k)\, S(l-k-\varepsilon) \qquad k = 0,1,\ldots,N-1 \tag{3.11}$$

Where $N$ is the total number of OFDM subcarriers, $X(k)$ is the modulated subcarrier and $S(l\text{-}k-\varepsilon)$ are the complex coefficients for the ICI components in the received signal. These coefficients are given by

$$S(l-k-\varepsilon) = \frac{\sin\,(\pi(l-k-\varepsilon))}{N sin(\frac{\pi(l-k-\varepsilon)}{N})}\; e^{(j(1-\frac{1}{N})\pi(l-k-\varepsilon))}$$

where $\varepsilon$ represents the normalized frequency offset.

The first term in the right hand side of (3.11) is the desired part of the received signal, and the second term represents the sum of interferences resulting from other subcarriers.

### 3.2.2 Equalization techniques for ICI reduction

In this section a brief review of some of the equalization techniques are explored for the comparison of time domain equalization technique which will be presented in Chapter four.

There are different methods used to reduce ICI available in the literature [32].
One of the approaches is statistically estimating the frequency offset and canceling this offset at the receiver. In this approach, the frequency-offset estimation is generally performed in two steps: coarse frequency-offset estimation, which estimates the part of frequency offset that is a multiple of the subcarrier spacing, and fine frequency-offset estimation, which estimates the remaining part of the offset that is smaller than half the subcarrier spacing. For example in reference [33] MMSE algorithm for frequency domain equalization is considered.

Another approach used to mitigate ICI is to use signal processing or coding to reduce the sensitivity of the OFDM system to the frequency offset. These methods can be used as low complexity alternatives to fine frequency-offset estimation. Some of these techniques are:

- ICI self cancellation
- Frequency domain equalization
- Time domain windowing

Self cancellation and frequency domain correlative coding are briefly presented in this section while time domain windowing equalization is investigated in Chapter four.

**ICI SELF CANCELLATION**

Self-cancellation method, also called Polynomial Cancellation Coding (PCC), is studied most among other ICI reduction methods. The method is investigated by different authors in [14, 15, 16, 34].

The main idea in self-cancellation is to modulate one data symbol onto a group of L subcarriers with predefined weighting coefficients to minimize the average carrier to interference ratio (CIR). Most papers consider a self-cancellation technique in which one data symbol modulates two subcarriers for bandwidth efficiency. Thus, in this thesis, we consider this relatively bandwidth efficient self-cancellation technique as a technique which the performance of time domain windowing is to be compared with.

Let L=2, thus its coefficients are (l,-l), which implies that modulating one data symbol onto two adjacent subcarriers with 180° phase different. At the receiver side, the signals received within a group are linearly combined with the same weighting coefficients.

Let $X(k)$ and $X_{sc}(k)$ be frequency domain symbols of the k^th subcarrier before and after the implementation of self-cancellation coding respectively. The transmitted data symbols for a given OFDM block can thus be represented as:

$$X_{sc}(k) = X(k) \; and \; X_{sc}(k+1) = -X(k) \; for \; k = 0, 2, \ldots, N-2. \qquad (3.14)$$

The ICI now depends on the difference between the adjacent ICI coefficients, $S(l-k) - S(k-l+1)$, rather than on the coefficients themselves. As the difference between adjacent coefficients is small, this results in substantial reduction in ICI. If adjacent coefficients were equal, then the ICI would be completely cancelled. Thus this process can be considered as cancelling out the component of ICI, which is constant, between adjacent pairs of coefficients. ICI cancellation depends only on the coefficients being slowly varying functions of the offset. It does depend not on the absolute values of the coefficients and so improves the performance for any frequency offset.

By mapping data onto larger groups of subcarriers, higher order ICI cancellation can be achieved. For the general case of mapping onto groups of subcarriers, the relative weightings of the subcarriers in the group are given by appropriate coefficients.

This scheme is very easy to implement without increasing the system complexity. The drawback of the scheme is that the reduction in bandwidth efficiency.

**FREQUENCY DOMAIN EQUALIZATION:**

Frequency domain correlative coding techniques for intercarrier interference reduction techniques are studied by many researchers as in [6, 20, 21, 35, 36]. The Doppler spread in the channel causes ICI in the OFDM demodulator. The pattern of ICI varies from frame to frame for the demodulated data but remains invariant for all symbols within a demodulated data frame. Compensation for fading distortion in the time domain introduces the problem of noise enhancement. So frequency domain equalization process is prefered for reduction of ICI by using suitable equalization techniques. We can estimate the ICI for each frame by inserting frequency domain pilot symbols in each frame as shown below.

The correlative coding between signals modulated on subsequent subcarriers is used to compress ICI in OFDM systems. Frequency-domain correlative coding is a simple solution to ICI problems, and makes OFDM systems less sensitive to frequency errors. In addition, system bandwidth efficiency will be not reduced by introducing correlative coding into the system.

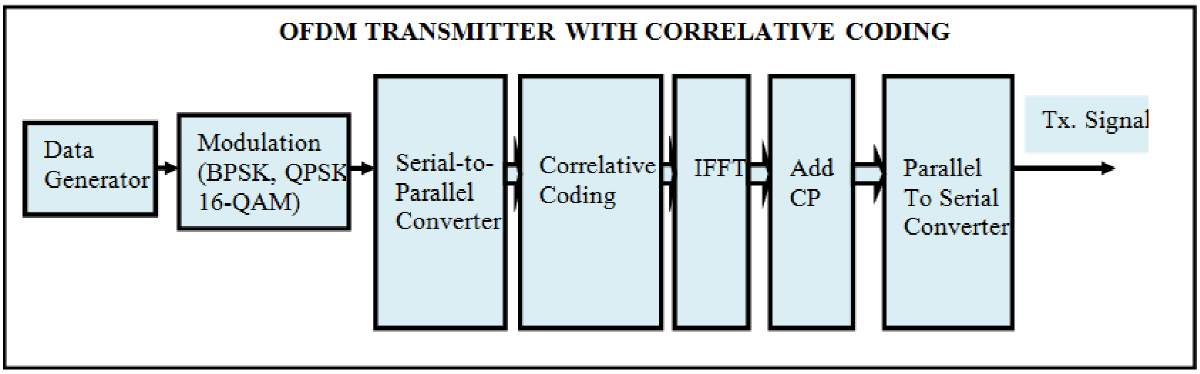



Figure 3.1-Correlative coding Transmitter model

The major source of ICI is due to the frequency mismatch between the transmitter and receiver, and the Doppler shift. The above method cannot address this. Again it is only suitable for flat fading channels, but in mobile communication the channels are frequency selective in nature because of multipath components. Here also the channel needs to be estimated for every frame. Estimation of channel is complex, expensive & time consuming. Hence the method is not efficient.

The block diagram of an OFDM system using correlative coding is shown in Figure 3.1. The input signal sequence $X(k)$, where $k$ is the subcarriers' index with $k = 0, 1, \dots, N-1$, takes the values -1, +1, that fulfill the zero mean and independence conditions. Using a coding correlation polynomial $F(D) = 1 - D$, where $D$ denotes the unit delay of the subcarrier index $k$, the coded symbols are expressed as:

$$X_{CC}(k) = X(k) - X(k-1), \quad for\ k = 0, 1, \dots, N-1 \tag{3.15}$$

The coded symbols modulate $N$ orthogonal subcarriers. The symbol $X_{CC}(k)$ takes three possible values {-2,0,2}. Thus equation (3.15) introduced a correlation between successive symbols ($X_{CC}(k), X_{CC}(k-1)$), and the independence condition is not maintained. Precoding is performed before the BPSK modulation to avoid the error propagation in the decoding process due to correlative coding.

If we consider other modified correlative coding techniques such as 1-D, $1\text{-}2D+D^2$, $1\text{-}D\text{-}D^2$ etc. we will have different output symbols at the receiver. Assuming BPSK modulation Table 3.1 shows the correlative coding techniques and the corresponding output symbols.

| Correlative coding | Output symbols | Number of symbols (M) |
|---|---|---|
| 1-D | -2, 0, 2 | 3 |
| $1\text{-}D\text{-}D^2$ | -3,-1,1,3 | 4 |
| $1\text{-}2D+D^2$ | -4, -2, 0, 2, 4 | 5 |

Table 3.1- common types of correlative coding techniques

**Time domain windowing**

Basically windowing is the process of multiplying the transmitted signal wave form by a suitable function. Windowing the signal makes the spectrum of the signal waveform more concentrated in order to diminish the interference. Common types of pulse shaping windows such as Hamming, Hanning, Raised cosine, Kaiser, etc. are usually used for pulse shaping purposes. References [37] and [38] show the performance and comparison of these pulse shaping window functions.

Even though they are not effective, the time-domain pulse shaping method can also reduce ICI through multiplying the transmitted time-domain signals by a well-designed pulse shaping function [39]. If we can reduce the side lobe significantly then the ICI power will also be reduced significantly. Hence a number of pulse shaping functions are proposed aiming to reduce the side lobe as much as possible.

**A** dual-window technique to reduce the sensitivity to carrier frequency offset (CFO) is studied in reference [8]. In this paper, two windows are defined and used alternately onto adjacent subcarriers for pulse shaping at the transmitter. The receiver selects one of the two windows to maximize the output of the desired subcarriers and suppresses the others by window- matching and anti-matching functions respectively.

A time-domain windowing of OFDM signal to reduce the sensitively to carrier frequency offset was also described in [17], where the Nyquist window was used to suppress the side lobes of subcarriers.

This thesis investigates a specialized time domain window, discussed in Chapter four, which is derived from the correlative coding principle in frequency domain. This time domain window is used to reduce the sensitivity to linear distortions and to reduce the sensitivity to frequency errors (ICI). In this time domain equalization technique, the time domain signals are multiplied by a well designed window function. The performance in terms of BER is compared to the existing frequency domain and ICI self cancellation techniques.

This type of window is first proposed by Kumar, Malarvizhi and Jayashri in reference [6]. They proposed a time domain equalization technique based on the window function which creates a correlation between two adjacent subcarriers. Their paper is limited to a simple window and does not show the performance compared to other techniques. We investigate the performance of different window orders and compare it with other equalization techniques such as self cancellation and frequency domain correlative coding.

# CHAPTER FOUR

## TIME DOMAIN EQUALIZATION WITH WINDOWING

In this chapter, we discuss the general and specific forms of the proposed time domain window, the carrier to interference ratio, effects of the parameters of the window and appropriate demodulation techniques.

The application of the windowing function tapers the start and ends of waveform reducing the transients and consequently the spectral spreading.

In time domain equalization technique, the time domain signals are multiplied by a well designed window function. The time domain window function, considered in this research, is defined to be equivalent to the correlation polynomial used in the frequency domain equalization technique. It reduces the sensitivity to frequency offset by creating the correlation between the neighborhood subcarriers.

### 4.1 Proposed window function and system model

The time-domain windowing function used has a general form $(\mathbf{1} - \boldsymbol{a} * \boldsymbol{e}^{\boldsymbol{j2\pi n/N}})^{\boldsymbol{L}}$**.** The coefficient $a$ and the correlation order L are the parameters of the window function; N is the number of data samples (equal to FFT size or number of subcarriers) and n = 0, 1, …, N-1. This corresponds to the frequency-domain correlative coding process of the form $(1 - a * D)^L$, where $D$ is the unit delay and $L$ is the order of the windowing. For example, if *(1-D)* correlative coding is adopted, the transmitted data on the k$^{th}$ subcarrier of the given OFDM block can be expressed as $X_w(k) = X(k) - X(k-1)$where the subcarrier indices *(k) and (k-1)* in *X (k) and X (k-1)* are in mod $N$ operations, where *X (k) and X (k-1)* are the $k^{th}$ and $(k\text{-}1)^{th}$ subcarriers frequency domain symbols, and $X_w$ *(k)* is the frequency domain signal of the $k^{th}$ subcarrier after the application of window.

It is a modified ICI suppression scheme based on correlative coding, which is more suitable for the block-wise modulation property of OFDM systems. The window function is reinitialized for each OFDM symbol or block and operates circularly on each block of data samples. The window function is implemented at the transmitter directly after the IFFT block for system complexity reduction as shown in Figure 4.1, the general system model.

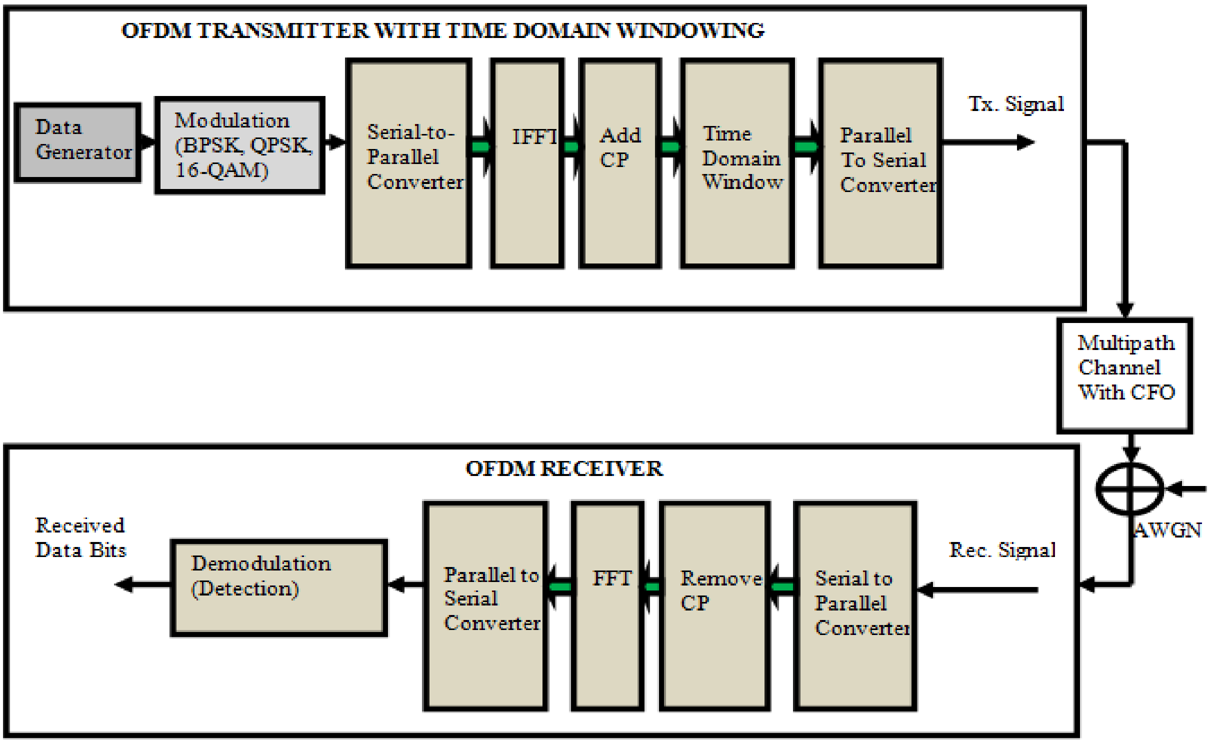



Figure 4.1-Time domain equalization with windowing system model

This arrangement ensures that each pair of adjacent subcarriers within an OFDM block will contribute to a self-canceled ICI term, unlike the non-blocked correlative coding scheme where the head and tail data samples within an OFDM block do not form self-canceled pairs of ICI.

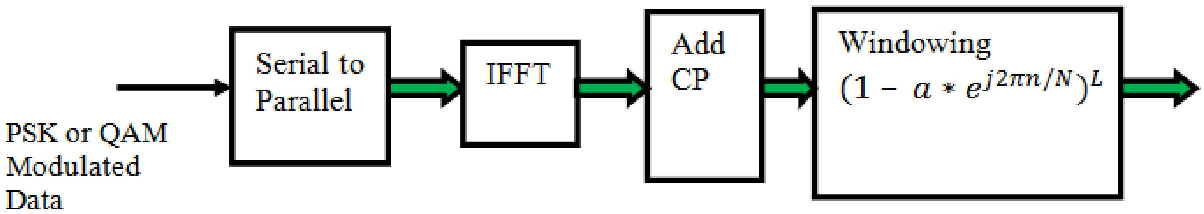



Figure 4.2- general form of window function in an OFDM transmitter

The relationship of this type of time domain window with the correlative coding scheme indicated in Chapter three is illustrated here through some mathematical analysis.

Consider the general form of the window function of the form:

$$w(n) = (1 - \alpha * e^{j2\pi n/N})^L \tag{4.1}$$

The time domain OFDM signal, x(n), is multiplied by the window function w(n) and fed into the channel.

$$x(n)w(n) = x(n)(1 - \alpha * e^{j2\pi n/N})^L \tag{4.2}$$

To visualize the frequency domain effect of the window at the receiver, assume an ideal channel condition, and then the DFT of the time domain signal in Equation (4.2) gives the receiver signal in frequency domain of a give subcarrier k, $Y(k)$.

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x(n)(1 - \alpha * e^{j2\pi n/N})^L e^{-j2\pi \frac{k}{N} n} \tag{4.3}$$

But,

$$(1 - \alpha * e^{j2\pi n/N})^L = (1 - \alpha * L * e^{\frac{j2\pi n}{N}} + \alpha^2 * \frac{L(L-1)}{2} * e^{\frac{j4\pi n}{N}} - \dots + (-\alpha)^L e^{\frac{j2\pi nL}{N}}) \tag{4.4}$$

Substitution of Equation (4.4) in to (4.3),

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x(n)(1 - \alpha * L * e^{\frac{j2\pi n}{N}} + \alpha^2 * \frac{L(L-1)}{2} * e^{\frac{j4\pi n}{N}} - \dots + (-\alpha)^L e^{\frac{j2\pi nL}{N}}) e^{-j2\pi \frac{k}{N} n}$$

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \Big\{ x(n) * e^{-j2\pi \frac{k}{N} n} - \alpha * L * x(n) * e^{\frac{-j2\pi n(k-1)}{N}} + \alpha^2 * \frac{L(L-1)}{2} * x(n) * e^{\frac{-j2\pi n(k-2)}{N}} - \dots + (-\alpha)^L * x(n) * e^{\frac{-j2\pi n(k-1)}{N}} \Big\} \tag{4.5}$$

The DFT implementation of each term in Equation (4.5) results in the following frequency domain expressions.

$$Y(k) = X(k) - \alpha * L * X(k-1) + a^2 * \frac{L(L-1)}{2} * X(k-2) - \dots + (-\alpha)^L X(k-L). \tag{4.6}$$

If L=1 ➔ $Y(k) = X(k) - \alpha * X(k-1)$

If L=2 ➔ $Y(k) = X(k) - 2 * \alpha * X(k-1) + \alpha * X(k-2)$ (4.7)

If L=3 ➔ $Y(k) = X(k) - 3 * \alpha * X(k-1) + 3 * \alpha * X(k-2) - \alpha * X(k-3)$

Equation (4.6) clearly shows that the time domain windowing of the form of Equation (4.1) is the correlative coding in frequency domain. Thus, the effect of the preceding symbol should be considered in the detection of the present symbol. This reduces the interference due to other carriers on the desired carrier.

**Example:** Consider the time domain windowing with leading coefficient $\alpha = 1 \; and \; correlation \; order \; L = 1$

$$w(n) = 1 - e^{j2\pi\frac{n}{N}}$$

Then, the signal after time domain windowing is:

$$x(n)w(n) = x(n)(1 - e^{j2\pi\frac{n}{N}})$$

The DFT of the time domain windowed signal at the receiver is given by

$$Y(k) = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} x(n)w(n)e^{-j2\pi\frac{k}{N}n}$$

$$Y(k) = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} x(n)\left(1 - e^{j2\pi\frac{n}{N}}\right)e^{-j2\pi\frac{k}{N}n}$$

$$Y(k) = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} x(n)e^{-j2\pi\frac{k}{N}n} - \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} x(n)\, e^{-j2\pi\frac{n}{N}(k-1)}$$

$$Y(k) = X(k) - X(k-1)$$

It can also be obtained from Equation (4.7) by substituting the leading coefficient $\alpha = 1$.

The effect of this typical time domain window after the FFT implementation is illustrated diagrammatically in Figure 4.3.

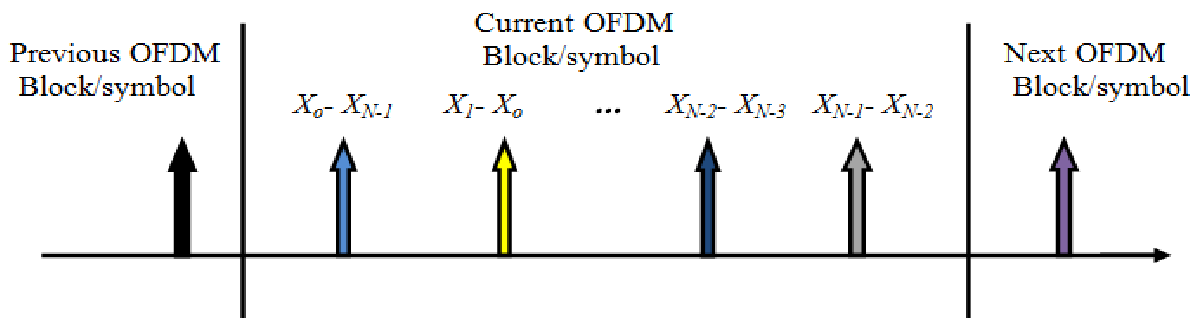



Figure 4.3-Frequency domain realization of windowing on OFDM receiver symbols

The frequency domain graphical representations of the corresponding window functions indicate that correlation between the desired subcarrier symbol and the adjacent subcarrier symbols has been created.

To clearly visualize the window, let's have graphical representations of some of the window functions in time domain. The frequency domain representations (after the FFT implementation of the time domain window functions) some of the two typical windows are also shown in Section 4.3.

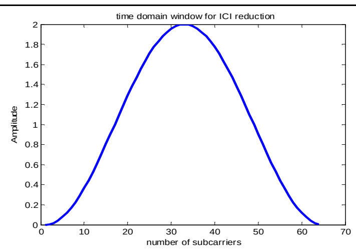
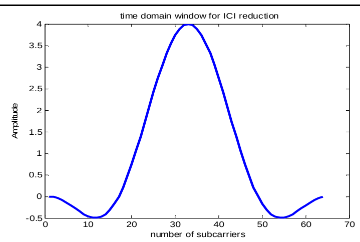
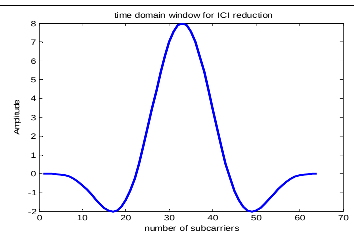
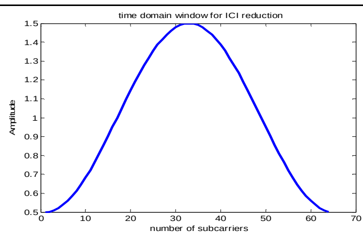
| Parameters | Window function | Time domain plot |
|---|---|---|
| $\alpha = 1, L = 1$ | $w(n) = 1 - e^{j2\pi n/N}$ |   |
| $\alpha = 1, L = 2$ | $w(n) = (1 - e^{j2\pi n/N})^2$ |   |
| $\alpha = 1, L = 3$ | $w(n) = (1 - e^{j2\pi n/N})^3$ |   |
| $\alpha = 0.5, L = 1$ | $w(n) = 1 - 0.5 * e^{j2\pi n/N}$ |   |

Table 4.1-typical window functions

The leading coefficient $\alpha$ is not a significant parameter but it has the effect on the demodulation complexity at the receiver by increasing the number of output symbols. Table 4.2 shows the impact of non integral values of $\alpha$ on the receiver symbols considering BPSK modulation. It produces a number of receiver symbols with non integral energy values which may complicate the decision functions of the receiver. The next section also shows $\alpha = 1$ is the optimum value from the analysis of carrier to interference ratio.

| Parameters | Window function | Output symbols | No. of symbols |
|---|---|---|---|
| $\alpha = 1, L = 1$ | $1 - e^{j2\pi n/N}$ | -2, 0, 2 | 3 |
| $\alpha = 0.5, L = 1$ | $1 - 0.5 * e^{j2\pi n/N}$ | -1.5, -0.5, 0.5, 1.5 | 4 |
| $\alpha = 1, L = 2$ | $(1 - e^{j2\pi n/N})^2$ | -4, -2, 2, 4 | 4 |
| $\alpha = 0.5, L = 2$ | $(1 - 0.5 * e^{j2\pi n/N})^2$ | -2.25, -1.75, -0.25, 0.25, 1.75, 2.25 | 6 |

Table 4.2- Effect of leading coefficient on receiver output symbols

### 4.2 Carrier to Interference Ratio analyses

Next let's describe how to optimize the parameters $\alpha$ and $L$ based on maximizing the CIR function. When the windowing function $(1 - \alpha * e^{j2\pi n/N})^L$ is applied, the data sample transmitted on the k$^{th}$ subcarrier can be expressed as (after FFT implementation at the receiver)

$$X_w(k) = X(k) - \alpha * L * X(k-1) + \cdots + (-\alpha)^L X(k-L) \quad (4.8)$$

$$= \sum_{l=0}^{L} C_l^L (-\alpha)^l X(k-l)$$

Where, $C_l^L = \frac{L!}{l!(L-l)!}$

From the ICI analysis in Chapter three, let's consider the impairments due to carrier frequency offset and optimize the parameters on the carrier to interference ratio of the windowed signal.

The received data sample at the l$^{th}$ subcarrier can be written as

$$Y_l = \sum_{k=0}^{N-1} X_w(k) \frac{\sin \pi(l-k-\varepsilon)}{N sin\left(\frac{\pi(l-k-\varepsilon)}{N}\right)} . \mathrm{e}^{\left(\frac{\mathrm{j}\pi(l-k-\varepsilon)(1-\mathrm{N})}{N}\right)} + n_l, l = 0,1, \dots, \mathrm{N}-1 \quad (4.9)$$

where, $X_w(k)$ is the windowed frequency domain signal.

This can be alternatively expressed as

$$Y_l = C_l + I_l + n_l \quad (4.10)$$

where, $C_l = X_w(l)S(-\varepsilon)\ and\ I_l = \sum_{k=0,k\neq l}^{N-1} X_w(k)S(l-k-\varepsilon)$

The function $S(l-k-\varepsilon)$ is regarded as the ICI coefficient which reflects the interfering magnitude from the k$^{th}$ subcarrier to the l$^{th}$ subcarrier and expressed as:

$$S(l-k-\varepsilon) = \frac{\sin \pi(l-k-\varepsilon)}{N sin\left(\frac{\pi(l-k-\varepsilon)}{N}\right)} . \mathrm{e}^{\left(\frac{\mathrm{j}\pi(l-k-\varepsilon)(1-\mathrm{N})}{N}\right)}$$

The first term in the right hand side of (4.10) is the desired part of the received signal, the second term represents the sum of interferences resulted from other subcarriers, and the third term represents the sample of a zero-mean Gaussian distributed random variable.

Thus, the demodulated data sample for the l$^{th}$ subcarrier can thus be of the form

$$Y_l = X_w(l)S(-\varepsilon) + \sum_{k=0,k\neq l}^{N-1} X_w(k)S(l-k-\varepsilon) + n_l, l = 0,1, \dots, \mathrm{N}-1 \quad (4.11)$$

where, N is the total number of subcarriers, $X_w(l)$ denotes the transmitted data sample on the l$^{th}$ subcarrier, and ε represents the normalized frequency offset with respect to the frequency spacing among subcarriers.

For the windowing parameters optimization, let’s investigate the theoretical CIR after applying the general $(1 - \alpha * e^{j2\pi n/N})^L$ windowing. The carrier to interference signal power ratio from equation (4.10) is

$$CIR = \frac{E\{|C_l|^2\}}{E\{|I_l|^2\}} = \frac{E\{|X_w(l)|^2|S(-\varepsilon)|^2\}}{E\left\{\sum_{k=0,k\neq l}^{N-1}\sum_{p=0,p\neq l}^{N-1} X_w(k)X_w(p)^* S(l-k-\varepsilon)S(l-p-\varepsilon)^*\right\}} \quad (4.12)$$

Assume that the set of data $\{X(k), k = 0,1, \dots, N-1\}$ are zero mean i.i.d. random variables. Then we can obtain

$$E\{X_w(k)X_w(p)^*\} = E\{\sum_{l=0}^{L} C_l^L(-\alpha)^l X(k-l)(\sum_{m=0}^{L} C_m^L(-\alpha)^m X(p-m))^*\} \tag{4.13}$$

Since we have $E\{X(k)X(p)^*\} = 0 \; for \; k \neq p \; and \; E\{X(k)X(p)^*\} = 1 \; for \; k = p$, Equation (4.13) can be expressed as:

$$E\{X_w(k)X_w(p)^*\} = \begin{Bmatrix} (C_0^L(-\alpha)^0)^2 + \cdots + (C_0^L(-\alpha)^L)^2, k = p \\ \sum_{j=0}^{L-j} C_j^L(-\alpha)^j C_{j+i}^L(-\alpha)^{j+i}, k = [p \pm i]_N, 1 \leq i \leq L \end{Bmatrix} \tag{4.14}$$

By substituting Equation (4.14) into Equation (4.12) and after going through some mathematical simplifications the theoretical CIR in Equation (4.12) can be expressed as;

$$CIR = \frac{|S(-\varepsilon)|^2}{\sum_{k=1}^{N-1}|S(k-\varepsilon)|^2 + \sum_{i=1}^{L}\sum_{\substack{k=1 \\ k\neq i}}^{N-1} d_i Re\{2S(k-\varepsilon)^* S(k-\varepsilon-i)\}} \tag{4.15}$$

Where, $d_i = \frac{\sum_{j=0}^{L-i} C_j^L(-\alpha)^j C_{j+i}^L(-\alpha)^{j+i}}{\sum_{l=0}^{L}(C_j^L(-\alpha)^j)^2}$

To optimize the windowing function, we can maximize the CIR function in (4.15), or minimize its denominator. From the numerical analysis of the first-order derivative of the denominator of (4.15) with respect to $\alpha$, we find that $\alpha$ =1 is not only one of the roots, but also happens to result in the global minimum of CIR if we substitute it into (4.15) regardless of the order of windowing L. This also agrees with the tabular analysis in table 4.2. Such an outcome implies that the correlative pairs would on the average contribute to the smallest ICI when the weighting magnitudes of each pair are equal.

The theoretical CIR obtained by substituting $\alpha$ =1 into (4.14) results in

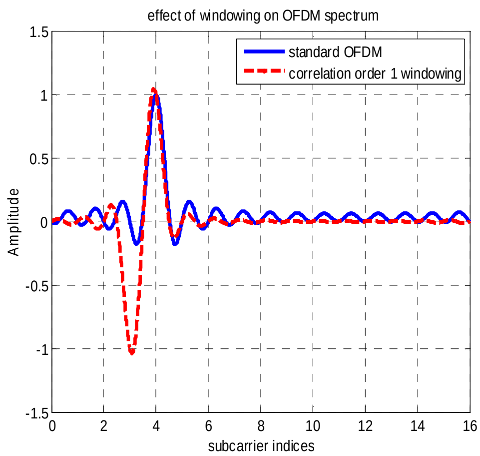
$$CIR = \frac{|S(-\varepsilon)|^2}{\sum_{k=1}^{N-1}|S(k-\varepsilon)|^2 + \sum_{i=1}^{L}\sum_{\substack{k=1 \\ k\neq i}}^{N-1} e_i Re\{2S(k-\varepsilon)^* S(k-\varepsilon-i)\}}$$

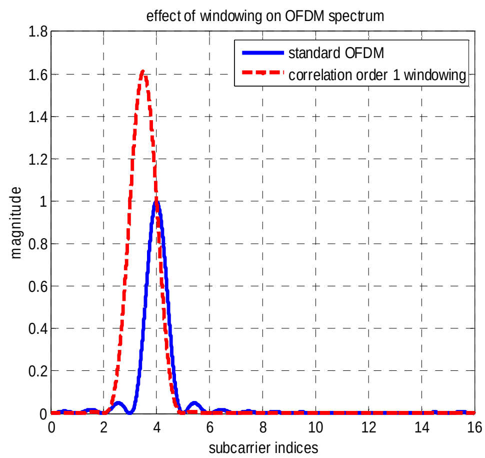
Where, $e_i = \binom{2L}{L-i}(-1)^i$

Having an optimum value of the leading coefficient, in the following discussions we will consider only the form of $(1 - e^{j2\pi n/N})^L$ for the time-domain windowing function.

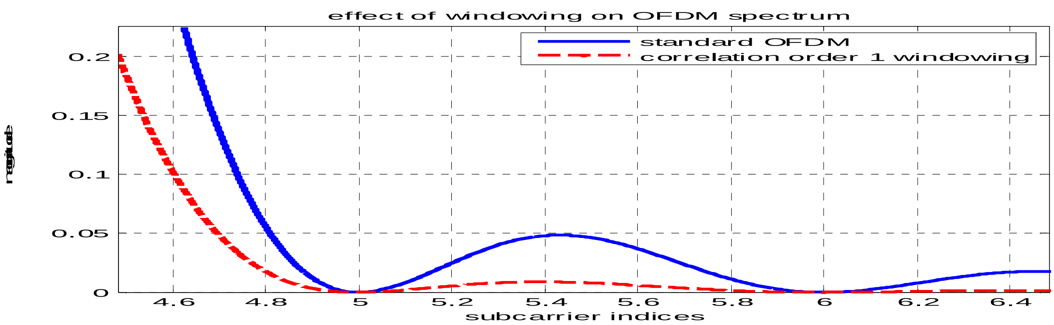
### 4.3 Effect of the window on subcarrier spectrum outband OFDM power spectrum

Here the subcarrier spectrum of the proposed equalization scheme for different orders of windowing is examined. As shown in Figure 4.4 and figure 4.5, when windowing is applied, the proposed approach produces a stronger main lobe and smaller side lobes in each subcarrier spectrum and this implies that smaller ICI would be introduced before the demodulation process.



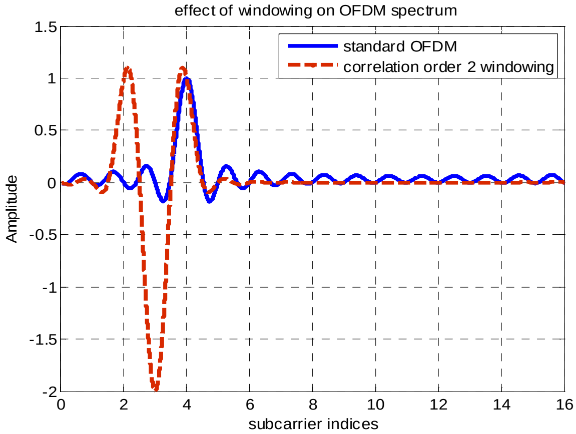

(a) Subcarrier spectrum at the 4th subcarriers (b) subcarrier power at the 4th subcarrier

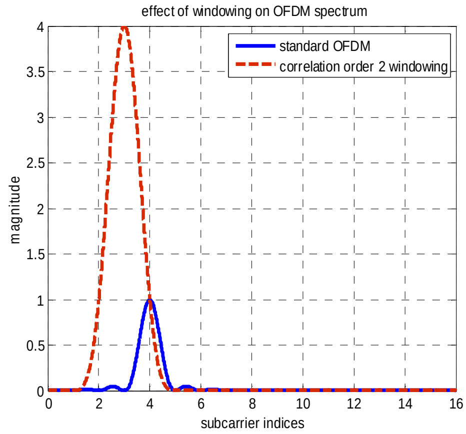



(c) Zoom in on side lobes of subcarrier power in (b)

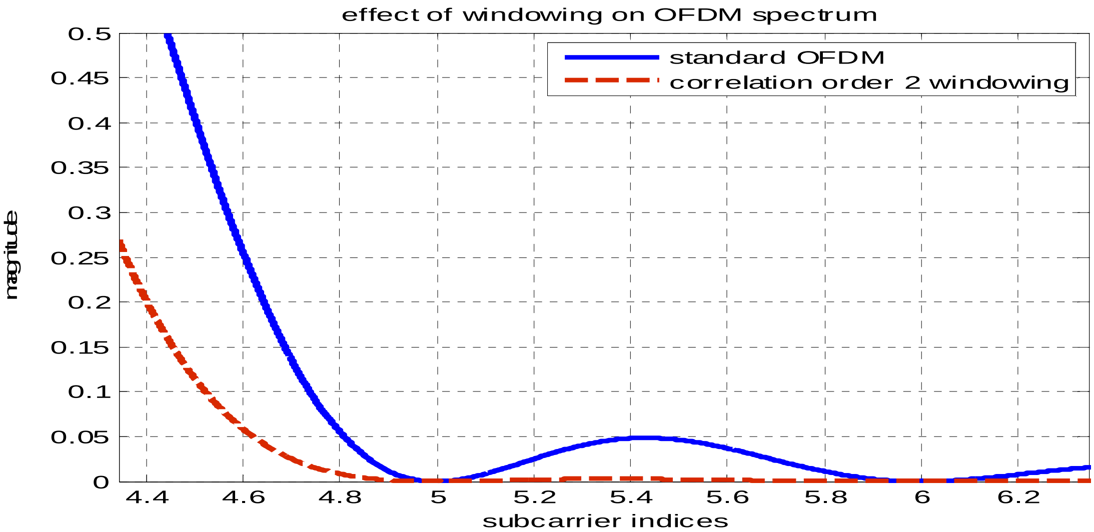
Figure 4.4- Effect of windowing of L=1 on subcarrier frequency spectrum

In figures 4.4 and 4.5 the spectrum at the 4th subcarrier in the band of 16 subcarriers is shown. The application of the proposed window introduces significant lobes to the adjacent L subcarriers and highly reduced side lobes for the other subcarriers. This produces an increased subcarrier power whose center is influenced by the adjacent subcarriers. The significant effect of the adjacent L subcarriers on the desired subcarrier power is under consideration by the appropriate demodulation techniques explained in chapter four.




(a) Subcarrier spectrum at the 4th subcarriers (b) subcarrier power at the 4th subcarrier

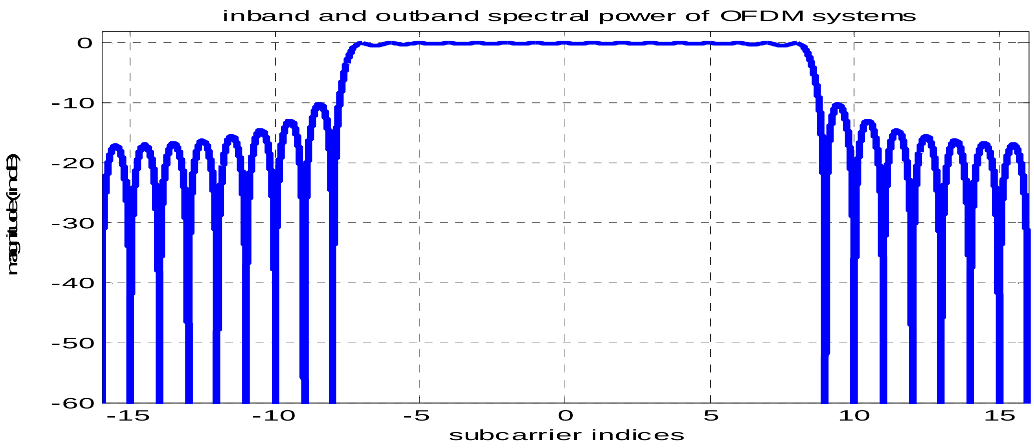



(c) Zoom in on side lobes of subcarrier power in (b)

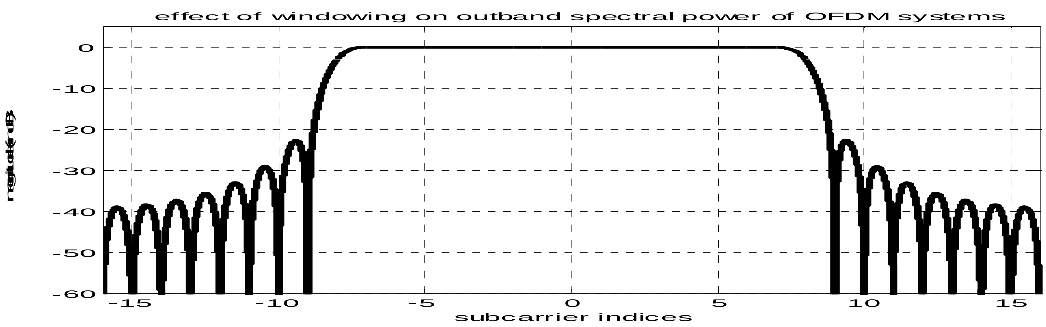
Figure 4.5- Effect of windowing of L=2 on subcarrier frequency spectrum

Although an increased window order produces stronger main lobe and highly reduced side lobes, a bit higher demodulation complexity would be required for the case using a higher order of windowing.

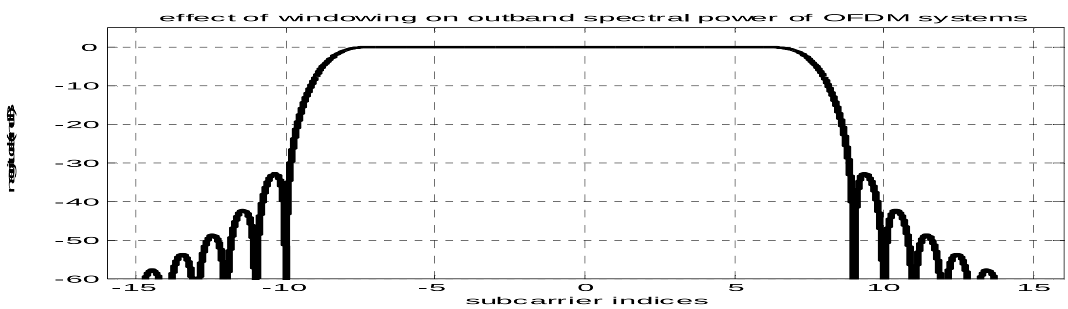
Figure 4.6 show that the power spectrum of the overall OFDM signal without any equalization and after the application of time domain windows of correlation orders 1 and 2. It can be seen that Figure 4.6 (b) & (c) results in more rapid spectral rolloff at the OFDM symbol boundaries compared to the spectrum of the standard OFDM system in figure 4.6 (a). The OOB power is considerably minimized and thus it can avoid interference between adjacent broadcast channels.




(a) Inband and outband power of standard OFDM systems




(b) Inband and outband power of OFDM with windowing of L=1




(c) Inband and outband power of OFDM with windowing of L=2

Figure 4.6 - Effect of windowing on outband power spectrum

The window with correlation order 2 has a better result in reducing the outband power spectrum compared to window of correlation order 1. However the demodulation complexity increases with correlation order which may introduce errors that may result in high BER value.

Note that due to time the proposed time domain windowing, each symbol at the given subcarrier is correlated to the L adjacent subcarriers. Thus the effect of the preceding L subcarrier symbols should be considered in the detection of the present subcarrier symbol.

### 4.4 Demodulation techniques for time domain equalization using windowing

If ICI can be suppressed well by applying well designed windowing function, then the remaining task that will affect the BER performance would be decided by the modulation and demodulation technique. There are two types of demodulation techniques proposed here: one is without using any training symbol and the other is using pilot symbols.

In the first demodulation process, there is no need of training or pilot signal and is suitable for the transmission scheme with windowing and correlative coding. It is very efficient and convenient for BPSK and QPSK modulation schemes.

Let $\Theta(.)$ is the decision function of the incoming received signal at the receiver, which is dependent on the modulation technique. For convenience, we assume that the windowing function $w(n) = 1 - e^{j2\pi\frac{n}{N}}$ is applied at the transmitter. Considering BPSK modulation this demodulation technique follows the following steps.

1. Make a random guess of the first symbol *X(0)* and denote by $\bar{X}(0)$. This symbol is used to start the demodulation process of the received symbols which are correlated to the correlation order of L due to the time domain window considered.
2. Sequentially demodulate the N symbols (symbols X(1) to X(N-1) and then X(0)) of the N subcarriers under the consideration of the correlation created by the window used. For the window stated here the received symbol in frequency domain (under ideal channel condition)has the form:

   $Y(k) = X(k) - X(k-1) \Rightarrow X(k) = Y(k) + X(k-1)$

   Thus the transmitted symbol of the k$^{th}$ subcarrier is obtained based on the decision function $\Theta(Y(k) + \hat{X}(k))$.

For example for BPSK modulation, $\theta(x) = \begin{Bmatrix} 0, for\ x < 0 \\ 1, for\ x \geq 0 \end{Bmatrix}$.

Thus demodulation sequence is:

$$\hat{X}(1) = \theta(Y(1) + \bar{X}(0))$$

$$\hat{X}(2) = \theta(Y(2) + \hat{X}(1))$$

$$\hat{X}(k) = \theta(Y(k) + \hat{X}(k-1))$$

Where $Y(k)$ is the received sequence of data of the k$^{th}$ subcarier.

3. Finally continuing the process, if $\hat{X}(0) = \bar{X}(0)$, which implies the predicted and the detected symbol matches and then the demodulation process is completed; otherwise, let $\bar{X}(0) = \hat{X}(0)$ and return to step 2 above.

Figure 4.7 summarizes the above process diagrammatically.

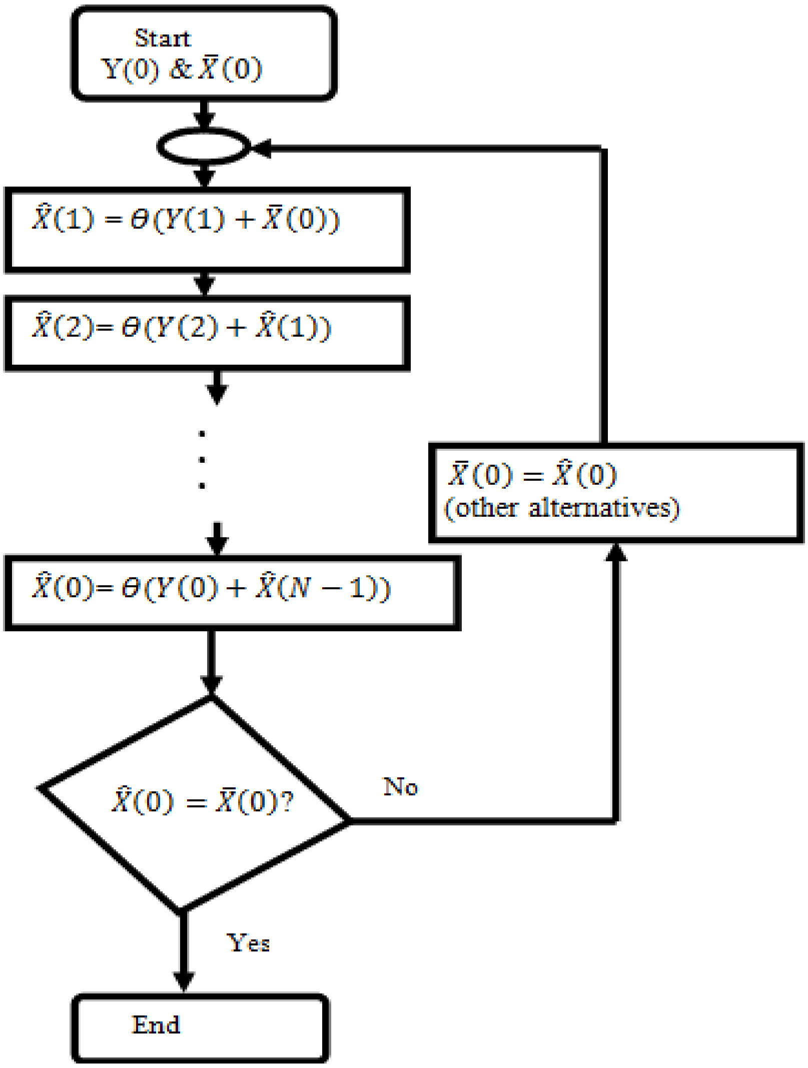



Figure 4.7- demodulation algorithm without prior knowledge of previous symbol

The computational complexity of this type of decoding however increases as the modulation order of windowing used increases.

Example:

- For BPSK modulation and order L windowing we may search up to $2^L$ cycles to get the correct initial predictions.
- For order 1 windowing and M-PSK or M-QAM modulation techniques we may have M iterations.
- For order L windowing and M-PSK or M-QAM modulation technique we may have $M*2^L$ iterations to get the correct initial predictions.

The second demodulation technique for time domain windowing used is by using the known symbols by the transmitter and the receiver for the set of subcarriers known as pilot symbols to be used at the starting point in the demodulation technique. For time domain windowing and correlative coding, the demodulation process can be initiated well by transmitting a pilot data or training sequences as the reference signal.

- For order 1 windowing the $N^{th}$ subcarriers carries the pilot symbol, denoted $X_p(N\text{-}1)$. Thus symbols from the $1^{st}$ subcarrier to the $N\text{-}1^{th}$ subcarrier are detected sequentially based on the decision function used and previously detected symbols.
- For order 2 windowing since the $1^{st}$ received symbol is circularly correlated with the $N\text{-}1^{th}$ and $N^{th}$ subcarrier symbols, the apriori knowledge of the symbols is necessary and we use the pilot symbols for this subcarriers whose symbols are known by the transmitter and receiver.
- Similarly, the $N\text{-}2^{th}$, $N\text{-}1^{th}$ and $N^{th}$ subcarriers carry the pilot symbols for order 3 time domain windowing.

The flow chart in Figure 4.8 clearly shows the aforementioned demodulation technique for the order 1 windowing. To compare and contrast with the $1^{st}$ demodulation technique described previously, we assume that the windowing function $w(n) = 1 - e^{j2\pi\frac{n}{N}}$ is applied at the transmitter and the modulation technique is assumed to be BPSK.

$\theta(.)$ is the decision function (for BPSK, $\theta(x) = \begin{Bmatrix} 0, for\ x < 0 \\ 1, for\ x \geq 0 \end{Bmatrix}$) and Y(k) and $\hat{X}(k)$ are the received OFDM symbol from the channel and the detected symbol of the k^th subcarrier respectively.

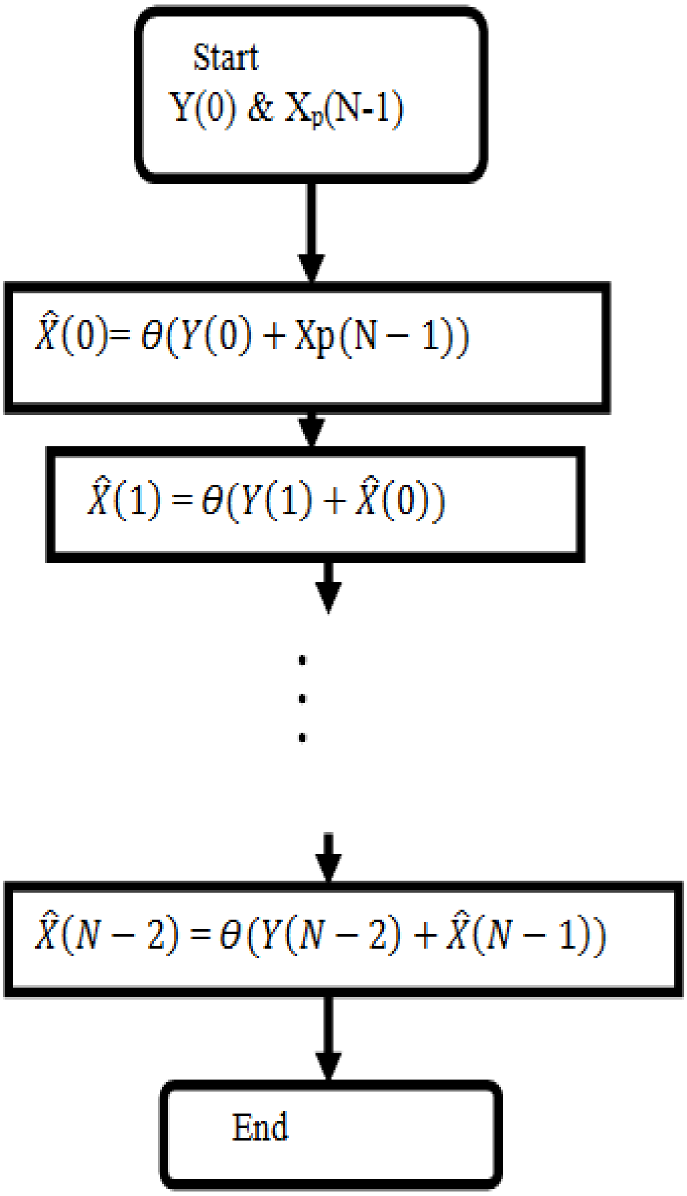



Figure 4.8- demodulation algorithm that using reference symbols

Note that the BER performance of the proposed time domain windowing scheme would be degraded due to error propagation for higher order windows. Thus there is no clear advantage to use more than order 2 windowing comparing the tradeoff between a better BER performance and demodulation complexity.

As we have seen in the above demodulation techniques the current subcarrier symbol detection depends on the previously detected symbol(s) of previous subcarrier(s). Thus error propagation may be the problem. Thus appropriate precoding has to be used at the transmitter before modulation. Precoding enables the detector to make symbol-by-symbol decisions that do not depend on previous decisions.

To conclude the theoretical analysis of the time domain window have made in this chapter. Chapter five presents different simulation results.

## CHAPTER FIVE

## SIMULATION RESULTS AND DISCUSSIONS

In this chapter different simulation results using appropriate parameters and based on the analysis in Chapter four, are presented with brief discussions on the outputs. The simulation results are plotted in terms of BER, carrier to interference power ratio and outband power analysis. We will consider QPSK modulation for the performance comparisons. This is because the symbols are at equal energy levels and the effect of carrier frequency offset is easily presented in QPSK modulation.

The standard BER that was used to determine the minimum performance of the OFDM system is minimum BER for voice transmission, that is ten to the power of minus three ($10^{-3}$). Analysis was done by observing the simulation result and tabulating the analysis results to make it more convenient to be read.

The theoretical BER in conventional OFDM systems assuming static channel conditions is illustrated in table 5.1 for reference to evaluate the simulation results [12].

| Modulation scheme | Theoretical BER | |
|---|---|---|
| | AWGN | One-path Rayleigh fading |
| BPSK | $\frac{1}{2}erfc\left(\sqrt{E_b/N_o}\right)$ | $\frac{1}{2}\left[1-\frac{1}{\sqrt{1+\frac{1}{E_b/N_o}}}\right]$ |
| QPSK | $\frac{1}{2}erfc\left(\sqrt{E_b/N_o}\right)$ | $\frac{1}{2}\left[1-\frac{1}{\sqrt{1+\frac{1}{E_b/N_o}}}\right]$ |
| 16-QAM | $\frac{3}{8}erfc\left(\sqrt{\frac{2}{5}E_b/N_o}\right)-\frac{9}{24}erfc^2\left(\sqrt{\frac{2}{5}E_b/N_o}\right)$ | $\frac{3}{8}\left[1-\frac{1}{\sqrt{1+\frac{5}{(2\,E_b/N_o)}}}\right]$ |

Table 5.1-Theoretical BERs in conventional OFDM system

We consider both the AWGN and multipath channels to justify the robustness of performance under different environments. The multipath channel is assumed to be a symbol spaced taped delay line model and characterized by a multipath intensity profile. For practical purposes, the tapped delay line channel model can be truncated at M taps as given in Proakis *et.al.* [40]. The multipath channel is assumed to have path gains following an exponential power delay profile and number of paths M=6 in the simulation results in this paper. Based on reference [41], from measurements made in different countries' urban areas delay spreads are mostly in the range of 100 nano second to 10 micro second RMS values. Thus, the delay spread of urban areas with maximum value of 10 $\mu s$ is considered here.

| **Parameter** | **Value** |
|---|---|
| Number of subcarriers | 128 |
| FFT size | 128 |
| Modulation | QPSK |
| Carrier frequency | 5 GHz |
| OFDM symbol duration | 100 µs |
| Maximum delay spread | 10 µs |
| Bit rate | 1.28 Mbits/s |
| OFDM symbols/blocks considered | 2000 |
| Cyclic prefix ratio | 1/4 |
| Channel | AWGN, multipath Rayleigh fading |
| Normalized frequency offset | 0.0:0.5 |

Table 5.2- Summary of simulation parameters

The following assumptions are made together with the parameters presented in Table 5.2.

- Perfect symbol timing synchronization at the receiver.
- Delay between consecutive paths in multipath Rayleigh fading are equal to integral multiple of sample duration.
- Cyclic prefix ratio is the ratio of cyclic prefix duration to OFDM symbol duration.
- Maximum multipath delay spread of 10 µs (urban areas) at carrier frequency of 5 GHz considered.
- Carrier frequency offset is the frequency error normalized by subcarrier frequency spacing.
- Vehicular and pedestrian frequency offsets from a speed of 3 kmph (for pedestrian) to 100 kmph (for vehicular are considered).

## 5.1 Number of carriers and cyclic prefix ratio

Figure 5.1 and Figure 5.2 show that the choice of number of subcarriers and cyclic prefix ratio is appropriate respectively.

As shown in the figure 5.1, as the number of subcarriers increases the effect of carrier frequency offset is boosted. Consider only ICI, apart from advantages for ISI reduction, the decrease in frequency separation between the subcarriers with increase in number of subcarriers results in increase in ICI. Thus for convenience we considered 128 subcarriers in a given OFDM channel for simulations to demonstrate the effect of frequency offset and its proposed solutions.

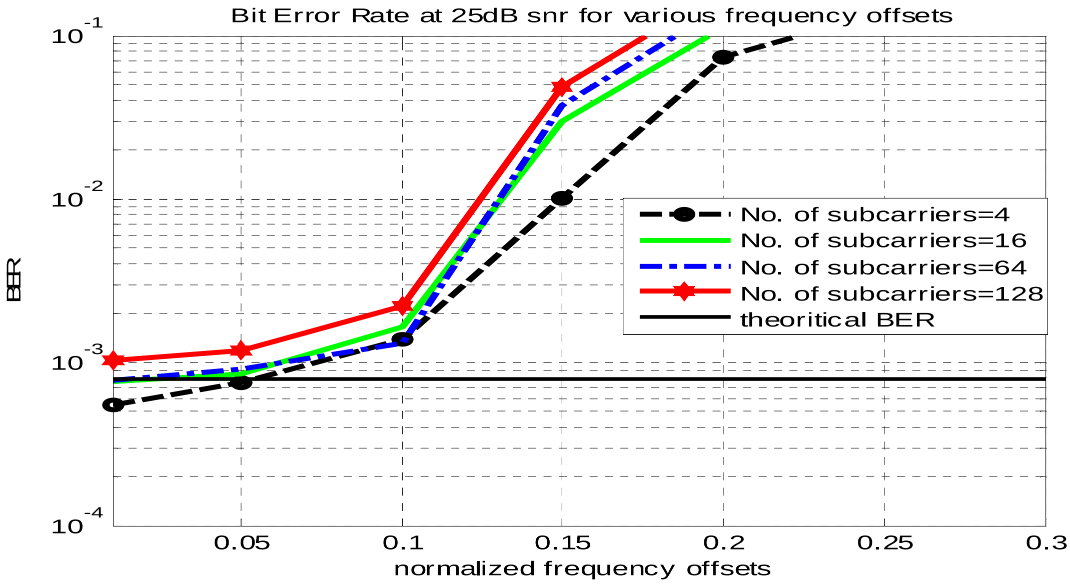



Figure 5.1- number of subcarriers sensitivity to carrier frequency offset

Cyclic prefix duration depends on the delay spread of the channel. With the specified maximum delay spread (10 us), the optimized cyclic prefix ratio is graphically illustrated in Figure 5.2 in terms of the BER performance. In the figure it can be observed that BER does not decrease values for cyclic prefix ratio less than 1/8 even for larger SNR values. This is because under the specified delay spread, BER start to increase when the length of delay spread reaches the length of effective guard period and will increase rapidly when the length of delay spread is longer than the length of guard period. It is previously stated that the effect of Inter Symbol Interference (ISI) due to multipath fading can be eliminated as long as the length of guard period is longer than the length of delay spread. Thus we have taken a cyclic prefix ratio of ¼ as an optimum value considering the delay spread of different channel conditions. Although it is better in avoiding ISI, cyclic prefix ration greater than ¼ is not effective in terms of transmission rate.

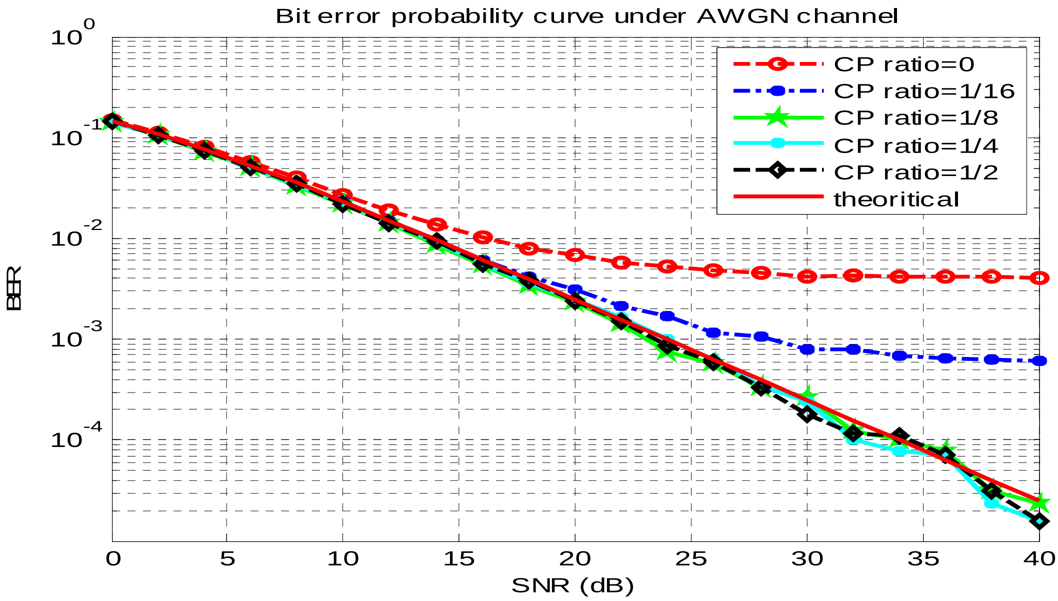
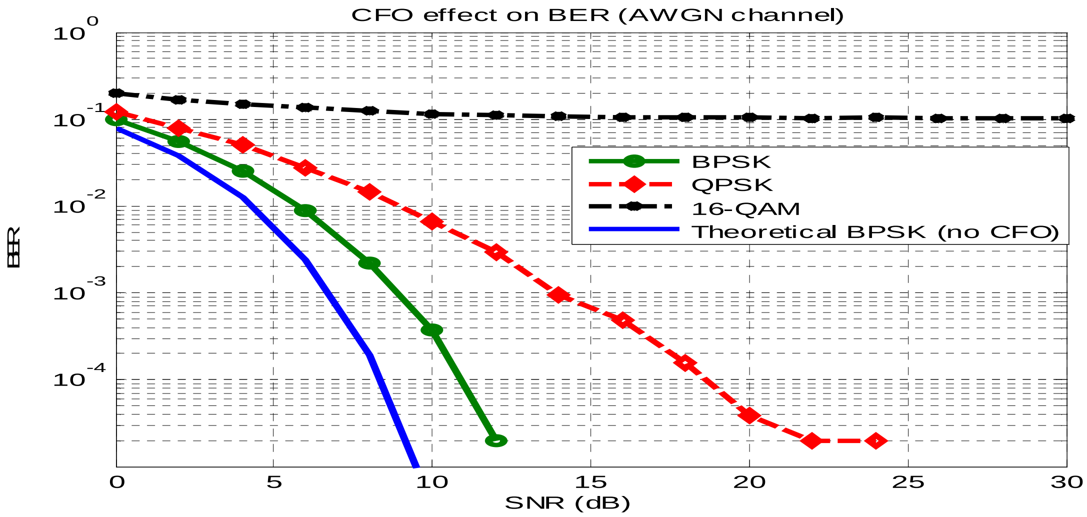



Figure 5.2- optimum cyclic prefix ratios in multipath fading channel

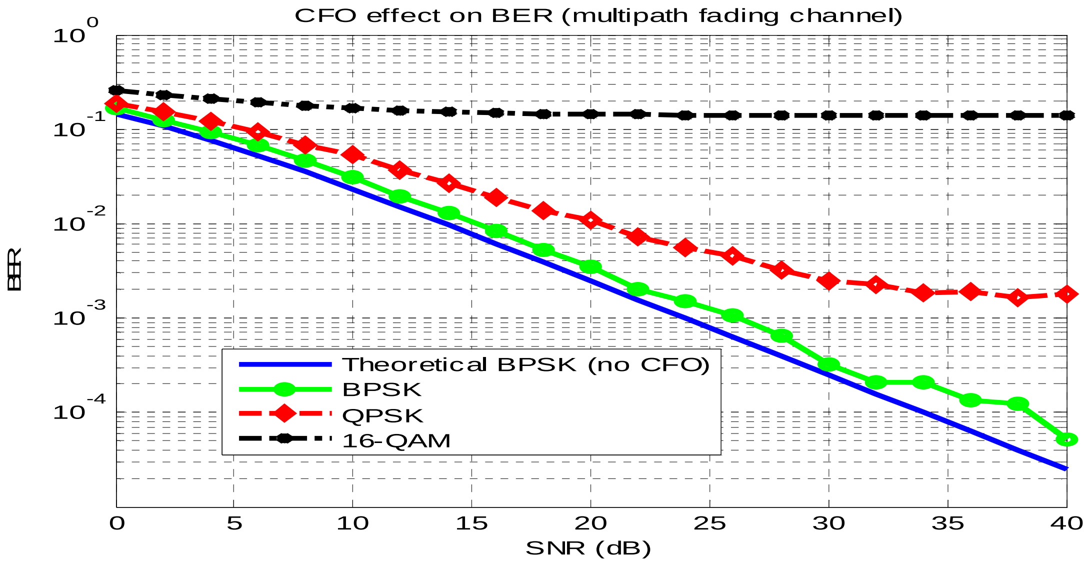
Note: theoretical indicates the theoretical BER performance of OFDM systems (from table 5.1) under static multipath channel conditions.

### 5.2 Effect of carrier frequency offset on BER performance

Before we show the effectiveness of time domain windowing in the reduction of the BER performance of OFDM systems which are affected by intercarrier interference, let's see graphically the effects of frequency offsets. Figure 5.3 shows BER effect of carrier frequency offset comparison across different modulation techniques. The effect of carrier frequency offset on OFDM system with different modulation techniques such as BPSK, QPSK and 16-QAM, for a normalized frequency offset value of $\varepsilon = 0.14$, are shown under AWGN and multipath fading channels in Figures 5.3 (a) and (b) respectively. As it is clearly seen in the figures for the 16-QAM are more vulnerable to carrier frequency offset errors than QPSK and BPSK. This is because the distance among the symbols in 16-QAM constellation is less than QPSK. Correspondingly BPSK is relatively less affected as compared to the others.




(a)




(b)

Figure 5.3- Effect of CFO on BER for $\varepsilon = 0.14$ (a) AWGN (b) multipath Rayleigh fading

The theoretical BER performance curves (from table 5.1) under both AWGN and static multipath channel conditions are plotted as a reference to show the effect of carrier frequency offset. For example, it can be observed from Figure 5.3 (a) that to achieve the BER of $10^{-3}$, the OFDM system using BPSK modulation needs an SNR of about 8 dB, the OFDM system using QPSK modulation needs at least 14 dB SNR and the OFDM system using 16-QAM modulation needs SNR of more than 30 dB.

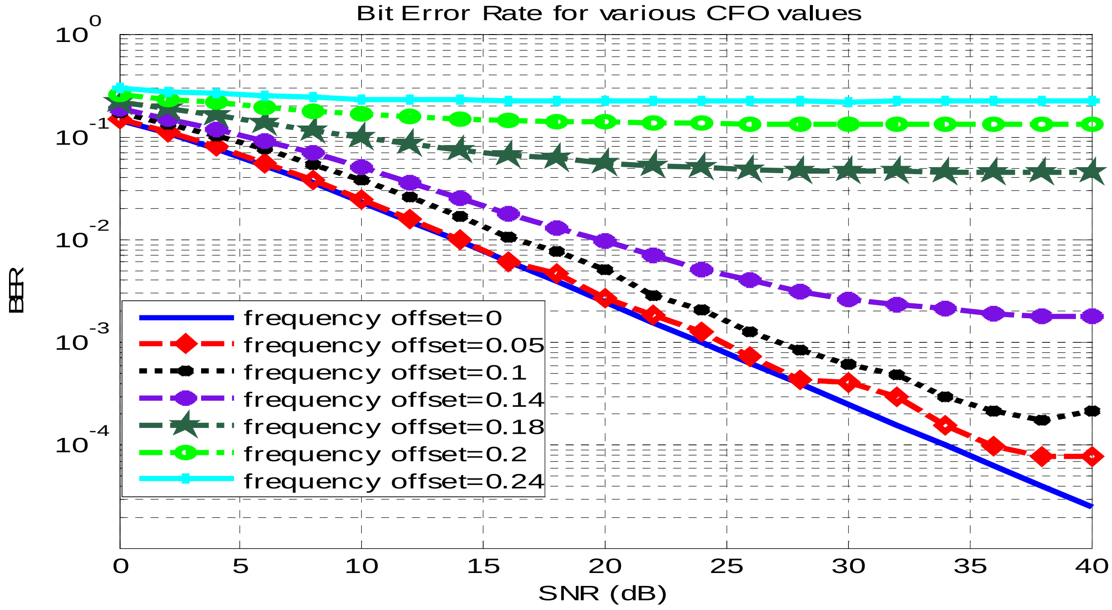
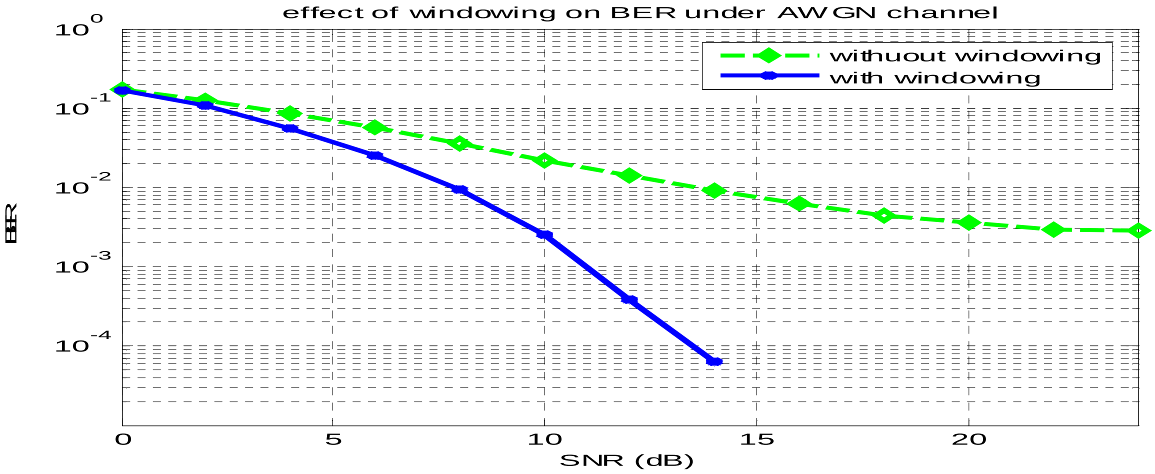



Figure 5.4- BER for various carrier frequency offset values using QPSK modulation

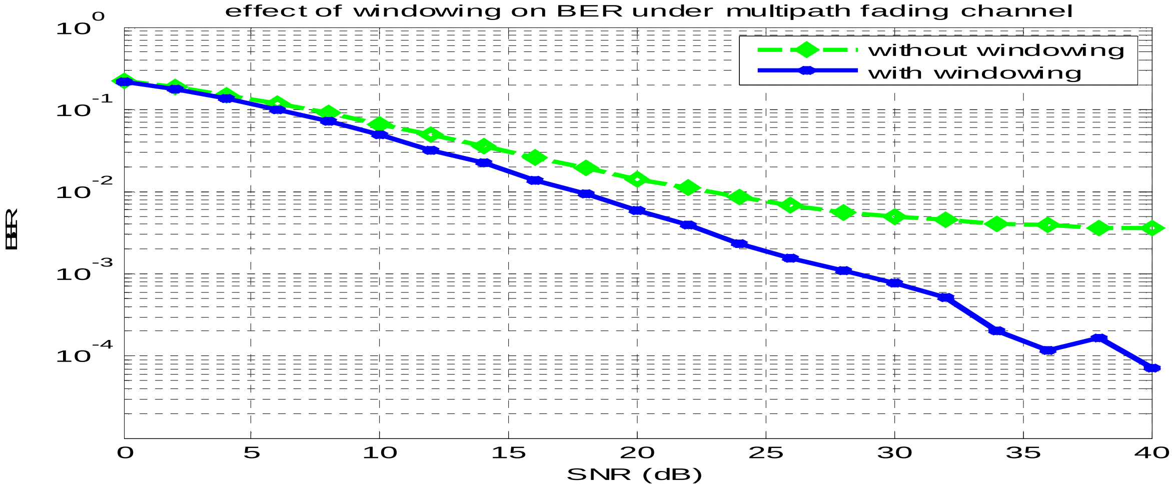
Figure 5.4 clearly shows that an increase in carrier frequency offset produces an increase in the BER of OFDM system for fading channel for the QPSK modulation case compared to the standard static channel conditions, when there is no movement between the transmitter and the receiver.

## 5.3 Effect of time domain windowing on BER performance

The performance of optimum time domain window for QPSK modulation technique which is selected for reasonable comparison is shown in Figure 5.5. Fast channel was simulated for mobile velocity of 100 km/hour (vehicular) while urban slow channel was simulated for mobile velocity of 3 km/hour (pedestrian). It can be observed from Figure 5.5 (a) that to achieve the BER of $10^{-2}$ at the frequency offset value of $\varepsilon = 0.14$, the OFDM system using windowing needs at least SNR of about 8 dB while it needs at least about 13 dB in the absence of windowing or other techniques. It provides us with an effective advantage of 5 dB in terms of SNR gain. Thus windowing in AWGN can reduce the effect due to the subcarrier's interference although it is not significant as in fading channels.




(a) Windowing effect under AWGN channel

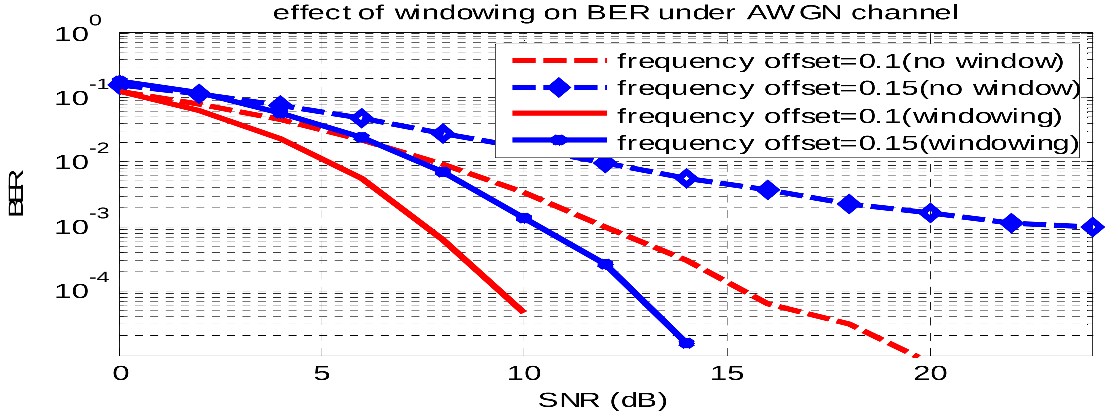



(b) Windowing effect under multipath Rayleigh fading channel

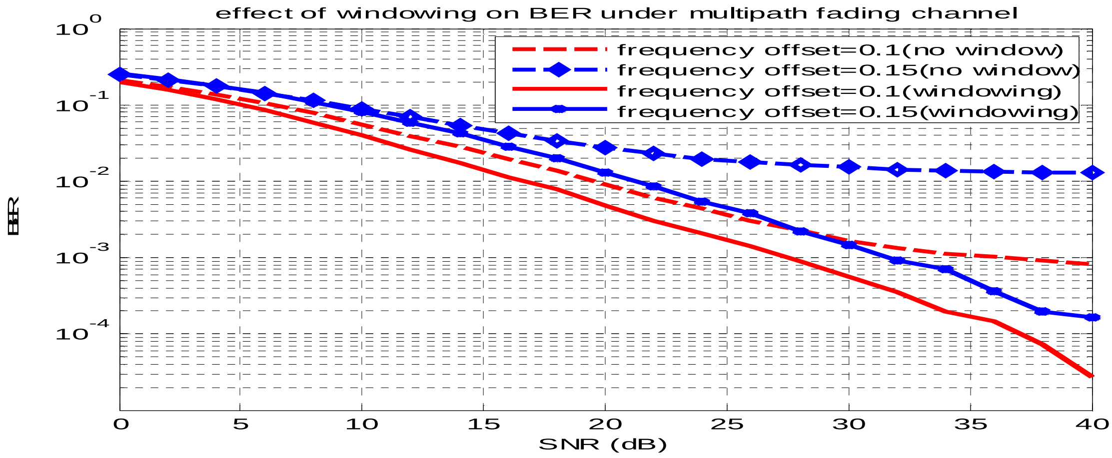
Figure 5.5-effect of windowing on BER reduction for $\varepsilon = 0.14$

Observing figure 5.5(b) at a BER of $10^{-2}$, a 6 dB performance gain can be obtained with employment of windowing of correlation order 1 than the standard OFDM system without any equalization techniques employed.

To see the consistency of this time domain windowing ICI reduction technique we should have to see the performance for various values of normalized frequency offsets as shown in Figure 5.6. For simplification and comparison it is enough to consider two different frequency offset values of, $\varepsilon = 0.1$ and $\varepsilon = 0.15$. Window with correlation order 1 is used.




(a) Windowing effect under AWGN channel




(b) Windowing effect under multipath Rayleigh fading channel

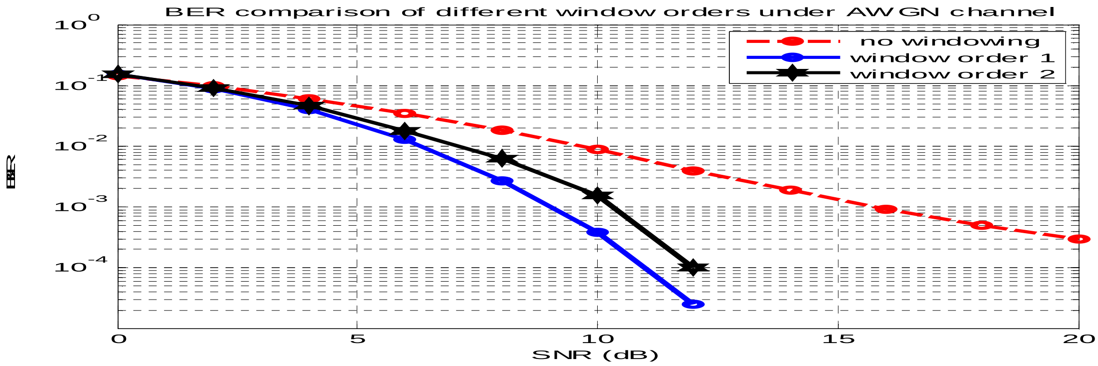
Figure 5.6 - performance of windowing for various frequency offsets

At the BER of $10^{-3}$ from Figure 5.6 (a), about 4.5 dB SNR gain is obtained under the normalized frequency offset of value 0.1 while about 12 dB SNR is gained under frequency offset of value 0.15 by using the proposed window compared to the standard OFDM systems with no proposed window. Of course, the proposed window at a frequency offset of value 0.15 cannot obtain as better BER performance as the window at a frequency offset of value 0.1.

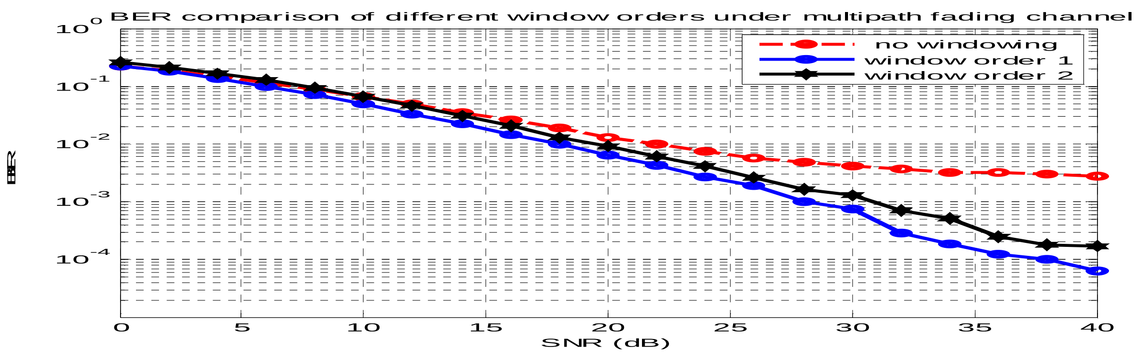
Similarly from figure 5.6(b) the SNR gain and BER reduction is better for frequency offset of 0.15 than that of 0.1 comparing OFDM system using the proposed window with the absence of the window.

Comparison of the BER performance in the Figure 5.6 implies that with the increase in the frequency offset value the BER increases and the time domain windowing technique offers a better performance compared to the standard OFDM system with no equalization. The optimized window function considered has a leading coefficient of $\alpha = 1$ and a correlation order of L=1 as derived in chapter four. The parameter $\alpha$ is not that much a significant parameter. But since the signal energy are transmitted in integral values, $\alpha$ must have an integral value for demodulation complexity reduction.

The use of different values of leading coefficient results in increased number of received symbols which are very adjacent to each other. Since it increases demodulation complexity it is advisable to use the theoretically derived optimum value of leading coefficient, $\alpha = 1$. Let's consider correlation orders of 1 and 2 under both AWGN channel and multipath fading channels.




(a) Comparison of windows under AWGN channel




(b) Comparison of windows under multipath fading channel

Figure 5.7- BER performance of windows of correlation order 1 and 2 for $\varepsilon = 0.13$

The curves in figure 5.7 show that the window with correlation order L=1 has a better performance than the window with L=2 (for normalized frequency offset of 0.13) by using the demodulation techniques designed in chapter four. As stated in the previous chapter the demodulation complexity increases with the increase in the correlation order of the window. Thus for convenience BERs of correlation order 1 and 2 windows are shown here.

### 5.4 comparison of time domain windowing with other equalization techniques

In this section, the BER performance measure of time domain windowing equalization technique is compared with other equalization techniques such as self cancellation, frequency domain correlative coding and also with the standard OFDM system where no equalization techniques implemented. In the figures 5.8 and 5.9 the performance is shown for the value ε =0.15.

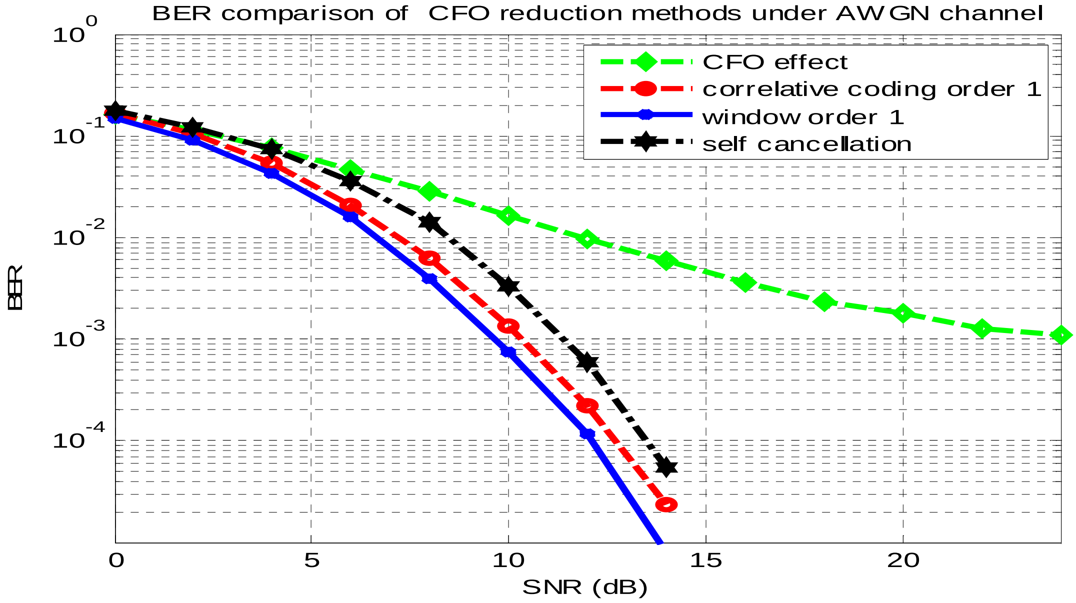



Figure 5.8- Comparison of ICI reduction techniques in AWGN channel

As shown in Figure 5.8 the BER performances of the time domain windowing, correlative coding and self cancellation schemes are reduced considerably compared to the standard OFDM without any equalization techniques.

Numerically, Table 5.3 below shows improvements brought and the SNRs saved by this windowing technique at the BER of $10^{-2}$, $10^{-3}$ and $10^{-4}$ which are equivalent to the results in Figure 5.8.

| BER values | SNRs (in dB) | | | |
|---|---|---|---|---|
| | Time domain window | Correlative coding | Self cancellation | Standard OFDM |
| $10^{-2}$ | 6.5 | 7 | 8.5 | 12 |
| $10^{-3}$ | 9.5 | 10.5 | 11.5 | 25 |
| $10^{-4}$ | 12 | 13 | 13.5 | >25 |

Table 5.3- SNR improvement comparison of different ICI reduction techniques

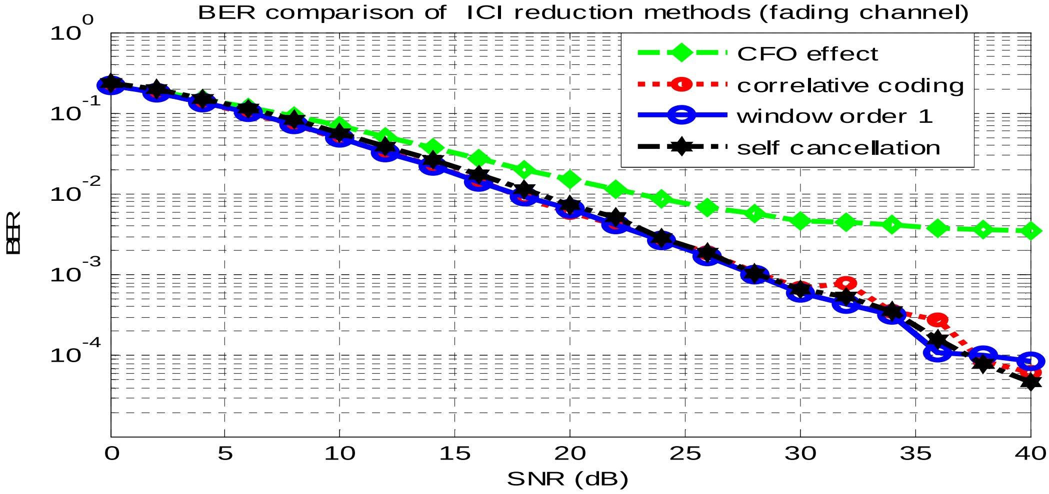



Figure 5.9- Comparison of ICI reduction techniques in multipath fading channel

Observing figure 5.9 at a BER of $10^{-2}$, a 7 dB performance gain can be obtained with employment of windowing of correlation order 1 than the standard OFDM system without any equalization techniques employed. Note that, when the SNR is high enough, the BER performance of the proposed scheme is better than that of the standard OFDM system.

The three equalization techniques considered here time domain windowing, frequency domain correlative coding of order 1 and self cancellation techniques show not exact performance in BER reductions. However, from the spectrum efficiency and outband power analysis we see that using the proposed windowing technique has extra advantages. Both time domain windowing and correlative coding techniques have almost 50% bandwidth efficient compared to self cancellation scheme. Time domain also effectively produces reduced outband power.

Table 5.4 shows the consistency of these techniques for the range of frequency offset values.

From the comparison of the average BER performance various ICI reduction techniques, the increase in the frequency offset value results in the increase in BER and the time domain windowing technique offers a better average BER performance compared to the correlative coding, self cancellation and standard OFDM system. Even though the three ICI reduction techniques have comparable performances the time domain window has a slightly reduced BER performance for a specified normalized frequency offset.

| Normalized Frequency Offset (ε) | Standard OFDM | Self cancellation | Correlative coding | Time domain window |
|---|---|---|---|---|
| | BER | BER | BER | BER |
| ε =0.1 | 0.02629 | 0.02632 | 0.02105 | 0.02103 |
| ε =0.15 | 0.04268 | 0.04108 | 0.03518 | 0.03388 |
| ε =0.2 | 0.06527 | 0.05699 | 0.05606 | 0.05455 |

Table 5.4-BER performance of various equalization techniques

Generally, it is better to use the time domain windowing since it has the best BER performance. The effect of the window on carrier to interference and spectral spreading is presented in the next sections.

**5.5 Effect of designed window on Carrier to Interference Ratio (CIR)**

Apart from BER, CIR is another performance measurement used to investigate the effect of time domain windowing. CIR serves as indication of good quality.

The intercarrier interference power and the carrier to interference signal power ratio for various frequency offset values are shown in Figures 5.10 and 5.11 below respectively. The curves are generated based on the analysis provided in Chapter four of Section 4.2 assuming that the standard transmitted data has zero mean and the symbols transmitted on different sub-carriers are statistically independent.

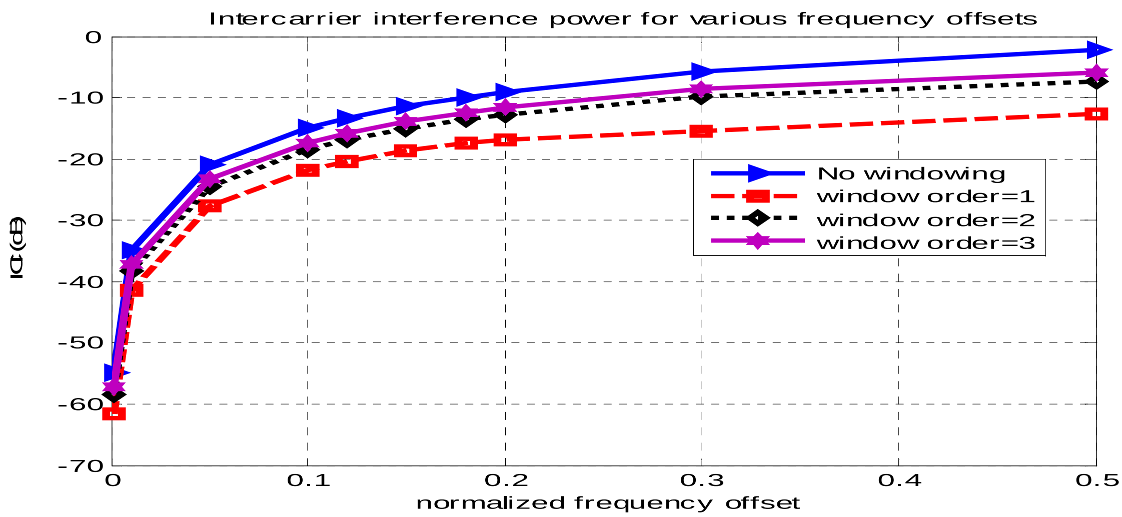



Figure 5.10- intercarrier interference power for various frequency offsets

The performances of windows of correlation order 1, 2 and 3 are shown in the figures and compared with the standard OFDM system with no ICI suppression scheme. The theoretical ICI power and CIR without the proposed windowing or other equalization techniques are plotted for comparison to evaluate the performances. The reduction of the ICI signal levels (figure 5.10) in the proposed windowing techniques leads to higher CIR.

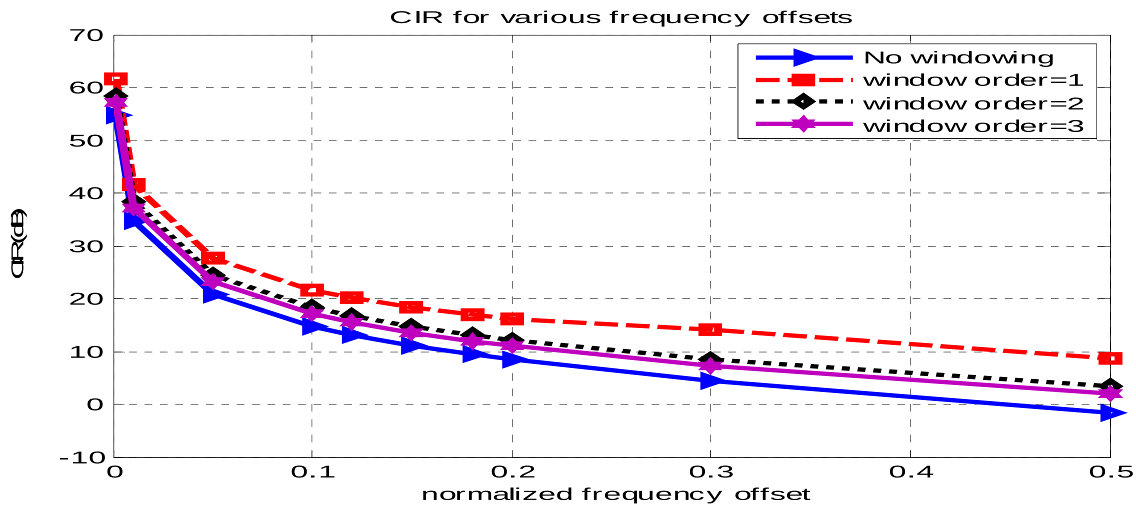



Figure 5.11- Carrier to interference signal power for various frequency offsets

It is shown that the designed windows of correlation orders L=1, 2 and 3 all have better performance in reducing the ICI power and increasing the CIR power compared to the standard OFDM systems without the consideration of windowing. The optimized value, the window with correlation order L=1, has better performance compared to windows with L=2 and L=3.

Figure 5.12 shows the CIR comparisons of the proposed windowing, the original correlative coding scheme and the self cancellation technique. It can be seen that the window of correlation order 1 outperforms the correlative coding, self cancellation technique and windows with L=1 and L=2.When the normalized frequency offset is large enough, the proposed scheme outperforms the correlative coding and self cancellation schemes, due to its better capabilities in suppressing ICI and preventing error propagation through OFDM symbols.

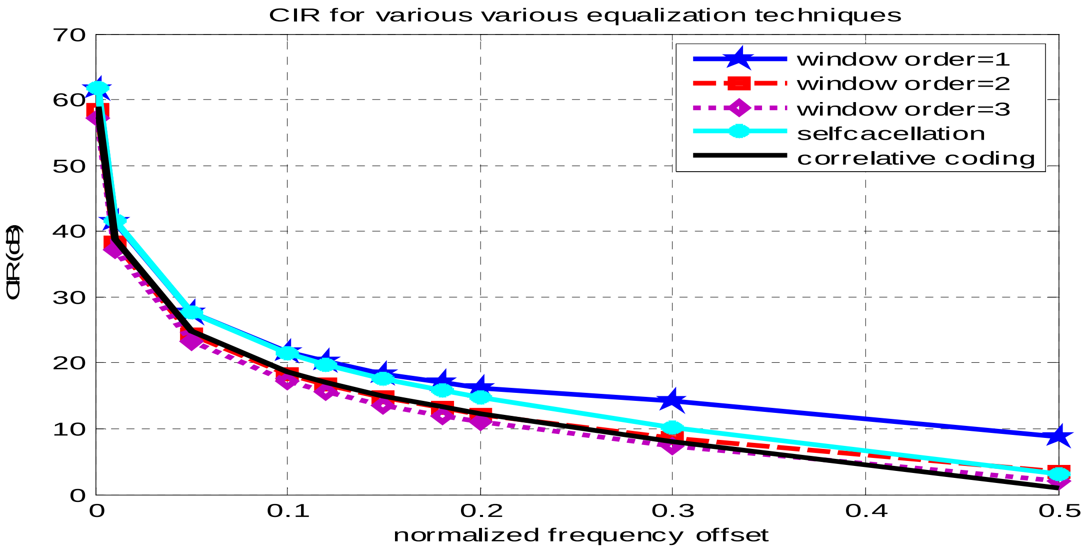



Figure 5.12- CIR comparison of different equalization techniques

From the simulated curves, it can be observed that, when the normalized frequency offset is zero the time domain equalization technique behaves same as the other techniques and the standard OFDM without any equalization technique. The effect of the time domain windowing can be explicitly seen from the Table 5.5, when the frequency offset value increases above 0.1.

| Normalized Frequency Offset (ε) | CIR (dB) | | | | |
|---|---|---|---|---|---|
| | Window (L=1) | Window (L=2) | Window (L=3) | Correlative coding | Self cancellation |
| ε =0.1 | 22 | 19 | 18 | 19 | 22 |
| ε =0.15 | 18 | 15 | 14 | 15 | 17 |
| ε =0.2 | 17 | 12 | 11 | 12 | 15 |

Table 5.5- CIR comparison of various ICI reduction techniques

As presented in chapter four, In the case of the demodulation process that needs no training or pilot signal the number of iterations increases as the window correlation order increases. Thus the demodulation complexity is measured based on the number of iterations in the proposed demodulation techniques in section 4.3 of chapter four.

For order L windowing and M-PSK or M-QAM modulation technique we may have a maximum of $M*2^L$ iterations to get correct detections.

- For QPSK modulation and order 1 windowing we may search up to 8 iterations to get the correct initial predictions.
- For QPSK modulation and order 2 windowing we may search up to 16 iterations to get the correct initial predictions.
- For QPSK modulation and order 3 windowing we may search up to 32 iterations to get the correct initial predictions.

Since the effect of previous subcarrier symbols on the present subcarrier symbol increases with the number of correlation order of the windows used, the probability of error propagation also increases. Thus the BER performance of the proposed time domain windowing scheme would be degraded due to error propagation for higher order windows.

Although higher order windows produce considerable reduction in outband power, they have increased demodulation complexity based on the number of iterations for correct detection, and relatively increased degraded BER due to error propagation. Thus there is no clear advantage to use more than order 2 windowing comparing the tradeoff between bandwidth efficiency and BER and demodulation complexity increase.

### 5.6 Summary of results

It has been shown that the proposed time domain windowing ICI reduction technique is able to mitigate ICI created by a time-varying channel. Compared to the estimation and correction methods, time domain windowing has very low complexity. In an estimation and correction method, aside from the computational complexity necessary to estimate the frequency offset, the correction alone requires higher computational complexity than the proposed system.

It is observed that at a BER of $10^{-3}$, about 10 dB (under AWGN) and more than 12 dB (under multipath fading channel) performance gains can be obtained with employment of windowing compared to the standard OFDM system without any equalization techniques employed at the normalized frequency offset of $\varepsilon = 0.15$. BER increases with the increase in the frequency offset value even though the time domain windowing technique offers a better performance.

The use of different values of the leading coefficient in windows results in increased number of received symbols which are very adjacent to each other. The use of fractional leading coefficient and increased window order increases demodulation complexity. Thus it is efficient to use the optimum values of $\alpha$=1 and L=1 for better performance and complexity reduction.

The BER performances of the proposed time domain windowing, correlative coding and self cancellation schemes are all considerably better compared to the standard OFDM without any equalization techniques. When the SNR is high enough, the BER performance of the proposed scheme is much better than that of the standard OFDM system. Even though the three ICI reduction techniques have comparable performances the time domain window has a slightly better BER performance compared to the correlative coding and self cancellation techniques for a given normalized frequency offset. The proposed windows have relatively better band width efficiency and outband power reduction.

From the CIR comparisons of the proposed windowing with the original correlative coding and the self cancellation schemes, it can be seen that the proposed window of correlation order 1 outperforms the correlative coding and self cancellation. When the normalized frequency offset is large enough, the proposed scheme outperforms the correlative coding and self cancellation schemes, due to its better capabilities in suppressing ICI and preventing error propagation through OFDM symbols.

This chapter clearly demonstrates the time domain window equalization technique is effective resulting in a better performance with less ICI. Generally, it is better to use the time domain windowing since it has the best BER performance and also bandwidth efficient technique. It has also better spectral spreading reducing capability.

## CHAPTER SIX

## CONCLUSION AND FUTURE WORKS

### 6.1 Conclusion

In this thesis work the time domain windowing equalization technique for intercarrier interference reduction is implemented and its performance is analyzed in terms of BER and CIR and compared with other equalization techniques such as frequency domain correlative coding, self cancellation technique, and also with standard OFDM system in which no ICI reduction technique is employed.

Over10 dB performance gain can be obtained with employment of windowing compared to the standard OFDM system without any equalization techniques employed at the normalized frequency offset of $\varepsilon = 0.15$ and BER of $10^{-3}$. BER increases with the increase in the frequency offset values even though the time domain windowing technique offers a better performance. When the SNR is high enough, the BER performance of the proposed scheme is much better than that of the standard OFDM system.

The BER performances of the proposed time domain windowing, correlative coding and self cancellation schemes are reduced considerably compared to the standard OFDM without any equalization techniques. Even though the three ICI reduction techniques have comparable performances the time domain window has a slightly reduced BER performance compared to the correlative coding and self cancellation techniques for a given normalized frequency offset. From the spectrum efficiency and outband power analysis we see that using the proposed windowing technique has extra advantages.

The proposed time domain windowing scheme outperforms the correlative coding and self cancellation schemes, due to its better capabilities in suppressing ICI and preventing error propagation through OFDM symbols when the normalized frequency offset is large enough.

The time domain windowing scheme shows better tolerance in frequency offset and by considerable reduction of the sensitivity to frequency errors. Since the windowing methods do not attempt to estimate a specific frequency shift, but instead reduce the level of ICI created as a result of the shift, they can also reduce the amount of ICI created by time variations in the channel without having to track the channel.

The conclusions of the study results are:

- ICI suppression scheme using time-domain windowing for OFDM systems, which can be regarded as an improved version of the correlative coding scheme, effectively reduces subcarrier frequency offset effects.
- Parameters for the windowing function have been optimized through theoretical analysis. The optimum well designed window is with correlation order 1 and leading coefficient 1 based on CIR maximization and results in reduced probability of error propagation and demodulation complexity.
- Appropriate demodulation algorithms: with no need of prior information of the transmitted sequence and with pilot symbols known by the transmitter and receiver has been developed.
- Computer simulation results have shown that the time domain windowing scheme gains improvements over the original correlative coding and self cancellation schemes in terms of both BER and CIR performances.
- The CIR performance for the proposed time domain windowing scheme is enhanced considerably compared to the self cancellation and correlative coding schemes, and the standard OFDM without any equalization techniques.
- When the normalized frequency offset is large enough, the proposed windowing scheme outperforms the correlative coding scheme, due to its better capabilities in suppressing ICI and preventing error propagation through OFDM symbols.

As the conclusion, the objectives of this research have been achieved. The study and investigation on the performance of time domain equalization using windowing in time varying radio channel conditions including radio channel impairment by AWGN have been done.

### 6.2 Future Works

Although the objective of this project has been achieved successfully, it still need further study in order to dig this project deeper and wider. Several suggested future works that can be done are listed below.

- Consider error correction techniques such as Forward error correction for mitigation of problems on bad channels especially when large numbers of subcarriers are considered.
- In depth investigation on the design of demodulation techniques when the correlation window orders increase. Implement the MLSD techniques to for better BER performance.
- Investigation has to be made on the tradeoff between demodulation complexity and bandwidth efficiency of OFDM channels to use outband power reduction ability of higher correlation order windows.
- Use the time domain windowing equalization with other complex equalization techniques to reduce the residual errors in highly dispersive channels.
- Adopt the length of the cyclic prefix to further increase spectral efficiency depending on the radio environment.

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