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NEW DEVELOPMENT AND PERFORMANCE EVALUATION OF TRANSFORM DOMAIN OFDM BASEBAND SYSTEM

A THESIS SUBMITTED IN PARTIAL REQUIREMENTS FOR THE DEGREE OF

BACHELOR OF TECHNOLOGY

IN

ELECTRONICS & COMMUNICATION ENGINEERING

By

Vijay Kumar

Roll No. : 10509026

Under the Guidance of

Prof. G. Panda

Department of Electronics & Communication Engineering
National Institute of Technology, Rourkela
Orissa 769008

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CERTIFICATE

This is to certify that the thesis entitled "**New Development And Performance Evaluation Of Transform Domain OFDM Baseband System**" submitted by Sri Vijay Kumar in partial fulfillment of the requirements for the award of Bachelor of Technology Degree in Electronics and Communication Engineering at National Institute of Technology, Rourkela (Deemed University) is an authentic work carried out by him under my supervision and guidance.

To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other University/ Institute for the award of any Degree or Diploma.

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CONTENTS

ABSTRACT

Wireless Communication has gained much popularity in the field of Communication because of its ability to transfer the data at high rate with much more increased high quality, low cost and better performance. Also it offers variety of services for the wide range of applications.

Orthogonal frequency division multiplexing (OFDM) is becoming widely applied in wireless communications system due to its high rate transmissions capability with high bandwidth efficiency and its robustness with regard to multipath and delay. It has been used in digital audio broadcasting systems (DAB), digital video broadcasting (DVB) systems, digital subscriber line (DSL) standards, and wireless LAN standards such as the American IEEE std. 802.11(WiFi) and WiMAX (stands for Worldwide Interoperability for Microwave Access), are one of the standards of IEEE which utilizes the idea of OFDM, and is aimed to provide high-throughput broadband connections over long distances.

Conventially we use discrete Fourier transform (DFT) in OFDM system. This thesis presents the simulation of 4 Qudrature Amplitude Modulation (QAM) orthogonal frequency divison multiplexing (OFDM) baseband system and channel estimation which uses inverse discrete Hartley transform (IDHT) and discrete Hartley transform (DHT). As the calculation of DHT and IDHT involves real operations hence the computational complexities are less as comapred to DFT and IDFT. Moreover as IDHT is same as DHT hence we can use same hardware for both, while the DFT and IDFT require separate hardware to implement. As compared to DFT based OFDM system, the simulated DHT based OFDM system achieves approximately the same transmission performance with less computational complexity and hardware requirements.

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ABBREVIATIONS USED

OFDM	Orthogonal Frequency Division Multiplexing
DFT	Discrete Fourier Transform
DHT	Discrete Hartley Transform
LS Channel Estimator	Least Squares Channel Estimators
MLS Channel Estimator	Modified Least Squares Channel Estimators
BER	Bit Error Rate
SNR	Signal to Noise Ratio
PED-B Channel	Pedestrian B Channel

Chapter 1

INTRODUCTION

1.1 Basics of Wireless Communication

Wireless communication is one of the most active areas of technology development. Wireless Broadband Technologies allow the simultaneous delivery of voice, data and video over fixed or mobile platforms **[1]**. Wi-Fi, WiMAX, LTE, UMB are some of the emerging technologies. WiMAX acronym stands for World Wide Interoperability for Microwave Access. LTE stands for Long Term Evolution. These standards are wireless technologies that provide high throughput broadband connections over long distances **[15].**

Frequency division multiplexing (FDM) is a technology that transmits multiple signals simultaneously over a wired or a wireless system. Each signal is limited by a specific frequency band, and is modulated by data stream. Orthogonal Frequency Division Multiplexing is a special case of this, where the data is distributed over a large number of carriers, which are 'orthogonal' to each other. OFDM is spectrally efficient compared to the conventional FDM systems, since it doesn't need guard bands between adjacent channels. This orthogonality property of the carriers is at the heart of OFDM, since the interference due to the other carriers is prevented, when the receiver demodulates a particular carrier. IEEE 802.16: Wireless MAN and WiMAX (stands for Worldwide Interoperability for Microwave Access), are one of the standards of IEEE which utilizes the idea of OFDM.

OFDM is a robust and efficient modulation scheme, which mitigates some of the channel impairments quite effectively and support high data rates. It combats multipath fading and narrow band interference efficiently **[16]**.

Here we shall discuss the motivation for the study as well as the insight into the significance of design of an OFDM system using Discrete Hartley Transform.

1.2 Motivation

One of the principal advantages of OFDM is its utility for transmission at very nearly optimum performance in unequalized channels and in multipath channels. Inter-symbol interference (ISI) and inter-carrier interference (ICI) can be entirely eliminated by the simple expedient of inserting between symbols a small time interval known as guard interval **[17].**

One of the principal disadvantages of OFDM is sensitivity to frequency o□set in the channel. There are two deleterious effects caused by frequency o□set; one is the reduction of signal amplitude in the output of the filters matched to each of the carriers and the second is introduction of ICI from the other carriers which are now no longer orthogonal to the filters. Timing errors also affects OFDM system performance by reducing the delay spread robustness and by introducing phase shift in the received spectrum**.**

Adaptive estimation of channel is necessary before the demodulation of OFDM signals since the wireless channel is frequency selective and time- varying. There are two main problems in designing the channel estimators for wireless OFDM systems. The first problem is the arrangement of pilot information, where the pilots means the reference signal used by the both the transmitter and the receiver. The second problem is to design an estimator with both low complexity and good channel tracking ability. Third problem is, while considering the design of system the major parameters are computational complexity and implementation cost. In the current DFT based OFDM system transceivers, the modulator needs to compute a long length inverse discrete Fourier transform (IDFT) and the demodulator needs to compute a long length

DFT, where transform length is up to 512 or more. For such long-length DFT/IDFT computations, a great number of complex multiplications are required. Hence DFT based OFDM system involves a lot of computational complexity and high implementation cost.

Motivated by above problem of OFDM timing acquisition and frequency synchronization we have implemented a robust timing acquisition (Scmidl-Cox algorithm) and frequency synchronization algorithm for OFDM and to cope with computational complexity and implementation cost we have designed a Discrete Hartley Transform (DHT) based OFDM system in the thesis report.

We have also implemented the following three algorithm of channel estimation is implemented to correct the timing error, which is like a rotating phasor in the frequency domain.

1) Least Squares (LS)/ Zero forcing method 2) Modified Least Squares (MLS)

1.3 Contribution

In the current DFT based OFDM system transceivers, the modulator needs to compute a long length inverse discrete Fourier transform (IDFT) and the demodulator needs to compute a long length DFT, where transform length is up to 512 or more. For such long-length DFT/IDFT computations, a great number of complex multiplications are required and each of them basically involves four real multiplications and two real additions. Clearly, the complexity of a DFT-based or OFDM-based transceiver would be reduced if the corresponding modulator/demodulator could be implemented using purely real transforms while fast algorithms similar to the fast Fourier transform (FFT) algorithm can still be applied. In this project report, we simulated an OFDM system which is based on Discrete Hartley transform (DHT) and inverse Discrete Hartley transform (IDHT) for modulation and demodulation. The

DHT involves only real-valued arithmetic and has an identical inverse. Hence the simulated OFDM system model has reduced computational complexity but same performance compared to DFT based model. Finally the results of OFDM system based on DFT and DHT are compared after channel estimation with the help of BER versus SNR curve.

1.4 Thesis Outline

The organization of the thesis is as follows. Following the brief introduction, motivation and contibution of the thesis, the basics of modulation and OFDM baseband system are outlined in chapter 2. Chapter 3 discusses OFDM timing and frequency synchronization and Schmidl-Cox algorithm. Chapter 4 OFDM channel estimation using least squares (LS) and modified least squares (MLS) channel estimator. In Chapter 5 DHT based OFDM system and its performance analysis and results in comparison with DFT based OFDM system are discussed. Finally, chapter 6 describes the concluding remark.

Chapter 2

BASICS OF MODULATION & OFDM BASEBAND MODEL

2.1 Single Carrier Modulation System

A single carrier system modulates information onto one carrier using frequency, phase, or amplitude adjustment of the carrier **[18]**. For digital signals, the information is in the form of bits, or collections of bits called symbols, that are modulated onto the carrier. As higher bandwidths (data rates) are used, the duration of one bit or symbol of information becomes smaller. The system becomes more susceptible to loss of information from impulse noise, signal reflections and other impairments. These impairments can impede the ability to recover the information sent. In addition, as the bandwidth used by a single carrier system increases, the susceptibility to interference from other continuous signal sources becomes greater. This type of interference is commonly labeled as carrier wave (CW) or frequency interference.

2.2 Frequency Division Multiplexing Modulation System

Frequency division multiplexing (FDM) extends the concept of single carrier modulation by using multiple subcarriers within the same single channel. The total data rate to be sent in the channel is divided between the various subcarriers. The data do not have to be divided evenly nor do they have to originate from the same information source. Advantages include using separate modulation/demodulation customized to a particular type of data, or sending out banks of dissimilar data that can be best sent using multiple, and possibly different, modulation schemes.

Current national television systems committee (NTSC) television and FM stereo multiplex are good examples of FDM. FDM offers an advantage over single-carrier modulation in terms of narrowband frequency interference since this interference will only affect one of the frequency sub bands. The other subcarriers will not be affected by the interference. Since each subcarrier has a lower information rate, the data symbol periods in a digital system will be longer, adding some additional immunity to impulse noise and reflections.FDM systems usually require a guard band between modulated subcarriers to prevent the spectrum of one subcarrier from interfering with another. These guard bands lower the systems effective information rate when compared to a single carrier system with similar modulation.

2.3 Basics of OFDM System

Orthogonal Frequency Division Multiplexing (OFDM), is now a popular technique for MCM (Multi-Carrier Modulation), is deployed in various standards of IEEE, especially in the wireless systems. It also looks promising for the 4G mobile technologies. OFDM converts a frequency selective fading channel into a collection of the flat fading sub-channels. The key ideas of OFDM were patented, on 1967-68. The wireless channels offers much more unpredictability and other challenges than their wire line (like twisted wire pairs or coaxial cables) counterparts, due to the presence of multipath, Doppler spread etc., This difficulty in the wireless channels is mainly due to the frequent change in the environment and other factors because of the mobility of the user, and presence of different environment conditions.

2.3.1 OFDM for Multicarrier Transmission

In a wireless communication system, the signal is carried by a large number of paths with different strengths and delays. Such multipath dispersion of the signal is commonly referred as "channel-induced inter symbol interference (ISI)." In fact, the multipath dispersion leads to an upper limitation of the transmission rate in order to avoid the frequency selectivity of the channel or the need of a complex adaptive equalization in the receiver. In order to mitigate the time-dispersive nature of the channel, single-carrier serial transmission at a high data rate is replaced with a number of slower parallel data streams. Each parallel stream will be then used to sequentially modulate a different subcarrier. By creating N parallel sub streams, we will be able to decrease the bandwidth of the modulation symbol by the factor of N, or, in other words, the duration of a modulation symbol is increased by the same factor. The summation of all of the individual subchannel data rates will result in total desired symbol rate, with the drastic reduction of the ISI distortion.

If the FDM system above had been able to use a set of subcarriers that were orthogonal to each other, a higher level of spectral efficiency could have been achieved. The guard bands that were necessary to allow individual demodulation of subcarriers in an FDM system would no longer be necessary. The use of orthogonal subcarriers would allow the subcarriers' spectra to overlap, thus increasing the spectral efficiency. As long as orthogonality is maintained, it is still possible to recover the individual subcarriers' signals despite their overlapping spectrums. If the dot product of two deterministic signals is equal to zero, these signals are said to be orthogonal to each other. Orthogonality can also be viewed from the standpoint of stochastic processes. If two random processes are uncorrelated, then they are orthogonal. Given the random nature of signals in a communications system, this probabilistic view of orthogonality provides an intuitive understanding of the implications of orthogonality in OFDM. As we know the sinusoids of the DFT form an orthogonal basis set, and a signal in the vector space of the DFT can be represented as a linear combination of the orthogonal sinusoids. One view of the DFT is that the transform essentially correlates its input signal with each of the sinusoidal basis functions. If the input signal has some energy at a certain frequency, there will be a peak in the correlation of the input signal and the basis sinusoid that is at that corresponding frequency. This transform is used at the OFDM transmitter to map an input signal onto a set of orthogonal subcarriers, i.e., the orthogonal basis functions of the DFT. Similarly, the transform is used again at the OFDM receiver to process the received subcarriers. The signals from the subcarriers are then combined to form an estimate of the source signal from the transmitter. The orthogonal and uncorrelated nature of the subcarriers is exploited in OFDM with powerful results. Since the basis functions of the DFT are uncorrelated, the correlation performed in the DFT for a given subcarrier only sees energy for that corresponding subcarrier. The energy from other subcarriers does not contribute because it is uncorrelated. This separation of signal energy is the reason that the OFDM subcarriers' spectrums can overlap without causing interference.

In OFDM, the orthogonal subcarriers are separated by a frequency interval of $\Delta f = 1/Ts$, where Ts is the OFDM symbol duration, as shown in Fig. 2.1.

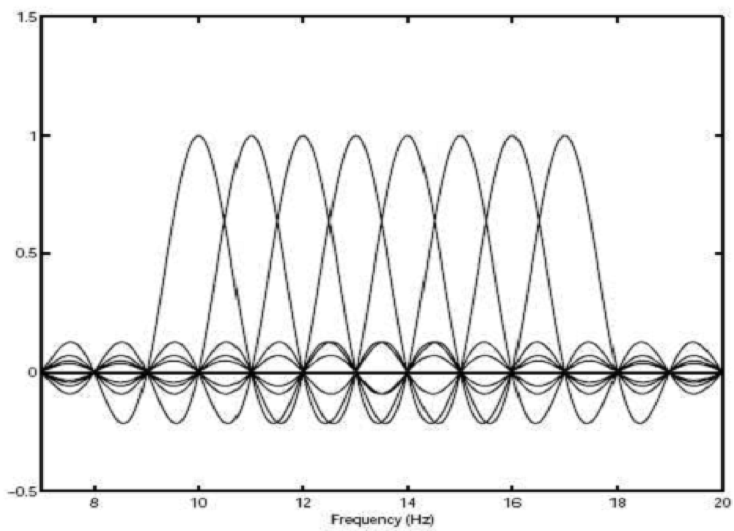

Fig.2.1 OFDM transmission spectrum

The frequency spectrum of the adjacent subchannel overlap with one another, but the orthogonality of subcarriers will eliminate in principle the inter-channel interference (ICI).

2.4 Implementation of DFT based OFDM System

The idea behind the analog implementation of OFDM can be extended to the digital domain by using the discrete Fourier Transform (DFT) **[2, 3]** and its counterpart, the inverse discrete Fourier Transform (IDFT). These mathematical operations are widely used for transforming data between the time-domain and frequency-domain. These transforms are interesting from the OFDM perspective because they can be viewed as mapping data onto orthogonal sub-carriers. For example, the IDFT is used to take in frequency-domain data and convert it to time-domain data. In order to perform that operation, the IDFT correlates the frequency-domain input data with its orthogonal basis functions, which are sinusoids at certain frequencies. This correlation is equivalent to mapping the input data onto the sinusoidal basis functions.

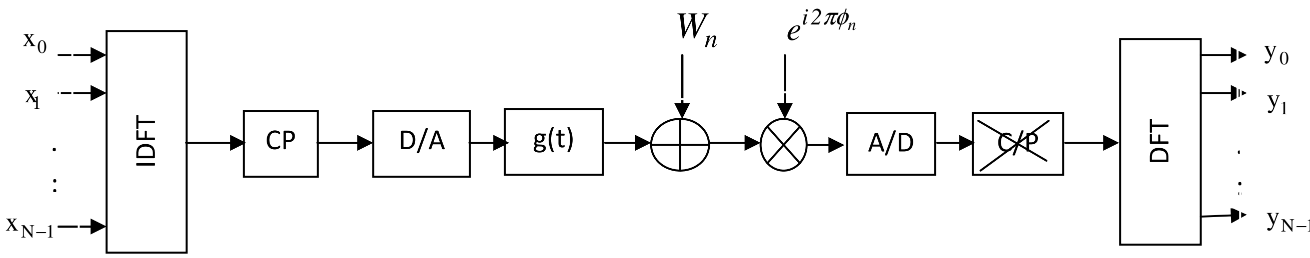

Fig 2.2 Typical OFDM baseband system, affected only by noise, timing offset and frequency offset. W_n represents AWGN noise, ϕ_n represents a time varying phase which (artificially mimics) implements frequency offset , CP refers to cyclic prefix A/D refers to analog to digital converter and D/A refers to digital to analog converter.

In practice, OFDM systems are implemented using a combination of fast Fourier Transform (FFT) and inverse fast Fourier Transform (IFFT) blocks that are mathematically equivalent versions of the DFT and IDFT, respectively, but more efficient to implement. An OFDM system treats the source symbols (e.g., the QPSK or QAM symbols that would be present in a single carrier system) at the transmitter as though they are in the frequency-domain. These symbols are used as the inputs to an IFFT block that brings the signal into the time domain. The IFFT takes in N symbols at a time where N is the number of subcarriers in the system. Each of these N input symbols has a symbol period of T seconds. Recall that the basis functions for an IFFT are N orthogonal sinusoids. These sinusoids each have a different frequency and the lowest frequency is DC. Each input symbol acts like a complex weight for the corresponding sinusoidal basis function. Since the input symbols are complex, the value of the symbol determines both the amplitude and phase of the sinusoid for that subcarrier. The IFFT output is the summation of all N sinusoids. Thus, the IFFT block provides a simple way to modulate data onto N orthogonal subcarriers. The block of N output samples from the IFFT make up a single OFDM symbol. The length of the OFDM symbol is NT where T is the IFFT input symbol period mentioned above.

After some additional processing, the time-domain signal that results from the IFFT is transmitted across the channel. At the receiver, an FFT block is used to process the received signal and bring it into the frequency domain. Ideally, the FFT output will be the original symbols that were sent to the IFFT at the transmitter. When plotted in the complex plane, the FFT output samples will form a constellation, such as 4-QAM. However, there is no notion of a constellation for the time-domain signal. When plotted on the complex plane, the time-domain signal forms a scatter plot with no regular shape. Thus, any receiver processing that uses the concept of a constellation (such as symbol slicing) must occur in the frequency-domain. The block diagram in Figure 2.2 illustrates the switch between frequency-domain and time-domain in an OFDM system.

2.5 Multipath Channels and Use of Cyclic Prefix

A major problem in most wireless systems is the presence of a multipath channel. In a multipath environment, the transmitted signal reflects off of several objects. As a result, multiple delayed versions of the transmitted signal arrive at the receiver. The multiple versions of the signal cause the received signal to be distorted. Many wired systems also have a similar problem where reflections occur due to impedance mismatches in the transmission line.

A multipath channel will cause two problems for an OFDM system **[9]**. The first problem is intersymbol interference. This problem occurs when the received OFDM symbol is distorted by the previously transmitted OFDM symbol. The effect is similar to the intersymbol interference that occurs in a single-carrier system. However, in such systems, the interference is typically due to several other symbols instead of just the previous symbol; the symbol period in single carrier systems is typically much shorter than the time span of the channel, whereas the typical OFDM symbol period is much longer than the time span of the channel. The second problem is unique to multicarrier systems and is called Intrasymbol Interference. It is the result of interference amongst a given OFDM symbol's own subcarriers. The next sections illustrate how OFDM deals with these two types of interference.

2.6 Intersymbol Interference

Assume that the time span of the channel is L_C samples long. Instead of a single carrier with a data rate of R symbols/second, an OFDM system has N subcarriers, each with a data rate of R/N symbols/second **[13]**. Because the data rate is reduced by a factor of N, the OFDM symbol period is increased by a factor of N. By choosing an appropriate value for N, the length of the OFDM symbol becomes longer than the time span of the channel. Because of this configuration, the effect of intersymbol interference is the distortion of the first L_C samples of the received OFDM symbol. By noting that only the first few samples of the symbol are distorted, one can consider the use of a guard interval to remove the effect of intersymbol interference. The guard interval could be a section of all zero samples transmitted in front of each OFDM symbol. Since it does not contain any useful information, the guard interval would be discarded at the receiver. If the length of the guard interval is properly chosen such that it is longer than the time span of the channel, the OFDM symbol itself will not be distorted. Thus, by discarding the guard interval, the effects of intersymbol interference are thrown away as well.

2.7 Intrasymbol Interference

The guard interval is not used in practical systems because it does not prevent an OFDM symbol from interfering with itself **[7, 13]**. This type of interference is called intrasymbol interference. The solution to the problem of intrasymbol interference involves a discrete-time property. Recall that in continuous-time, a convolution in time is equivalent to a multiplication in the frequency domain. This property is true in discrete-time only if the signals are of infinite length or if at least one of the signals is periodic over the range of the convolution. It is not practical to have an infinite-length OFDM symbol; however, it is possible to make the OFDM symbol appear periodic. This periodic form is achieved by replacing the guard interval with something known as a cyclic prefix of length L_P samples. The cyclic prefix is a replica of the last LP samples of the OFDM symbol where $L_P > L_C$. Since it contains redundant information, the cyclic prefix is discarded at the receiver. Like the case of the guard interval, this step removes the effects of intersymbol interference. Because of the way in which the cyclic prefix was formed, the cyclically-extended OFDM symbol now appears periodic when convolved with the channel. An important result is that the effect of the channel becomes multiplicative. In a digital communications system, the symbols that arrive at the receiver have been convolved with the time-domain channel impulse response of length L_C samples. Thus, the effect of the channel is convolutional. In order to undo the effects of the channel, another convolution must be performed at the receiver using a time-domain filter known as an equalizer. The length of the equalizer needs to be on the order of the time span of the channel. The equalizer processes symbols in order to adapt its response in an attempt to remove the effects of the channel. Such an equalizer can be expensive to implement in hardware and often requires a large number of symbols in order to adapt its response to a good setting.

In OFDM, the time-domain signal is still convolved with the channel response. However, the data will ultimately be transformed back into the frequency-domain by the FFT in the receiver. Because of the periodic nature of the cyclically-extended OFDM symbol, this time-domain convolution will result in the multiplication of the spectrum of the OFDM signal (i.e., the frequency-domain constellation points) with the frequency response of the channel. The result is that each subcarrier's symbol will be multiplied by a complex number equal to the channel's frequency response at that subcarrier's frequency. Each received subcarrier experiences a complex gain (amplitude and phase distortion) due to the channel. In order to undo these effects, a frequency-domain equalizer is employed. Such an equalizer is much simpler than a time-domain equalizer. The frequency-domain equalizer consists of a single complex multiplication for each subcarrier. For the simple case of no noise, the ideal value of the equalizer's response is the inverse of the channel's frequency response.

Chapter 3

OFDM FREQUENCY & TIMING SYNCHRONIZATION

3.1 OFDM Frequency Synchronization Errors

OFDM modulation encodes the data symbols onto orthogonal subchannels, where orthogonality is assured by the subcarrier separation $\Delta f = 1/Ts$. The subchannels may overlap in the frequency domain, as shown in Fig. 2.1 for a rectangular pulse shape in time (sinc function in frequency). In practice, the frequency separation of the subcarriers is imperfect, so Δf is not exactly equal to 1/Ts. This is generally caused by mismatched oscillators, Doppler frequency shifts, or timing synchronization errors **[20]**.

Carrier frequency errors result in a shift of the received signal in the frequency domain. If the frequency error is an integer multiple of the subcarrier spacing Δf, then the received frequency domain quadrature amplitude modulated (QAM) subcarriers are shifted by n subcarrier positions. The subcarriers are still mutually orthogonal but the received data symbols, which were mapped to the OFDM spectrum, are now in the wrong position in the demodulated spectrum, resulting in BER degradation.

If the carrier frequency error is not an integer multiple of the subcarrier spacing, then energy spills over between the subcarriers, resulting in loss of their mutual orthogonality. Interference is then observed between the subcarriers, leading to ICI.

3.2 OFDM Timing Synchronization Errors

Unlike the frequency mismatch discussed above, time synchronization errors do not result in inters subcarrier interference. However, even small misalignments of the FFT window result in an evolving phase shift in the frequency domain symbols, leading to BER degradation. If the receiver’s FFT window is shifted in the received sampling stream, then the time shift property of the Fourier transform can be formulated as

$$f(t) \leftrightarrow F(\omega) \tag{3.1}$$

$$f(t-\tau) = e^{-j\omega\tau} F(\omega) \tag{3.2}$$

Any misalignment τ of the receiver's FFT window will introduce a phase error of 2πΔfτ/Ts between two adjacent subcarriers. If the timing errors are so high that the FFT window of the receiver includes samples outside the data and guard segments of the current OFDM symbol, then the consecutive OFDM symbols interfere, severely affecting the system's performance. When the guard interval is followed by the data samples, a moderately delayed FFT receiver window may overlap with the next OFDM symbol, while an early FFT window will include samples of the data segment and the guard interval. The latter case will not introduce any interference, while the former case is much more detrimental to the performance.

Therefore, timing and frequency synchronization between the transmitter and the receiver are of crucial importance in terms of the performance of an OFDM link **[20]**.

3.3 OFDM Timing & Frequency Offsets

We consider an OFDM system with N subcarriers, which includes N_G subcarriers for guard band and the zero-DC subcarrier. A cyclic length of N_{CP} is assumed, where typically N_{CP} = N/L, L = 32, 16, 8, or 4. The samples of the transmitted baseband sequence corresponding to the n^{th} OFDM symbol is then given by

$$s_{n,k} = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} C_{n,m} e^{j\frac{2\pi}{N}km}, 0 \le k \le N-1 \tag{3.3}$$

Where $C_{n,m}$ is complex modulated data on the m^{th} subcarrier.

After transmission over a multipath fading channel, the k^{th} sample of the received OFDM symbol can be written as

$$y_{n,k} = \frac{1}{N}\sum_{m=0}^{N-1} C_{n,m} H_{n,m} e^{j\frac{2\pi}{N}k(m+\partial f)} e^{j\frac{2\pi}{N}\tau m} + \omega_{n,k} \tag{3.4}$$

Where $H_{n,m}$ is the frequency response of the multipath channel on the m^{th} subcarrier, $\omega_{n,k}$ represents circular complex Gaussian noise samples, and τ represents the timing error. Here, ∂f is the normalized frequency offset given by the ratio of the actual frequency offset (in Hz) to the inter-subcarrier spacing. The DFT output of the received symbol can be expressed as

$$Y_{n,m} = a_{n,m} C_{n,m} H_{n,m} e^{2\pi\phi_{n,m}} + I_{n,m}(\partial f, \tau) + W_{n,m} \tag{3.5}$$

where $a_{n,m}$ and $\phi_{n,m}$ are the attenuation and the phase rotation factors on the m^{th} subcarrier respectively, $I_{n,m}(\partial f, \tau)$ is the inter-carrier interference and inter-block interference due to frequency and timing errors respectively, and $W_{n,m}$ is the noise component on this subcarrier.

3.4 Preamble & OFDM Symbol Generation

The Preamble samples are generated using IDcell 9 specification in IEEE 802.16d/e standard Ref[1] as shown in Fig.3.1. The amplitude of the sample is kept such that the average power of preamble is 3dB more the rest of the OFDM symbols. Out of the 512 subcarriers first 42 and last 41 subcarriers are null subcarriers. The DC subcarrier (257th subcarrier) is also a null subcarrier. The non-zero subcarriers extend from 43rd symbol index to 471th symbol index with an interleaved null subcarrier between successive non-zero subcarriers. There are in effect 214 non-zero subcarriers in the Preamble and a 512 point IFFT is performed on the preamble.

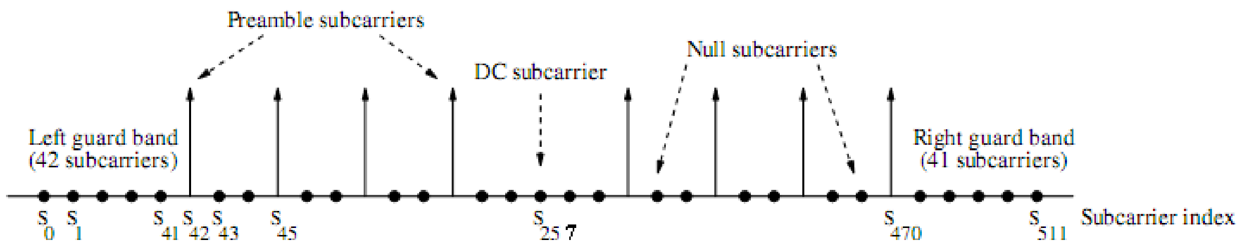

Fig.3.1. OFDM Preamble Symbol Structure for N=512 as specified in IEEE 802.16e standard, OFDM transmitted signal.

3.5 OFDM Timing Acquisition & Frequency Offset Estimation

3.5.1. Schmidl Cox Algorithm for OFDM Timing Acquisition

The transmitted samples are send through a noisy channel (i.e., channel gain unity and only white Gaussian noise of specified variance is added to the transmitted samples) and the received samples, $r(n)$ so obtained, are used to find the start of frame (timing recovery) as well as fractional frequency offset is estimated using the Schmidl Cox Algorithm **[12].**

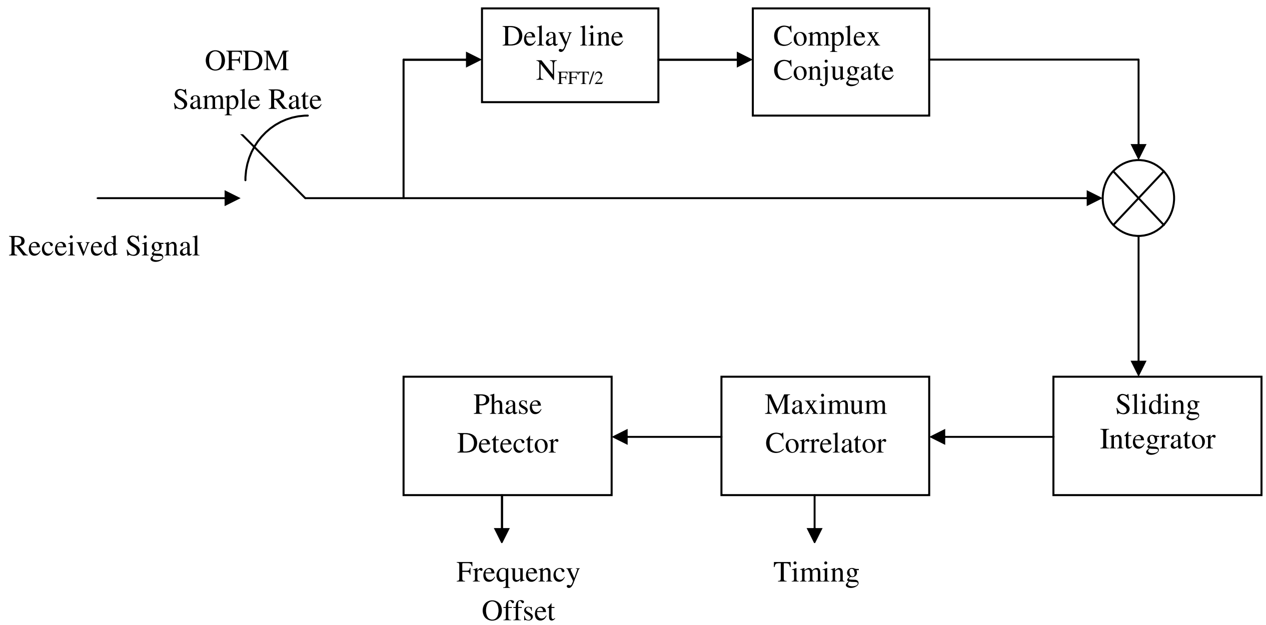

Fig.3.2 OFDM Synchronization using Schmidl Cox Algorithm

Autocorrelation is performed on the received sequence to find the start of frame. The l^{th} sample of autocorrelation output can be written as

$$z(l)=\sum_{m=0}^{N/2} r(l+m) * r^{*}(l+m+N/2) \tag{3.6}$$

The correlation $z(l)$ will have a constant "Plateau" of 64 samples, when there is no delay spread in the channel. Also, for example, if the channel is of length 26 samples, then the plateau length will be reduced to (64-26) =38 samples. Actually, any declaration of "start-of-symbol" (which specifies the FFT window) within this "plateau top" is good enough for OFDM receiver to correctly block decode the data. Indeed, this is the strong point of OFDM and OFDMA modulation. However, from a system view point where a frame of OFDM symbols are transmitted starting with a preamble symbol, it is important accurately estimate the start of the frame.

One method to exactly find the "Edge of the Plateau on the right Corner" is by differentiation of the smoothed samples $Z(l)$, as given below:

$$Y(l) = Z(l) - Z(l-1) \tag{3.7}$$

$$D(i) = \sum_{k=i-63}^{i} abs\ (Y(i)) \tag{3.8}$$

$$[c, m] = \min(D) \tag{3.9}$$

m+1 give the start of frame.

3.5.2 Fractional Frequency Offset Estimation

Once the timing acquisition is done, the autocorrelation output can be used to estimate the fractional frequency offset. The phase of the correlation output is equal to the phase drift between the samples that are N OFDM samples apart. The estimate of the normalized fractional frequency offset is given by

$$\Delta \hat{f}_{frac} = \frac{1}{\pi} angle\ (Z(m)) \tag{3.10}$$

The fractional frequency offset estimated is removed by multiplying the received sample with

the $\exp(-\frac{2\pi j * \Delta \hat{f} n}{512})$ (3.11)

3.5.3 Integer Frequency Offset Estimation

The integer part of the frequency offset can be estimated by using guard and null carriers in the Preamble OFDM symbol. As shown in Fig. 3.1, there are guard bands on both sides and 214 BPSK modulated subcarriers in every second position. By searching for these subcarriers in frequency domain, integer frequency offset can be estimated. At different subcarrier positions around the start of the non guard subcarriers, the total power in a window of length equal to number of non-guard subcarriers can be found. The number of subcarriers positions by which the subcarrier index, corresponding to maximum power in the window, is away from the 42nd subcarrier index (first no guard subcarrier without integer offset) gives the residual integer frequency.

The integer frequency offset is removed by multiplying the samples with

$$\exp(-j\frac{2\pi}{512}(\Delta f_{\text{int}} + \Delta f_{frac})n) \quad (3.12)$$

.

3.5.4 Phase Offset Estimation

There is a constant Phase rotation of the QPSK constellation when passed through the kit due to different attenuations for the in phase and Quadrature phase components. This can be estimated by doing zero forcing on any one non-zero subcarrier of the Preamble. The estimated phase offset is removed from all the samples in the OFDM symbols by multiplying with exp (-φ) where φ is the estimated phase offset.

3.6 Observations & Analysis

One OFDM frame in our experiment consists of 10 OFDM symbols. Each OFDM symbol has 428 non-zero subcarriers (i.e., from 43 to 471). Thus, 4280(428*10) QPSK symbols are generated and a 512 point IFFT is performed on the OFDM symbols. Therefore, an OFDM frame consists of a Preamble and 10 OFDM symbols. Preamble is always the first symbol of an OFDM frame followed by 10 OFDM symbols. Cyclic Prefix of length $N_{FFT}/8$ is inserted at the start of each OFDM symbol as shown in Fig 3.3.

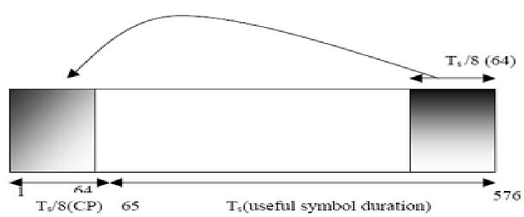

Fig.3.3 Addition of the Cyclic Prefix to form the transmitted OFDM symbol

Frequency offset will be introduced when we send the samples from card-to-card after IF modulation. However, in the base-band loop-back mode, or even when going from card to card via base-band, there will not be a (considerable) frequency error introduced on the received samples. In such situations, both integer and fractional frequency offsets are modeled, by multiplying the nth transmit sample with $\exp(-j\frac{2\pi}{512}(\Delta fn)$. For example offset=4.3 means integer offset=4 and fractional frequency offset=0.3. The real and the imaginary parts of each sample in the transmitted frame are the In-phase and the Quadrature-phase components, respectively. Since we are using 8 bits, I and Q samples are then converted into range -127 to 127. The samples are written into a file as I Q I Q I Q…. as expected by the WiCOMM-T kit driver. The QPSK data generated are also written into a file, because known symbols are required at the receiver for BER calculation.

3.6.1 Observations

The following observations are made when the simulation was carried out for DFT OFDM baseband transmission:

1. In the autocorrelation curve if it has two plateaus, sometime it will detect the first plateau and sometime it will second plateau.

2. Estimated integer frequency offset always does not come as per the expectation i.e. around 4±2 for the present case.

3. Estimated fractional frequency offset always does not come as per the expectation i.e. around 0.3±0.2 for the present case.

3.6.2 Analysis

The statistical analysis has been done for the above observations by plotting the histogram for the different values of SNR and 150 numbers of iterations.

1. For SNR=10

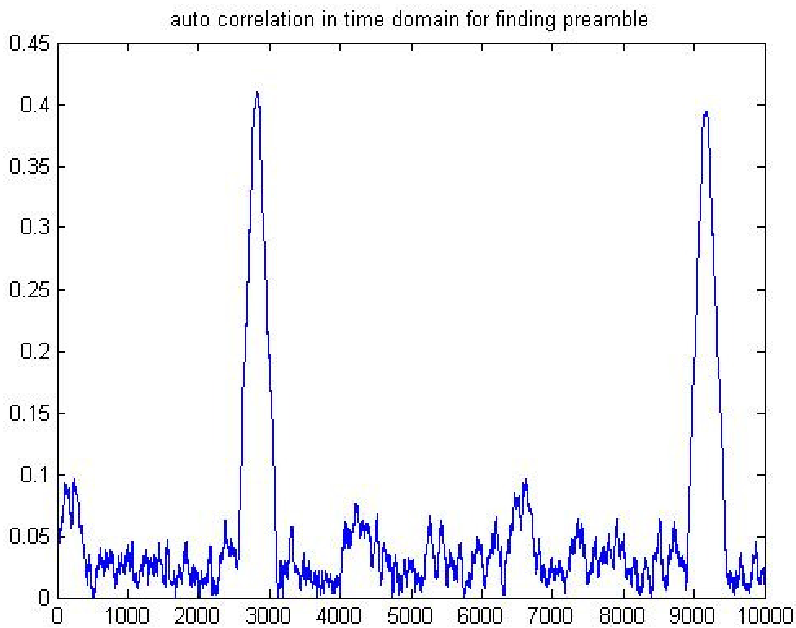

Fig.3.4. Auto Correlation plot in time domain at SNR=10

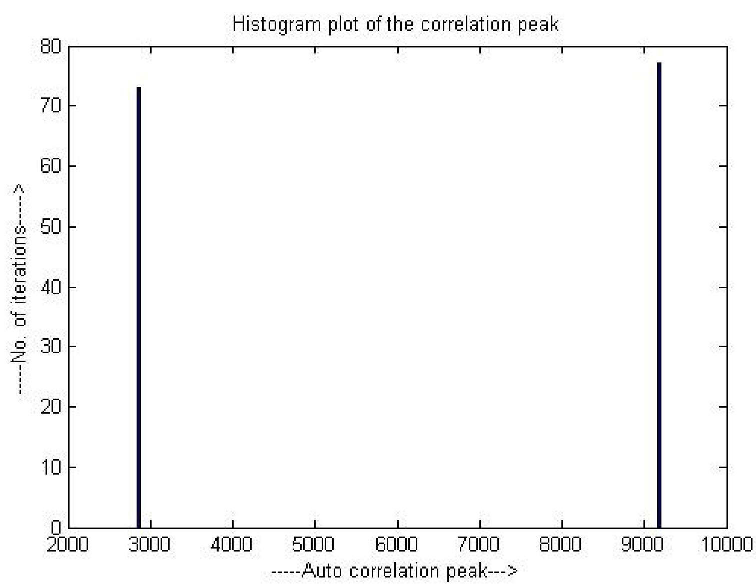

Fig.3.5. Histogram plot at SNR=10 shows that in almost half of the cases it detects first plateau and in rest the second plateau

2. For SNR = 20

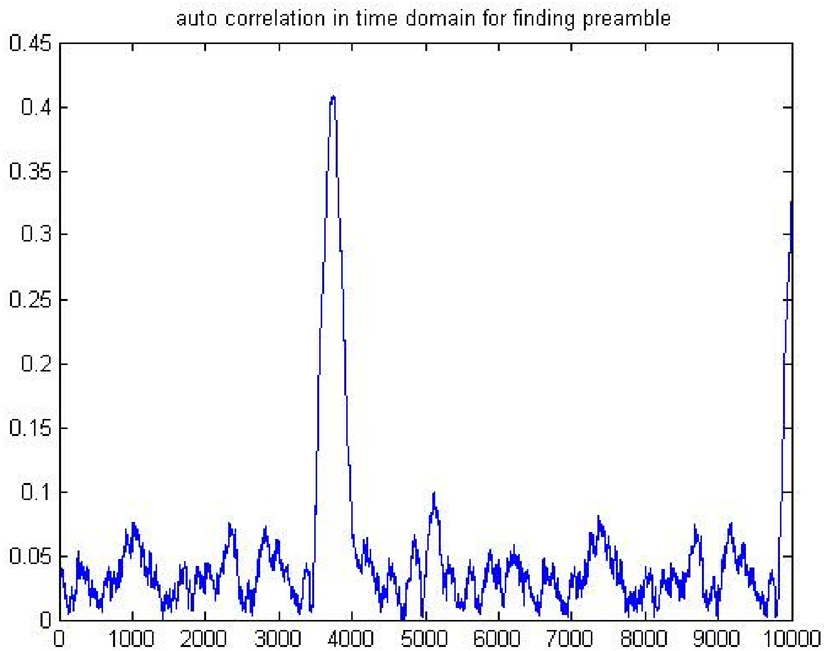

Fig.3.6. Auto Correlation plot in time domain at SNR=20

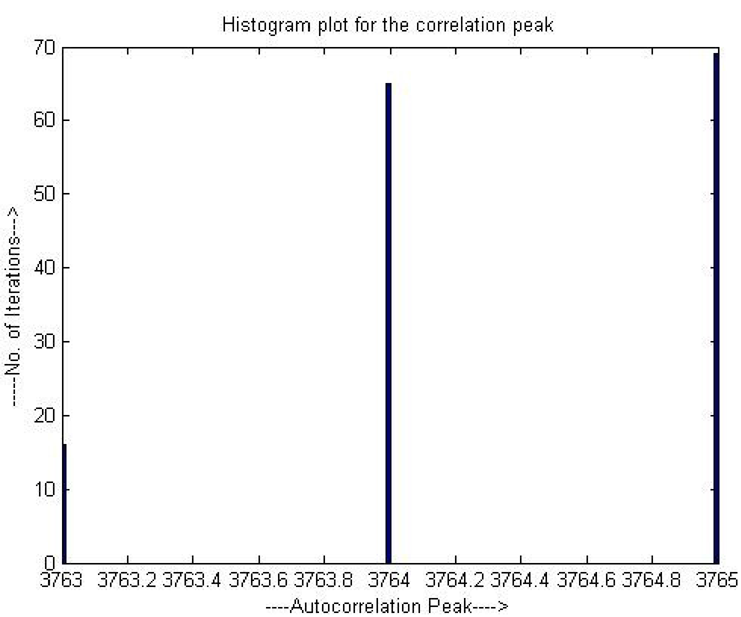

Fig.3.7. Histogram plot at SNR=20 shows that in almost in all the cases it detects first plateau in the range of 3763 to 3765. .

3. For SNR = 30

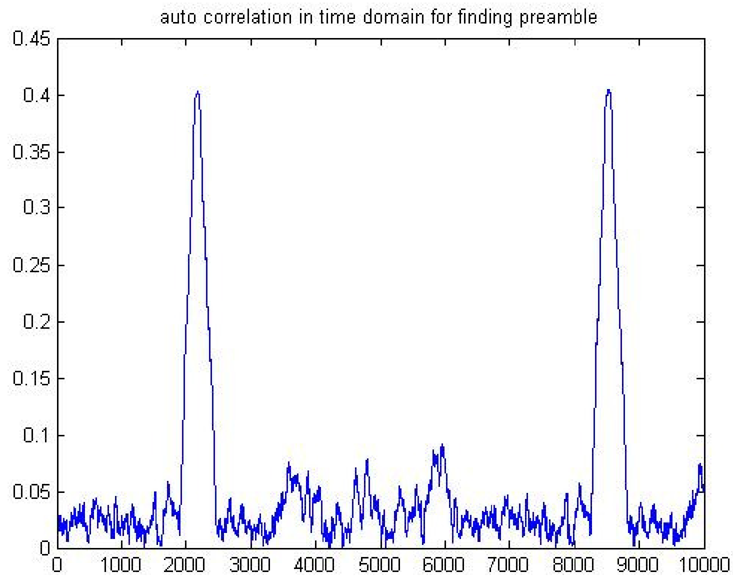

Fig.3.8. Auto Correlation plot in time domain at SNR=30

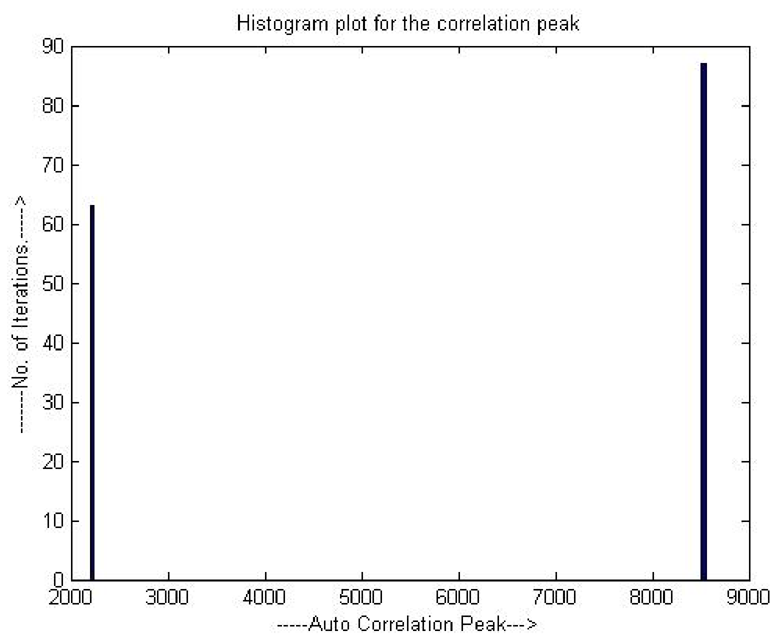

Fig.3.9. Histogram plot at SNR = 30 shows that for than half of the cases it detects second plateau and in rest the first plateau.

Chapter 4

OFDM CHANNEL ESTIMATION

The channel is the medium through which the signal travels from the transmitter to the receiver. Unlike the wired channels that are stationary and deterministic, wireless channels are extremely random in nature. Some of the features of wireless communication like mobility, places fundamental limitations on the performance in wireless system. The transmission path between the transmitter and receiver can vary from line of sight to that is severely obstructed by buildings, terrain & foliage. Efficient channel estimation strategies are required for coherent detection & decoding. Adaptive estimation of the channel is necessary before the demodulation of the OFDM signals since the wireless channel is frequency selective and time-varying. The channel estimation in OFDM can be classified into the two categories

1. Pilot Based Channel Estimation: Known symbol called pilots are transmitted.
2. Blind Channel Estimation: No pilots required. It uses some underlying mathematical properties of data sent.

The Blind channel estimation methods are computationally complex and hard to implement. The Pilot based channel estimation methods are easy to implement but they reduces the bandwidth efficiency. The Pilot based methods are most popular now a days. IEEE 802.16e, 3GP LTE standards support the pilot based channel estimation.

4.1 OFDM System Modeling

Let the cyclic extension of time length T_G, chosen to be larger than the expected delay spread, is inserted to avoid intersymbol and intercarrier interferences. The D/A converter contain low-pass filters with bandwidth $1/T_S$, (as shown in the Fig.2.2) where T_S is the sampling interval.

The channel is modeled as an impulse response g(t) followed by the complex additive white Gaussian noise (AWGN) n(t), where α_m is a complex values and $0 \leq \tau_m T_S \leq T_G$. Hence we treat the cannel response g(t) as a time –pulse train of the form **[4]**

$$g(t) = \sum_{m=1}^{M} \alpha_m \partial(t - \tau_m T_s) \tag{4.1}$$

The entire impulse response lies in the guard space. The system is then modeled using the N-point *discrete-time Fourier transform (DFT_N)* as

$$y = DFT_N(IDFT_N(x) \otimes \frac{g}{\sqrt{N}} + \hat{n}) \tag{4.2}$$

Where $\otimes$ denotes cyclic convolution, x=$[x_0 x_1 --- x_{N-1}]^T$, y=$[y_0 y_1 .... y_{N-1}]^T$, $\hat{n}$ is a vector of i.i.d. complex Gaussian variable, & g= $[g_0 g_1 .... g_{N-1}]^T$ is determined by the cyclic equivalent of sinc functions.

The system described by above equation can be written as a set of N independent Gaussian channels,

$$y_k = h_k x_k + n_k \tag{4.3}$$

Where h_k is the complex channel attenuation given by h=$[h_0 h_1 .... h_{N-1}]^T = DFT_N(g)$ and

n=$[n_0 n_1 .... n_{N-1}]^T = DFT_N(\hat{n})$ is an i.i.d. complex zero mean Gaussian noise vector.

As a matter fact we can write the above equation in the matrix form as

$$y = XFg + n \tag{4.4}$$

Where $X = diag\ (x)$ matrix and

$$F = \begin{bmatrix} W_N^{00} & \cdots & W_N^{0(N-1)} \\ \vdots & \ddots & \vdots \\ W_N^{(N-1)0} & \cdots & W_N^{(N-1)(N-1)} \end{bmatrix} \tag{4.5}$$

is the DFT matrix with

$$W_N^{nk} = \frac{1}{\sqrt{N}} e^{-j2\pi \frac{nk}{N}} \tag{4.6}$$

4.2 Channel Estimation

We will derive several estimators based on the system model in the previous section. These estimation techniques all have the general structure presented in Fig. 4.1. The transmitted symbols x_k, appearing in the estimator expressions, are either training symbols or quantized decision variables in a decision-directed estimator **[4, 5]**.

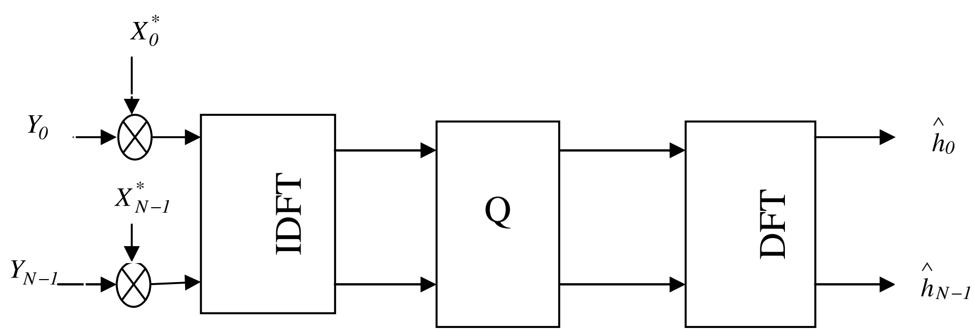

Fig.4.1 General channel Estimator Structure

4.2.1 Least Square/Zero Forcing Channel Estimators

The LS estimator for the cyclic response g minimizes $(y - XFg)^H (y - XFg)$ and generates the

$$\hat{h}_{LS} = FQ_{LS} F^H X^H y \tag{4.7}$$

Where

$$Q_{LS} = (F^H X^H XF)^{-1} \tag{4.8}$$

As the $\hat{h}_{LS}$ also corresponds to the estimator structure shown in the Fig.4.1. So $\hat{h}_{LS}$ reduces to

$$\hat{h}_{LS} = X^{-1} y \tag{4.9}$$

The channel estimates at data subcarriers can be obtained using 1D interpolation. As the spacing between the pilot subcarriers increase, the accuracy of this method drops. This method ignores the frequency domain correlation of the channel.

4.2.2 Modified Least Square Channel Estimators

This is a time domain method of channel estimation. Usually the number of taps in the impulse response of the channel is less than the number of subcarriers in the transfer function. Therefore it is advantageous to estimate the impulse response of the channel than its frequency domain counterpart.

As the performance of LS estimator is low in terms of mean square error, so to improve this we can assume that the most energy is concentrated into the first few samples of the impulse response. Intuitively, excluding low energy taps of g will to some extent compensate for this shortcoming since the energy of g decreases rapidly outside the first L taps, whilst the noise energy is assumed to be constant over the entire range **[4, 6].**

Taking only the first L taps of g into account, thus implicitly using channel statistics, the modified LS estimator becomes

$$\hat{h}_{LS} = TQ'_{LS}T^{H}X^{H}y \tag{4.10}$$

Where T denotes the first L columns of DFT matrix F and

$$Q_{LS} = (T^{H}X^{H}XT)^{-1} \tag{4.11}$$

4.3 PED-B Channel

The transmitted frame is passed through a frequency selective channel. PED-B model is used here as the channel. The number of taps in PED-B (pedestrian channel) is 6.

Tap Positions (μs)	Tap Gain (dB)
1	0
2	-1
5	-9
7	-10
12	-15
19	-20

Table 4.1 PED-B Channel Model

4.4 Observations & Analysis

The following constellation plot has been plotted for DFT OFDM base band transmissions for $N_{FFT}=512$

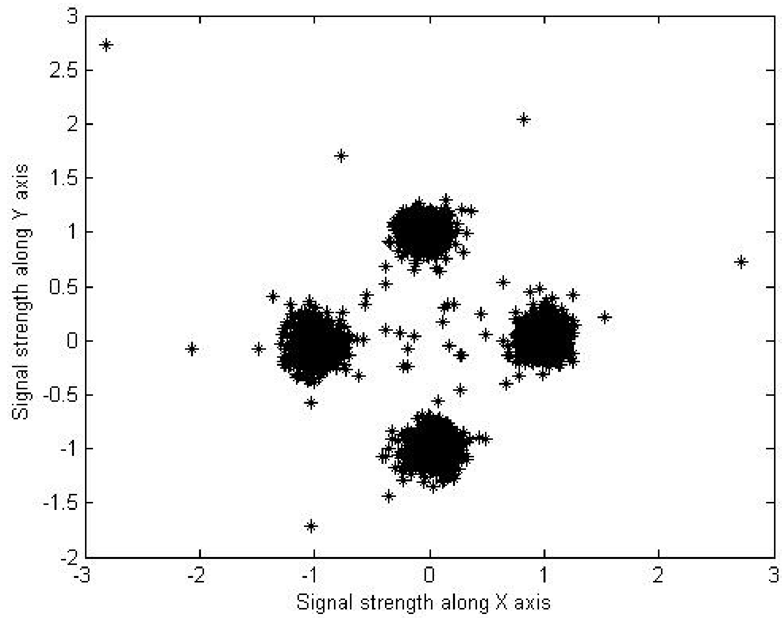

Fig.4.2 Constellation obtained by LS channel estimators for DFT OFDM baseband transmission

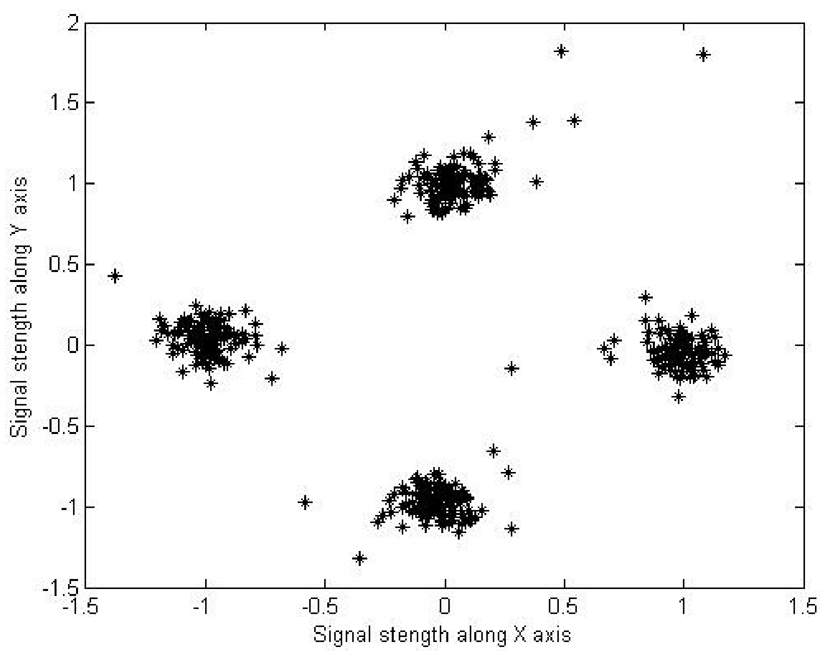

Fig.4.3 Constellation obtained by MLS channel estimator for DFT OFDM baseband transmission

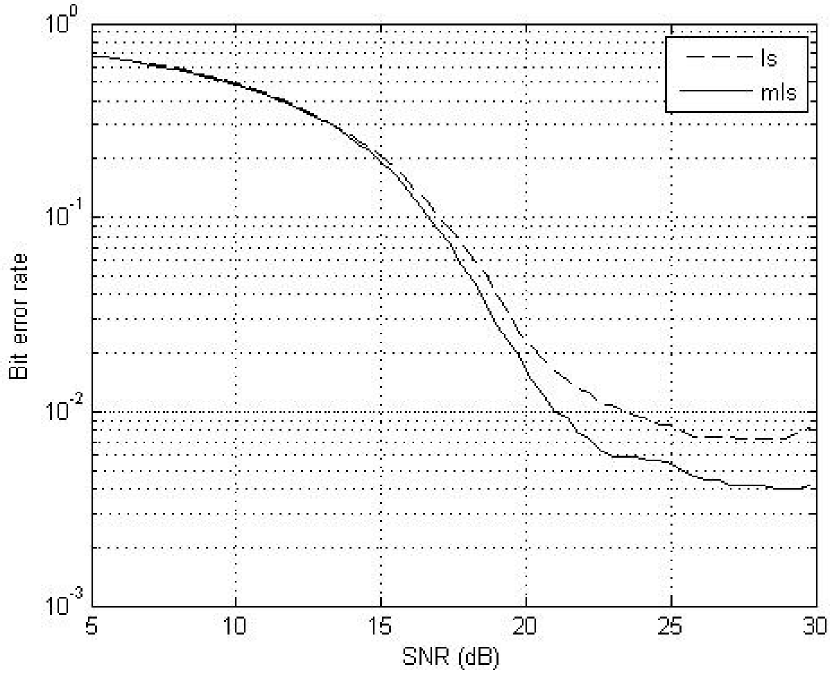

Fig.4.4 BER Versus SNR curve for DFT based OFDM system after channel estimation

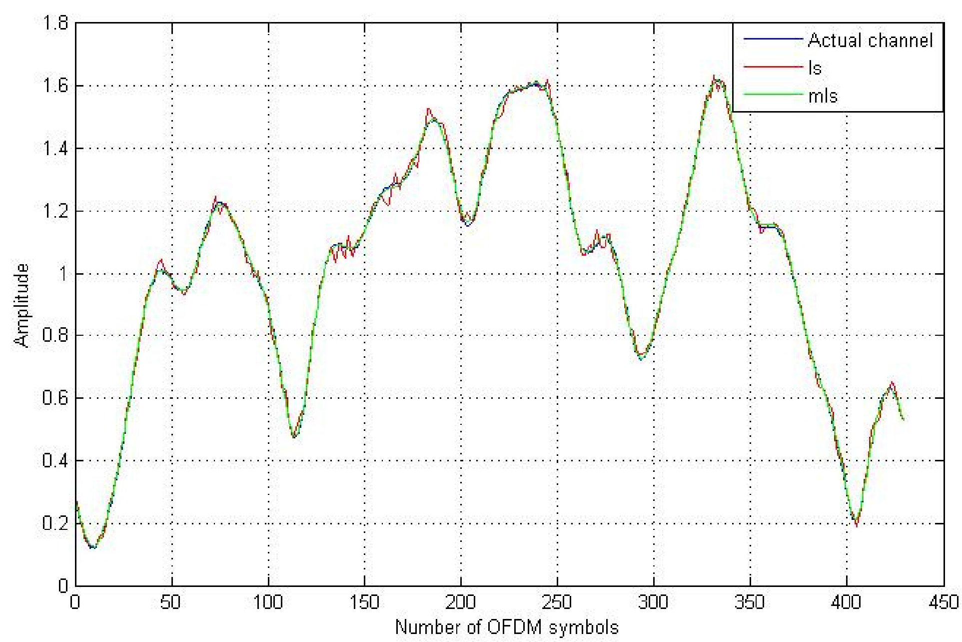

Fig.4.5 Channel estimation plot for DFT based OFDM system

It can be clearly observed from the Fig. 4.2, Fig. 4.3, Fig. 4.4 & Fig 4.5 that constellations & BER versus SNR curve performance for MLS channel estimator is better than LS channel estimator.

Chapter 5

DHT BASED OFDM BASEBAND SYSTEM

For the current DFT-based or OFDM-based transceivers, the modulator needs to compute a long-length inverse discrete Fourier transform (IDFT), and the demodulator needs to compute a long-length DFT, where the transform length is up to 512 or more. For such long-length IDFT/DFT computations, a great number of complex multiplications are required and each of them basically involves four real multiplications and two real additions. Clearly, the complexity of a DFT-based or OFDM-based transceiver would be reduced if the corresponding modulator/demodulator could be implemented using purely real transforms while fast algorithms similar to the fast Fourier transform (FFT) algorithm can still be applied.

In this report, we propose a novel digital form of MCM that is based on the discrete Hartley transform (DHT) and its inverse (IDHT) for modulation and demodulation **[10].** The DHT involves only real-valued arithmetic and has an identical inverse. Like the DFT, there have been a number of fast algorithms and hardware architectures available for the DHT computation.

Hence DHT-based MCM method achieves the same transmission performance as the DFT-based MCM method, but requires less computational complexity **[10, 14]**.

5.1 Some Properties of the DHT

The N-point DHT of a real sequence is defined by **[11]**

$$S_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} s_n H_N^{kn}, k = 0,1,2,\cdots,N-1 \tag{5.1}$$

Where $H_N^k = \sin(2\pi k / N) + \cos(2\pi k / N)$. Letting $s = [s_{N-1,} s_{N-2,} \ldots, s_1, s_0]^T$ be the transform input vector and $S = [S_{N-1}, S_{N-2}, \cdots, S_1, S_0]^T$ be the transform output vector, we can express the above equation in matrix-vector form as follows:

$$S=Hs \tag{5.2}$$

$$H = \frac{1}{\sqrt{N}} \begin{bmatrix} H_N^{(N-1)^2} & H_N^{(N-1)(N-2)} & \cdots & H_N^{N-1} & 1 \\ H_N^{(N-1)(N-2)} & H_N^{(N-2)^2} & \cdots & H_N^{N-2} & 1 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ H_N^{N-1} & H_N^{N-2} & \cdots & H_N^1 & 1 \\ 1 & 1 & \cdots & 1 & 1 \end{bmatrix} \tag{5.3}$$

Being the DHT transform matrix. Note that the DHT has an identical inverse, i.e., the DHT and IDHT transform matrices are the same ($H=H^{-1}$) and we have

$$s=H^{-1}S=HS \tag{5.4}$$

Also

$$Q = H.L \tag{5.5}$$

Where Q is IDFT matrix and is given by

$$Q = \frac{1}{\sqrt{N}} \begin{bmatrix} W_N^{(N-1)^2} & W_N^{(N-1)(N-2)} & \cdots & W_N^{N-1} & 1 \\ W_N^{(N-1)(N-2)} & W_N^{(N-2)^2} & \cdots & W_N^{N-2} & 1 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ W_N^{N-1} & W_N^{N-2} & \cdots & W_N^1 & 1 \\ 1 & 1 & \cdots & 1 & 1 \end{bmatrix} \tag{5.6}$$

And L is a unitary matrix given by

$$
L=\begin{bmatrix}
\frac{1+j}{2} & 0 & \cdots & 0 & \cdots & 0 & \frac{1-j}{2} & 0 \\
0 & \ddots & 0 & \cdots & 0 & \cdot^{\cdot^{\cdot}} & 0 & \vdots \\
\vdots & 0 & \frac{1+j}{2} & 0 & \frac{1-j}{2} & 0 & \vdots & 0 \\
0 & \vdots & 0 & 1 & 0 & \vdots & 0 & \vdots \\
\vdots & 0 & \frac{1-j}{2} & 0 & \frac{1+j}{2} & 0 & \vdots & 0 \\
0 & \cdot^{\cdot^{\cdot}} & 0 & \cdots & 0 & \ddots & 0 & \vdots \\
\frac{1-j}{2} & 0 & \cdots & 0 & \cdots & 0 & a_1 & 0 \\
0 & 0 & \cdots & 0 & \cdots & 0 & 0 & a_0
\end{bmatrix} \tag{5.7}
$$

With L^{-1} equal to the transpose of complex conjugate of L.

As we know received signal in DFT based OFDM system is given by **y**:

$$
y=Px+n \tag{5.8}
$$

Where **P** is a square N×N matrix, **x** is N×1 channel input vector, **y** is the corresponding N×1 channel output vector and n is noise.

P can be rewritten as

$$
P=HL\Omega L^{-1}H^{-1}= HL\Omega L^{-1}H=H\Phi H \tag{5.9}
$$

Where $\Phi=L\Omega L^{-1}$

$$
\Phi=\begin{bmatrix}
a_1 & 0 & \cdots & 0 & \cdots & 0 & -b_1 & 0 \\
0 & \ddots & \ddots & \vdots & \ddots & \cdot^{\cdot^{\cdot}} & 0 & 0 \\
\vdots & \ddots & a_{N/2-1} & 0 & -b_{N/2-1} & \cdot^{\cdot^{\cdot}} & \vdots & \vdots \\
0 & \cdots & 0 & b_0 & 0 & \cdots & 0 & 0 \\
\vdots & \cdot^{\cdot^{\cdot}} & b_{N/2-1} & 0 & a_{N/2-1} & \ddots & \vdots & \vdots \\
0 & \cdot^{\cdot^{\cdot}} & \cdot^{\cdot^{\cdot}} & \vdots & \ddots & \ddots & 0 & 0 \\
b_1 & 0 & \cdots & 0 & \cdots & 0 & a_1 & 0 \\
0 & 0 & \cdots & 0 & \cdots & 0 & 0 & a_0
\end{bmatrix} \tag{5.10}
$$

and

$$\Omega = \begin{bmatrix} a_1 + jb_1 & 0 & \cdots & 0 & \cdots & 0 & 0 & 0 \\ 0 & a_2 + jb_2 & \ddots & \vdots & \ddots & \iddots & 0 & 0 \\ \vdots & \ddots & \ddots & 0 & 0 & \iddots & \vdots & \vdots \\ 0 & \cdots & 0 & b_0 & 0 & \cdots & 0 & 0 \\ \vdots & \iddots & \vdots & 0 & \ddots & \ddots & \vdots & \vdots \\ 0 & \iddots & 0 & \vdots & 0 & a_2 - jb_2 & 0 & 0 \\ 0 & 0 & \cdots & 0 & \cdots & 0 & a_1 - jb_1 & 0 \\ 0 & 0 & \cdots & 0 & \cdots & 0 & 0 & a_0 \end{bmatrix} \tag{5.11}$$

5.2 DHT Based OFDM System and its Implementation

Based upon the properties of DHT described above, we propose a DHT-based OFDM system as follows. Assume that the real data sequence to transmitted, the channel impulse response, and the channel noise are same as those used in DFT based OFDM system. Then we can directly use the data sequence to form a real data vector for modulation, i.e. S=$[S_{N-1}, S_{N-2}, \ldots S_1, S_0]^T$ = $[\alpha_{N-1}, \alpha_{N-2}, \ldots, \alpha_1, \alpha_0]^T$. In contrast to the DFT based OFDM transmission process, the DHT based OFDM can be described as follow

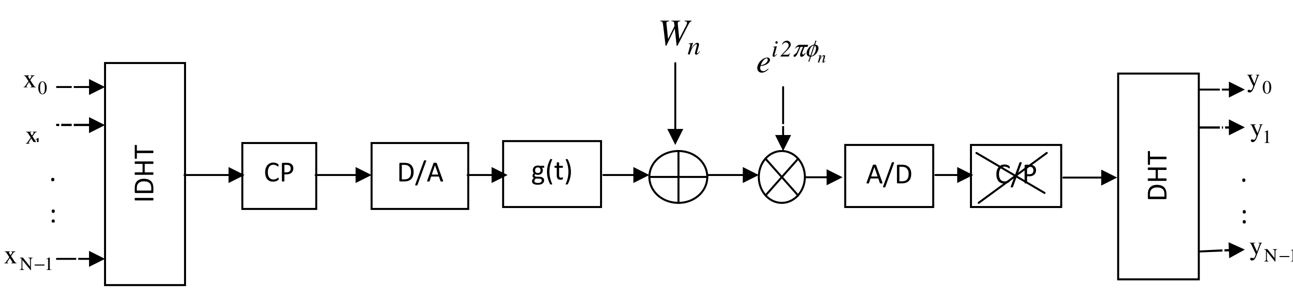

Fig.5.1. Typical DHT based OFDM baseband system, affected only by noise, timing offset and frequency offset. W_n represents AWGN noise, ϕ_n represents a time varying phase which (artificially mimics) implements frequency offset , CP refers to cyclic prefix A/D refers to analog to digital converter and D/A refers to digital to analog converter.

1. *Modulation*: Compute the N-point IDHT of **S,** i.e. $s=\left[s_{N-1,}s_{N-2,}\ldots,s_{1},s_{0}\right]^{T}$ **=HS** (5.12)

2. *Adding a cyclic prefix*: Cyclic prefix is added to form a channel input sequence as given by $s=\left[s_{-v,}s_{-v+1,}\ldots,s_{-1},s_{0}\cdots,s_{N-1}\right]$ where $s_{-k}=s_{N-k}$ for k=1,2,…,v.

3. *Channel Output*: After neglecting the first v sample at the receiver, the received signal vector $r=\left[r_{N-1,}r_{N-2,}\ldots,r_{1},r_{0}\right]^{T}$ and the IDHT modulated s vector have the following relation: (5.13)

   r=Ps+n = HΦHs=HΦS+n

4. *Demodulation:* Compute the N-point DHT of r, i.e. R=$[R_{N-1},R_{N-2},\ldots,R_{1},R_{0}]^{T}$=Hr=ΦS+Hn (5.14)

The demodulated **R** contains two components: the first component **ΦS** is due to the transmitted data and the second one Hn due to additive channel noise.

5. *Equalization and Detection:* To recover the data vector S, appropriate frequency equalization is needed. This can be described by the following matrix-vector multiplication operation:

$$\Phi^{-1}R=S+\Phi^{-1}Hn=S+n^{'} \tag{5.15}$$

In our program we have used Schmidl-Cox algorithm for timing and frequency synchronization.

6. *Channel Estimation:* The channel is estimated by using two algorithms

   a) Least Square Channel Estimation (described in section 4.2.1)

   b) Modified Least Square Channel Estimation (described in section 4.2.2)

Here in this case also we have used PED-B Channel (described in section 4.3)

5.3 Observation & Analysis

5.3.1 Frequency and Timing Synchronization

The observations and analysis for the DHT based OFDM system is same as that of DFT based OFDM system described in the section 3.6.

5.3.2 Channel Estimation

Channel estimation for both the DHT and DFT based OFDM system has been done using least squares (LS) channel estimator and modified least square (MLS) channel estimator and compared with the actual channel. Fig. 2 shows the channel estimation plot for DFT based OFDM system and Fig. 3 shows the channel estimation plot for DHT based OFDM system. It is clear from the Fig. 2 and Fig. 3 shows that LS and MLS channel estimators estimates the channel equally well and better at some points in case of DHT based OFDM system in comparison to DFT based OFDM system.

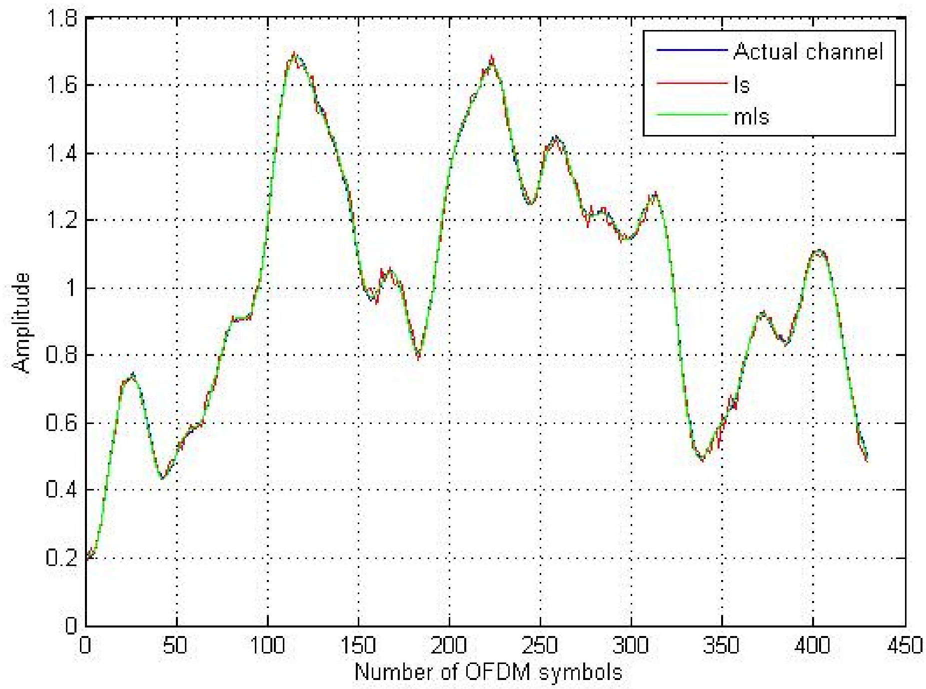

Fig.5.2 Channel estimation plot for DHT based OFDM system

5.3.3 Analysis using BER versus SNR curve

The performance of the proposed DHT based OFDM system has been analysed by plotting bit error rate (BER) versus SNR curve after channel estimation and compared it with that of DFT based OFDM system **[19]**. Fig. 5.3 shows BER versus SNR for DHT based system after channel estimation. . Fig. 4.4 shows BER versus SNR for DFT based system after channel estimation. Table 5.1 shows BER comparison of LS and MLS channel estimator for DHT and DFT based OFDM system at different SNR. It can be observed clearly from the Fig. 4 and Fig. 5 that at different SNR, the bit error rate performance for DHT and DFT based OFDM is almost same or better in case of DHT based OFDM system which justifies the importance of proposed method

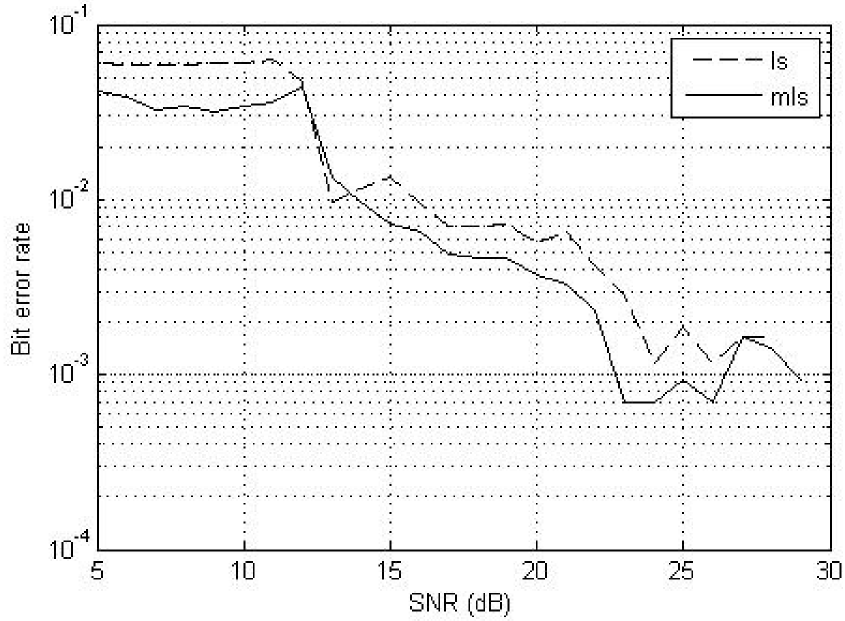

Fig.5.3 BER Versus SNR curve for DHT based system after channel estimation

SNR (dB)	BER in DFT Based OFDM System for Least Squares Channel estimator	BER in DHT Based OFDM System for Least Squares Channel estimator	BER in DFT Based OFDM System for Modified Least Squares Channel estimator	BER in DHT Based OFDM System for Modified Least Squares Channel Estimator
5.0	0.682	**0.06028**	0.682	**0.04029**
10.0	0.4761	**0.6021**	0.4642	**0.03458**
15.0	0.2068	**0.01355**	0.1949	**0.007423**
20.0	0.02196	**0.005607**	0.01572	**0.003738**
25.0	0.008645	**0.001869**	0.005374	**0.0009346**

Table 5.1. Showing BER comparison of LS and MLS channel estimator for DHT and DFT based OFDM system at different SNRs

Chapter 6

CONCLUSIONS & FUTURE WORK

5.1 Conclusions

In the wireless communication system it is very important to transfer the data at a very high rate with sufficient robustness and with less computationally complexites. We have simulated DHT based OFDM system. The DHT is real valued transform and having identical inverse while the DFT is a complex transform and not having identical inverse. So the DHT based OFDM is having less compuational complexities than DFT based OFDM system. As DHT is having identical inverse so implementation of DHT based results in reduction in hardware requirement as the same hardware can be used for inverse DHT on receiver side. The simulation results shows that DHT based OFDM system is having same transmission efficiency, better BER performance as that of DFT based OFDM system with less computational complexities & hardware requirement.

5.2 Future Work

The proposed methods of OFDM synchronization & channel estimation has been simulated for DFT based OFDM system and DHT based OFDM system. Further we have studied the application of OFDM in optical communication and theoretically found that like the wireless communication in optical communication also it offers efficient and fast trasmission of data with low requirement of bandwidth. The future works includes the hardware system implementation of the proposed scheme for both wireless & optical communication.

PUBLICATIONS

Publication Based on the Thesis

[1] Vijay Kumar, G. Panda, P. K. Sahu, "DHT Based 4 QAM OFDM Baseband System and Channel Estimation", **Accepted** for publication in *2nd IEEE International Conference on Computer Science and Information Technology (IEEE ICCSIT 2009)*, 8 - 11, August 2009, Beijing, China.

Other Publications

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