CHAPTER 1

INTRODUCTION

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1.1 INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is a popular modulation technique because of its spectral efficiency and robustness to frequency selective channels. It is well-known that OFDM systems are sensitive to the carrier frequency offset (CFO). CFO destroys the orthogonality between subcarriers and induces inter-carrier interference. Therefore, if the CFO is not corrected at the receiver, the performance of OFDM systems can degrade seriously. In the past, many methods have been proposed for CFO estimation . These methods can be divided into two groups: dataaided and non data-aided (blind) schemes. This project focuses on blind CFO estimation.

In the authors utilize the frequency analysis (FA) for blind CFO estimation, but it works only for the constant modulus constellations. Other blind CFO estimation schemes for non constant modulus modulation have been reported in literatures. In a kurtosis-based cost function was proposed for blind CFO estimation, and its complexity is quite low. Another simple method based on cyclic prefix (CP) was proposed in , but it suffers performance degradation in multipath environment. Several blind CFO estimation methods employing the null subcarriers or the virtual carriers (VC) have been developed. The multiple signal classification (MUSIC) method was proposed in which requires a one-dimensional search and its complexity depends on the search resolution. The ESPRIT-based scheme has a closed form solution, but its performance is highly dependent on the channel impulse response.

In this project, we propose a new blind CFO estimation for OFDM systems based on the so-called remodulated received vectors . A closed form formula is derived. The proposed method works for both constant and non constant modulation symbols, and for OFDM systems with or without VCs. Compared with existing methods, the proposed method achieves a better results at a fraction of their complexity.

1.2 BACKGROUND

With the rapid growth of Digital Communications in recent years the need for high speed data transmission is increased. OFDM is a promising solution for achieving high data rates in mobileenvironment, due to its resistance to ISI which is a common problem found in high speed data communication.

As Orthogonal Frequency Division Multiplexing (OFDM) is considered under development of LTE(long term evolution) which is commonly used in wireless communication systems due to its key advantages such as its too flexibility as well as bandwidth efficiency. The CFO estimate is given by a closed form formula. The proposed method has very low complexity and its performance is robust to different modulation symbols and the presence of virtual carriers.

1.3 OBJECTIVES

- To perform a low complexity blind cfo estimation for ofdm systems.
- To reduce inter symbol interference (ISI).
- Co – channel interference reduction.
- It works both for constant and non constants modulation symbols.

CHAPTER 2

LITERATURE REVIEW

CHAPTER 2

LITERATURE SURVEY

“A Novel Approach for Blind CFO Estimation in OFDM Systems ”Shivkanya Vishnu Doke Department of Electronics & Telecommunication Shree Ramchandra College of Engineering, Pune, Dr. K. Sujatha Department of Electronics & Telecommunication Shree Ramchandra College of Engineering, Pune,2019

The wireless communication has effectively contributed to the development of countries like India in last decade. India is looking forward to use the 5G technology to strengthen the present communication infrastructures. The rate of data transfer plays a vital role when we think of a wireless technique for communication. Authors have untaken the experimentation for learning of the carrier frequency offset based on the cyclic prefix. The implemented method is found less complex and robust than the timing synchronization method.

“Estimation And Reduction Of CFO In OFDM System ”Prof. A.R.Khedkar Prachi Admane MKSSS’S Cummins COE For Women,Karvenagar,Pune MKSSS’S Cummins COE For Women,Karvenagar,Pune Dept. of Electronics And Telecommunication Engineering Dept. of Electronics And Telecommunication Engineering,june 2016.

Orthogonal Frequency Division Multiplexing (OFDM) is an emerging multi-carrier modulation scheme. OFDM is sensitive to the carrier frequency offset (CFO) which may be caused due to difference in the carrier frequencies of transmitter and receiver. In this project work, we have done the estimation of CFO using pilot based.

"CFO and Channel Estimation for MISO-OFDM Systems "Abdelhamid Ladaycia, Karim Abed-Meraim, Ahmed Bader and M.S. Alouini,2017

This study deals with the joint channel and carrier frequency offset (CFO) estimation in a Multiple Input Single Output (MISO) communications system. This problem arises in OFDM (Orthogonal Frequency Division Multiplexing) based multi-relay transmission protocols such that the geo-routing. In this work, two approaches are considered: The first is based on estimating the overall channel (including the CFO) as a timevarying one using an adaptive scheme under the assumption of small or moderate CFOs while the second one performs separately, the channel and CFO parameters estimation based on the considered data model. The two solutions are analyzed and compared in terms of performance, cost and convergence rate.

M. Morelli and M. Moretti, "Carrier Frequency Offset Estimation for OFDM Direct-Conversion Receivers," IEEE Transactions on Wireless Communications, vol. 11, no. 7, pp. 2670-2679, July 2012.

We investigate the problem of carrier frequency offset (CFO) recovery in an OFDM direct-conversion receiver plagued by both dc-offset and frequency-selective I/Q imbalance. In order to enlarge the frequency acquisition range, the CFO is divided into an integer part, which is multiple of the subcarrier spacing, plus a remaining fractional part. The fractional CFO is firstly estimated by resorting to the least-squares (LS) principle using a suitably designed training sequence. Since the exact LS solution requires a complete search over the frequency uncertainty range, we propose a simpler scheme that dispenses from any peak-search procedure. We also derive an approximated closed-form expression of the estimation accuracy that reveals useful for assessing the impact of various design parameters on the system performance. After computing the fractional CFO, the integer frequency error is eventually retrieved by following a weighted LS approach. Numerical simulations and theoretical analysis indicate that the proposed scheme can be used to obtain accurate CFO estimates with affordable complexity.

L. Wu, X-D. Zhang, P-S. Li, and Y-T. Su, "A closed-form blind CFO estimator based on frequency analysis for OFDM systems," IEEE Transactions on Wireless Commun., vol. 57, no.6, pp. 1634-1637, June 2009.

This proposes a blind carrier frequency offset (CFO) estimator for orthogonal frequency-division multiplexing (OFDM) systems based on the frequency analysis of the received signal, and derives a closed-form CFO estimate. Since only one OFDM symbol is utilized instead of multi-symbol averaging, the proposed method is effective even when the CFO is time varying. Finally, analysis and simulation results indicate the outstanding performance of the proposed estimator.

Y. Yao and G. B. Giannakis, "Blind Carrier Frequency Offset Estimation in SISO, MIMO, and Multiuser OFDM Systems," IEEE Transactions on Communications, vol. 53, no. 1, pp. 173-183, Jan. 2005

Relying on a kurtosis-type criterion, we develop a low-complexity blind carrier frequency offset (CFO) estimator for orthogonal frequency-division multiplexing (OFDM) systems. We demonstrate analytically how identifiability and performance of this blind CFO estimator depend on the channel's frequency selectivity and the input distribution. We show that this approach can be applied to blind CFO estimation in multi-input multi-output and multiuser OFDM systems. The issues of channel nulls, multiuser interference, and effects of multiple antennas are addressed analytically, and tested via simulations.

F. Gao, Y. Zeng, A. Nallanathan, and T-S. Ng, "Robust Subspace Blind Channel Estimation for Cyclic Prefixed MIMO OFDM Systems: Algorithm, Identifiability and Performance Analysis," IEEE Journal on Selected Areas in Communications, vol. 26, no. 2, pp. 378-388, Feb. 2008.

A novel subspace (SS) based blind channel estimation method for multi-input, multi-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems is proposed in this work. With an appropriate re-modulation on the received signal blocks, the SS method can be effectively applied to the cyclic prefix (CP) based MIMO-OFDM system when the number of the receive antennas is no less than the number of transmit antennas. These features show great compatibility with the coming fourth generation (4G) wireless communication standards as well as most existing single-input single-output (SISO) OFDM standards, thus allow the proposed algorithm to be conveniently integrated into practical applications. Compared with the traditional SS method, the proposed algorithm exhibits many advantages such as robustness to channel order over-estimation, capability of guaranteeing the channel identifiability etc. Analytical expressions for the mean-square error (MSE) and the approximated Cramer-Rao bound (ACRB) of the proposed algorithm are derived in closed forms. Various numerical examples are conducted to corroborate the proposed studies.

CHAPTER 3

SYSTEM MODEL

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SYSTEM MODEL

Every concept in system theory can be viewed differently depending upon the system environment, in which it is used, and also based on how it is processed in a given algorithm. For example if we take the process of convolution, it could be done for a continuous or a discrete signal, in time or frequency domain. The system impulse response, with which the input signal is convoluted, can be used to represent an electronic circuit, a human vocal fold, a room environment or the echos as in multi-path signal etc. This chapter introduces some of the concepts of system theory in the view of Orthogonal Frequency Division Multiplexing (OFDM) communication system and synchronization of such OFDM systems.

3.1 Linear Time Invariant systems

Most of the systems in communication are assumed to be Linear Time Invariant (LTI) systems. These are the systems which satisfy the linearity property and the time invariance property. These properties will be defined and explained in this section.

3.1.1 Linearity Property

The general class of systems can be subdivided into linear systems and nonlinear systems. A system that satisfies the superposition principle is called linear system. Superposition principle requires that the response of the system to a weighted sum of the signals (inputs) be equal to the corresponding weighted sum of the responses (output) of the system to each of the individual input signals.

Let $x1(n)$ and $x2(n)$ be two arbitrary input sequences, and a1 and a2 be two arbitrary constants. Let T represent the system under consideration. Then T is linear if and only if,

$$T [a1x1(n) + a2x2(n)] = a1\, T [x1(n)] + a2\, T [x2(n)].$$

In the equation (3.1), if $a2 = 0$, then,

$$T [a1\, x1 (n)] = a1T [x1 (n)]$$

equation (3.2) shows the multiplicative property of a linear system. That is, if the input to a system is scaled by a factor a1 , then the corresponding response is also scaled by a1 .

In the equation (3.1), if $a1 = a2 = 1$,then

$$T [x1 (n) + x2 (n)] = T [x1(n)] + T [x2(n)].$$

This equation (3.3) shows the additivity property of a linear system. That is, the response of the system to sum of the signals is equal to the corresponding sum of the response of the system to each of the individual signals. The additivity and multiplicative properties constitute the superposition principle as it applies to linear systems.

3.1.2 Time-invariance property

A system is called time-invariant or shift invariant, if its input-output characteristics do not change with time. Let $y(n)$ be the response of the system T to the input signal $x(n)$, represented as,

$$y(n) = T [x(n)]$$

then if the same input signal is shifted by k units of time to yield, $x(n - k)$, and applied to the system (time invariant system), then the response of the system should also be delayed by k units to yield $y(n - k)$. That is,

$$T\,[x(n-k)] = y(n-k)$$

for the system to be time invariant, the equation (2.5) should hold for all possible values of k. If it does not hold even for one value of k, then the system is time variant.

3.2 Impluse response of a LTI system

Linear time-invariant systems (LTI systems) are a class of systems used in signals and systems that are both linear and time-invariant. Linear systems are systems whose outputs for a linear combination of inputs are the same as a linear combination of individual responses to those inputs. Time-invariant systems are systems where the output does not depend on when an input was applied. These properties make LTI systems easy to represent and understand graphically.

LTI systems are superior to simple state machines for representation because they have more memory. LTI systems, unlike state machines, have a memory of past states and have the ability to predict the future. LTI systems are used to predict long-term behavior in a system. So, they are often used to model systems like power plants. Another important application of LTI systems is electrical circuits. These circuits, made up of inductors, transistors, and resistors, are the basis upon which modern technology is built.

LTI systems are those that are both linear and time-invariant.

Linear systems have the property that the output is linearly related to the input. Changing the input in a linear way will change the output in the same linear way.

$$T[a1x1(t)+a2x2(t)]=a1T[x1(t)]+a2T[x2(t)]=a1y1(t)+a2y2(t),$$

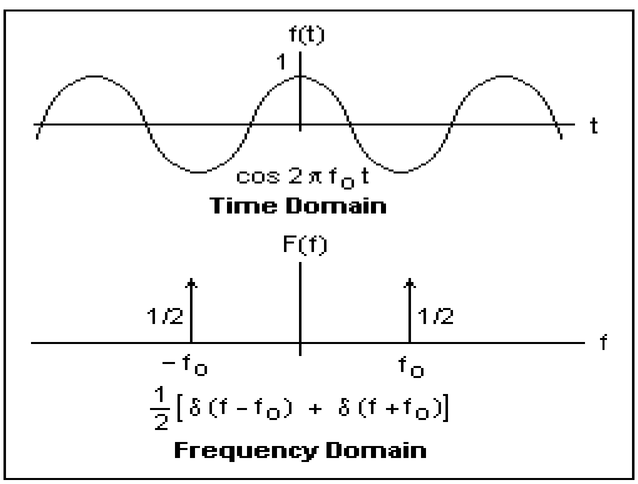
where a1 and a2 are constants.

Further, the output of a linear system for an input of 0 is also 0.

Time-invariant systems are systems where the output for a particular input does not change depending on when that input was applied. A time-invariant systems that takes in signal x(t) and produces output y(t)will also, when excited by signal x(t+σ), produce the time-shifted output y(t+σ).

Thus, the entirety of an LTI system can be described by a single function called its impulse response. This function exists in the time domain of the system. For an arbitrary input, the output of an LTI system is the convolution of the input signal with the system's impulse response.

Conversely, the LTI system can also be described by its transfer function**.** The transfer function is the Laplace transform of the impulse response. This transformation changes the function from the time domain to the frequency domain. This transformation is important because it turns differential equations into algebraic equations, and turns convolution into multiplication. In the frequency domain, the output is the product of the transfer function with the transformed input. The shift from time to frequency is illustrated in the following figure 3.2:



In addition to linear and time-invariant, LTI systems are also memory systems, invertible, casual, real, and stable. That means they have memory, they can be inverted, they depend only on current and past events, they have fully real inputs and outputs, and they produce bounded output for bounded input.

3.2.1 The Impulse Response

The impulse response is an especially important property of any LTI system. We can use it to describe an LTI system and predict its output for any input. To understand the impulse response, we need to use the unit impulse signal, one of the signals described in the Signals and Systems wiki. It has many important applications in sampling. The unit impulse signal is simply a signal that produces a signal of 1 at time = 0. It is zero everywhere else.

3.3.2 Convolution

Convolution is a representation of signals as a linear combination of delayed input signals. In other words, we're just breaking down a signal into the inputs that were used to create it. However, it is used differently between discrete time signals and continuous time signals because of their underlying properties. Discrete time signals are simply linear combinations of discrete impulses, so they can be represented using the convolution sum. Continuous signals, on the other hand, are continuous. Much like calculating the area under the curve of a continuous function, these signals require the convolution integral**.**

Convolution has many important properties:

1. Commutativity: $x(t)*h(t)=h(t)*x(t)$
2. Associativity: $[x(t)*h1(t)]*h2(t)=x(t)*[h1(t)*h2(t)]$
3. Distributivity of Addition: $x(t)*[h1(t)+h2(t)]=x(t)*h1(t)+x(t)*h2(t)$
4. Identity Element: $x(t)*h(t)=h(t)$

3.3 Fast Fourier Transform

A FFT is an algorithm that computes the discrete Fourier transform (DFT) of a sequence, or its inverse (IDFT). Fourier analysis converts a signal from its original domain (often time or space) to a representation in the frequency domain and vice versa. The DFT is obtained by decomposing a sequence of values into components of different frequencies. This operation is useful in many fields, but computing it directly from the definition is often too slow to be practical. An FFT rapidly computes such transformations by factorizing the DFT matrix into a product of sparse (mostly zero) factors. As a result, it manages to reduce the complexity of computing the DFT from , which arises if one simply applies the definition of DFT, to , where is the data size. The difference in speed can be enormous, especially for long data sets where N may be in the thousands or millions. In the presence of round-off error, many FFT algorithms are much more accurate than evaluating the DFT definition directly or indirectly. There are many different FFT algorithms based on a wide range of published theories,from simple complex-number arithmetic to group theory and number theory.

Property / Pair	Signal	FT in f	FT in ω
Linearity	$ax_1(t)+bx_2(t)$	$aX_1(f)+bX_2(f)$	$aX_1(\omega)+bX_2(\omega)$
Time delay	$(xt-t_0)$	$(X)fe^{-j2\pi f t_0}$	$(X)\,\omega e^{-j\omega t_1}$
Frequency Translation	$x(t)e^{j2\pi f_1 t}$	$X(f-f_0)$	$X(\omega-\omega_0)$
Convolution	$x_1(t)*x_2(t)$	$X_1(f)\cdot X_2(f)$	$X_1(\omega)\cdot X_2(\omega)$
Multiplication	$x_1(t)\cdot x_2(t)$	$X_1(f)*X_2(f)$	$\frac{1}{2\pi}X_1(\omega)*X_2(\omega)$
Parseval' s Theorem	$\int_{-\infty}^{\infty}\lvert x(t)\rvert^2\,dt$	$\int_{-\infty}^{\infty}\lvert X(f)\rvert^2\,df$	$\frac{1}{2\pi}\int_{-\infty}^{\infty}\lvert X(\omega)\rvert^2\,d\omega$
Rectangle	$\Pi\left(\frac{t}{\tau}\right)$	$\tau\,\mathrm{sinc}(f\tau)$	$\tau\,\mathrm{sinc}\left(\frac{\omega\tau}{2\pi}\right)$
sinc( )	$2W\,\mathrm{sinc}(2Wt)$	$\Pi\left(\frac{f}{2W}\right)$	$\Pi\left(\frac{\omega}{4\pi W}\right)$
Triangle	$\Lambda\left(\frac{t}{\tau}\right)$	$\tau\,\mathrm{sinc}^2(f\tau)$	$\tau\,\mathrm{sinc}^2\left(\frac{\omega\tau}{2\pi}\right)$
Exponential	$e^{-at}(u),t \quad a>0$	$\frac{1}{a+j2\pi f}$	$\frac{1}{a+j\omega}$
Impulse	$A\delta(t)$	A	A
Constant	A	$A\delta(f)$	$2\pi A\delta(\omega)$
Complex exponential	$e^{j2\pi f_1 t}$	$\delta(f-f_0)$	$2\pi\delta(\omega-\omega_0)$

Assume that $x_1(t)$ and $x_2(t)$ have FTs $X_1(f)$ and $X_2(f)$ respectively.

Figure 3.3 Properties of FFT

Summary

In this chapter the basic concepts of signals and systems are introduced. Most of the systems in communication are assumed to be linear time invariant. This assumption helps to model the system using the impulse response of the system. Knowing the impulse response, we can find the response of the system to any arbitrary input signal. This response to arbitrary input signal is obtained by convolution of the input signal with the impulse response of the system.

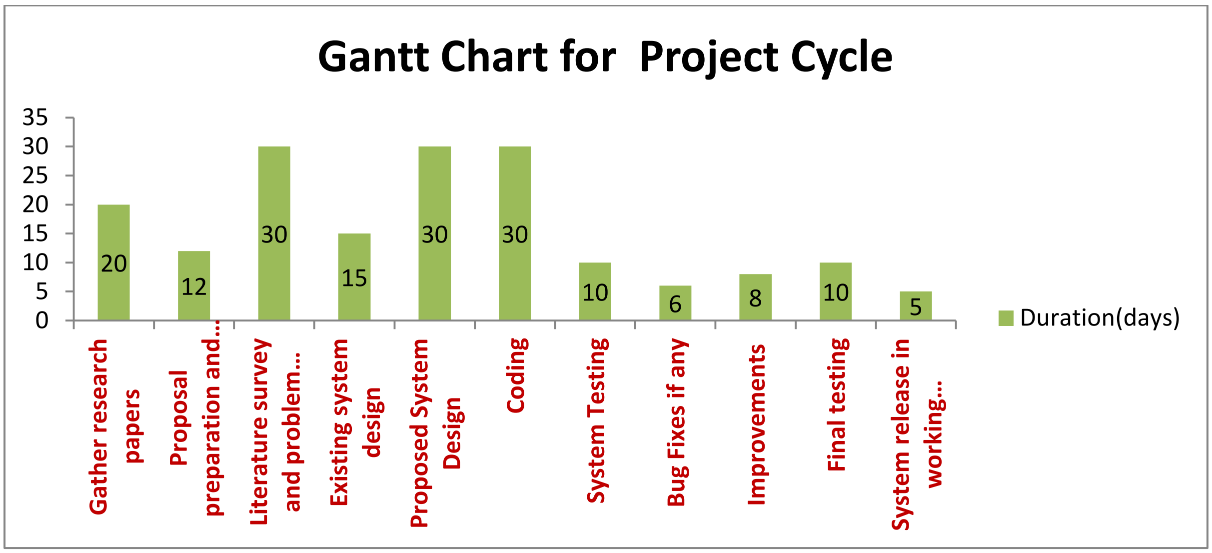
CHAPTER 4

IMPLEMENTATION

4.1 SCHEDULE

Gather research papers	20	15-07-2019	08-05-2019
Proposal preparation and submission	12	08-06-2019	17-08-2019
Literature survey and problem study	30	09-02-2019	10-02-2019
Existing system design	15	10-03-2019	18-10-2019
Proposed System Design	30	12-02-2019	01-02-2020
Coding	30	01-03-2020	02-03-2020
System Testing	10	02-04-2020	14-02-2020
Bug Fixes if any	6	15-02-2020	20-02-2020
Improvements	8	21-02-2020	28-02-2020
Final testing	10	03-01-2020	03-10-2020
System release in working condition	5	03-11-2019	15-03-2020

4.2 GANTT CHART



CHAPTER 5

OFDM

CHAPTER 5

OFDM

OFDM (ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING):

It is a form of signal waveform or modulation that provides some significant advantages for data links.Accordingly, OFDM, Orthogonal Frequency Division Multiplexing is used for many of the latest wide bandwidth and high data rate wireless systems including Wi-Fi, cellular telecommunications and many more.The fact that OFDM uses a large number of carriers, each carrying low bit rate data, means that it is very resilient to selective fading, interference, and multipath effects, as well providing a high degree of spectral efficiency.Early systems using OFDM found the processing required for the signal format was relatively high, but with advances in technology, OFDM presents few problems in terms of the processing required.

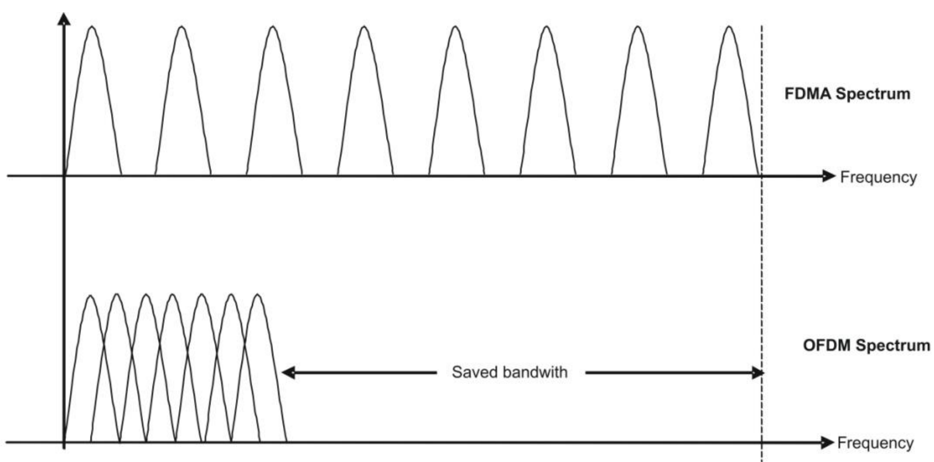


Figure 5.1 Difference between FDMA and OFDM

5.1 DEVELOPMENT OF OFDM

The use of OFDM and multicarrier modulation in general has come to the fore in recent years as it provides an ideal platform for wireless data communications transmissions.However the concept of OFDM technology was first investigated in the 1960s and 1970s during research into methods for reducing interference between closely spaced channels. IN addition to this other requirements needed to achieve error free data transmission in the presence of interference and selective propagation conditions.Initially the use of OFDM required large levels of processing and accordingly it was not viable for general use.Some of the first systems to adopt OFDM were digital broadcasting - here OFDM was able to provide a highly reliable form of data transport over a variety of signal path conditions. Once example was DAB digital radio that was introduced in Europe and other countries. It was Norwegian Broadcasting Corporation NRK that launched the first service on 1st June 1995. OFDM was also used for digital television.Later processing power increased as a result of rising integration levels enabling OFDM to be considered for the 4G mobile communications systems which started to be deployed from around 2009. Also OFDM was adopted for Wi-Fi and a variety of other wireless data systems.

5.2 WHAT IS OFDM?

OFDM is a form of multicarrier modulation. An OFDM signal consists of a number of closely spaced modulated carriers. When modulation of any form - voice, data, etc. is applied to a carrier, then sidebands spread out either side. It is necessary for a receiver to be able to receive the whole signal to be able to successfully demodulate the data. As a result when signals are transmitted close to one another they must be spaced so that the receiver can separate them using a filter and there must be a guard band between them. This is not the case with OFDM. Although the sidebands from each carrier overlap, they can still be received without the interference that might be expected because they are orthogonal to each another. This is achieved by having the carrier spacing equal to the reciprocal of the symbol period.

5.3 KEY FEATURES OF OFDM

The OFDM scheme differs from traditional FDM in the following interrelated ways:

- Multiple carriers (called subcarriers) carry the information stream
- The subcarriers are orthogonal to each other.

- A guard interval is added to each symbol to minimize the channel delay spread and intersymbol interference.

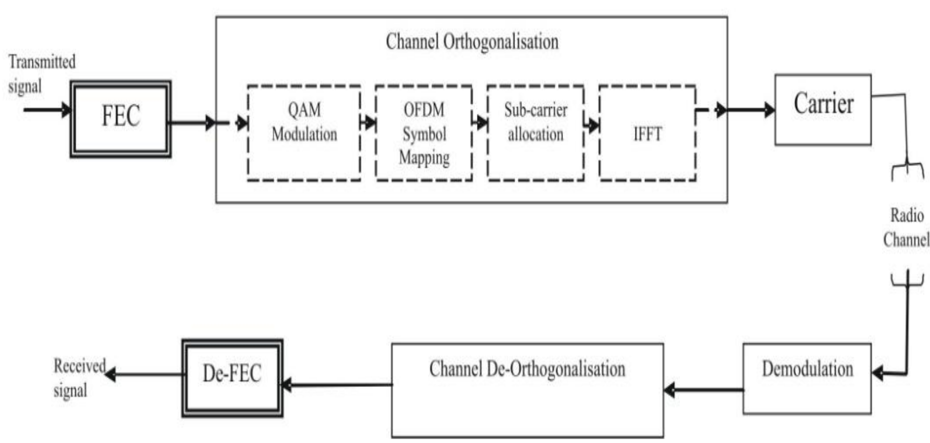
5.4 DATA ON OFDM

- The traditional format for sending data over a radio channel is to send it serially, one bit after another. This relies on a single channel and any interference on that single frequency can disrupt the whole transmission.
- OFDM adopts a different approach. The data is transmitted in parallel across the various carriers within the overall OFDM signal. Being split into a number of parallel "substreams" the overall data rate is that of the original stream, but that of each of the substreams is much lower, and the symbols are spaced further apart in time.

- This reduces interference among symbols and makes it easier to receive each symbol accurately while maintaining the same throughput.

- The lower data rate in each stream means that the interference from reflections is much less critical. This is achieved by adding a guard band time or guard interval into the system. This ensures that the data is only sampled when the signal is stable and no new delayed signals arrive that would alter the timing and phase of the signal. This can be achieved far more effectively within a low data rate substream.

5.5 Block diagram of OFDM systems



OFDM system block architecture can be divided into 3 main sections, see Figure 4.5, namely the transmitter, the channel and the receiver. The model used in this thesis is tested without the using the Forward Error Correction (FEC) coding

5.5.1 Transmitter

Forward Error Correction (FEC)

FEC coding is a technique that improves digital channel quality through the ad-dition of redundant data (parity bits) at the sending node. The redundant data is then decoded at the receiver to detect and correct errors. It is used in OFDM systems to reduce error probability; it trades off data rate and error probability.

There are several coding schemes including Convolution Codes (CC), Binary Turbo Codes (TC) and Turbo Trellis Coded Modulation (TTCM) for an OFDM modu-lated transmission. In our thesis, FEC is not considered in detail.

Quadrature Amplitude Modulation (QAM) and Symbol Mapping

Many transmitted signals are analog or time continuous in nature, such signals can be converted to their digital or discrete form in order to improve the noise resistance. Consequently many transmission systems make of use digital modulation techniques. QAM is one of such techniques; it utilizes the mathematical property that input signals are divided and carried on different components of a single frequency carrier wave, and at the receiver they are resolved successfully into inputs. It is also desirable for high data rate performance .The different QAM modulation schemes commonly used are, e.g., 4-QAM (QPSK), 8-QAM, 16-QAM, 32-QAM and 64-QAM with gray coding in the constellation map. QPSK would be used for simulation purposes in this thesis project, unless other-wise stated. The coded bits (uncoded, if FEC is not used) are then mapped to the constellation map to form the data symbols.

IFFT

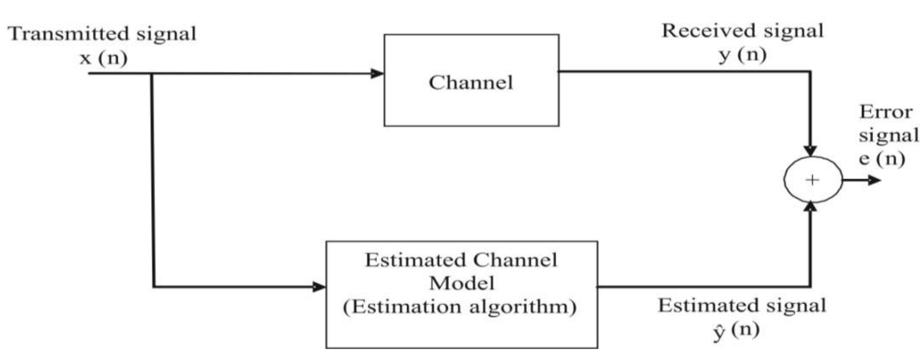
After the OFDM symbols are stacked up into the frame, it is then converted to time domain using the Inverse Discrete Fourier Transform (IDFT). An efficient way of implementing IDFT is IFFT (Inverse Fast Fourier Transform). IFFT is useful for OFDM because it generates samples of a waveform with frequency components satisfying orthogonality conditions, i.e., the IFFT modulates each sub-channel onto a precise orthogonal carrier .

5.5.2 Receiver

The receiver part of the OFDM system will not be discussed in details . This would focus on the estimation of the carrier frequency offset before the signal is processed through all the different blocks at the receiver side.A typical OFDM receiver first removes the cyclic prefix (added to prevent ISI). The data is passed through the serial to parallel converter and then fed to the Fourier Frequency Transform (FFT) for frequency domain transformation. In order to reconstruct the original data from the received data (distorted by the channel), channel estimation and interpolation operations are performed. For this purpose the pilot sub-carriers are used. In this thesis several estimation methods would be mentioned but the Maximum Likelihood (ML) estimation of OFDM carrier frequency offset would be explored .

Channel Estimation

Channel estimation is the process of characterizing or analysing the effect of the physical medium on the input sequence (transmitted data). The basic channel block diagram of channel estimation procedure is shown in Figure 5.5.2. The primary importance of channel estimation is that it allows the receiver to take into ac-count the effect of channel on the transmitted signal, secondly channel estimation is essential for removing ISI, noise rejection techniques etc. In wideband mobile communications systems, a dynamic estimation of the channel is essential before the demodulation of OFDM signals because the radio channel is time-varying and frequency selective .There are two main types of channel estimation methods, namely blind methods and training sequence methods. In blind methods, mathematical or statistical properties of transmitted data are used. This makes the method extremely computationally intensive and thus hard to implement on real time systems. In training sequence methods or non-blind methods, the transmitted data and training sequences known to the receiver are embedded into the frame and sent through the channel. Generally, the length of the training sequence is twice or thrice the order of the channel and it is computationally simple compared to blind methods. One of the popular methods is to make use of the training bits (pilot symbols) known to the receiver. The transmitter periodically, inserts the symbol from which the receiver derives its amplitude and phase reference. Although training sequence method is much less computationally intensive than the blind methods, the channel bandwidth is not put into effective use by the transmission of training sequences. Another channel estimation method is called semi-blind method. The semi-blind methods use information from both training sequence and statistical properties of the transmitted signal, which makes them more robust than the blind methods while they still require less training compared to the non-blind methods. It is preferable to estimate the channel before converting the received signal to time domain so as to reduce or eliminate the risk of compounded error. Therefore in this project, frequency domain channel estimator is designed and simulated. In OFDM system, data are modulated on frequency domain sub-channels and scaled by different sub-channel frequency response coefficients after passing through the multipath channel. For coherent detection, these sub-channel frequency responses must be estimated. This estimation is usually done using training symbols which are embedded in the symbol. In this thesis pilots are used for channel estimation. A possible way of performing channel estimation is illustrated in Figure 5.5.2



CHAPTER 6

CFO

CHAPTER 6

CFO

Carrier frequency offset (CFO) is one of many non-ideal conditions that may affect in baseband receiver design. In designing a baseband receiver, we should notice not only the degradation invoked by non-ideal channel and noise, we should also regard RF and analog parts as the main consideration. Those non-idealities include sampling clock offset, IQ imbalance, power amplifier, phase noise and carrier frequency offset nonlinearity.

Carrier frequency offset often occurs when the local oscillator signal for down-conversion in the receiver does not synchronize with the carrier signal contained in the received signal. This phenomenon can be attributed to two important factors: frequency mismatch in the transmitter and the receiver oscillators; and the Doppler effect as the transmitter or the receiver is moving.

When this occurs, the received signal will be shifted in frequency. For an OFDM system, the orthogonality among sub carriers is maintained only if the receiver uses a local oscillation signal that is synchronous with the carrier signal contained in the received signal. Otherwise, mismatch in carrier frequency can result in inter-carrier interference (ICI). The oscillators in the transmitter and the receiver can never be oscillating at identical frequency. Hence, carrier frequency offset always exists even if there is no Doppler effect.

TYPES:

AIDED: Channel estimation is done on known data both at transmitter and receiver.

BLIND/NON AIDED: Estimation is only on received data, without any known transmitted sequence.

Here in this project we are going for blind/non aided type of estimation and we are going to reduce the complexity.

CHAPTER 7

METHODOLOGY

CHAPTER 7

METHODOLOGY

Consider an OFDM system. Let N and L be respectively the size of the discrete Fourier transform (DFT) and the length of the cyclic prefix (CP). Let xk = [xk(0) xk(1) ··· xk(N − 1)] T be the kth time domain vector at the output of the inverse DFT. The transmitter adds L CP samples xk,cp = [xk(N − L) ··· xk(N − 2) xk(N − 1)] T to form the (N + L) × 1 cyclic prefixed vector x k = [xT k,cp xT k] T . Assume that the channel order does not exceed L. Then at the receiver, the kth received vector of size N + L is given by

$$\mathbf{r}'_k = e^{j2\pi\frac{k(N+L)}{N}\theta}\mathbf{D}_{N+L}(\theta)\mathbf{H}\begin{bmatrix}\mathbf{x}_{k-1,\mathrm{cp}}\\ \mathbf{x}'_k\end{bmatrix} + \mathbf{n}_k,$$

where θ is the normalized CFO and H is a (N +L)×(N +2L) Toeplitz matrix expressed as

$$\mathbf{H} = \begin{bmatrix} h(L) & \cdots & h(0) & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & h(L) & \cdots & h(0) \end{bmatrix}$$

with [h(0) ··· h(L)] being the channel impulse response. The last term nk in (1) is the kth noise block which is assumed to be an additive white Gaussian noise (AWGN) vector with variance σ2 n.

In this project, we will show how to blindly estimate the normalized CFO θ from the received signal. It does not need to know the channel information and works for both constant modulus and non constant modulus modulation symbols.

A remodulation of the received signal for blind channel estimation (in the absence of CFO). In this paper, we will show how the remodulated signal can be used for blind CFO estimation when there is CFO. Let us consider the remodulated vector:

$$\tilde{\mathbf{r}}_k \triangleq \begin{bmatrix} \mathbf{r}'_{k-1}(L : N + L - 1) \\ \mathbf{r}'_k(0 : L - 1) \end{bmatrix}.$$

That is, ˜rk is a (N + L) × 1 vector formed by the last N entries of r k−1 and the first L entries of r k. The vector ˜rk can be represented by

$$\tilde{\mathbf{r}}_k = e^{j2\pi \frac{k(N+L)-N}{N}\theta} \mathbf{D}_{N+L}(\theta)\mathbf{H} \begin{bmatrix} \mathbf{x}'_{k-1} \\ \mathbf{x}_{k,\mathrm{cp}} \end{bmatrix} + \tilde{\mathbf{n}}_k,$$

where n˜k = [nk−1(L : N + L − 1)T nk(0 : L − 1)T] T . Now let us consider the following vector

$$\mathbf{y}_k(\xi) = \mathbf{r}'_k - e^{j2\pi\xi}\tilde{\mathbf{r}}_k.$$

Utilizing (1) and (4), we can express yk(ξ) as

$$\begin{aligned} \mathbf{y}_k(\xi) =& e^{j2\pi \frac{k(N+L)}{N}\theta} \mathbf{D}_{N+L}(\theta)\mathbf{H} \left(\begin{bmatrix} \mathbf{x}_{k-1,\mathrm{cp}} \\ \mathbf{x}'_k \end{bmatrix} \right. \\ & \left. - e^{j2\pi(\xi-\theta)} \begin{bmatrix} \mathbf{x}'_{k-1} \\ \mathbf{x}_{k,\mathrm{cp}} \end{bmatrix} \right) + \underbrace{\left(\mathbf{n}_k - e^{j2\pi\xi}\tilde{\mathbf{n}}_k\right)}_{\boldsymbol{\eta}_k}. \end{aligned}$$

Computing the autocorrelation matrix of yk(ξ), we have

$$\begin{aligned} \mathbf{R}_y(\xi) =& \mathrm{E}\{\mathbf{y}_k(\xi)\mathbf{y}_k^{\dagger}(\xi)\} \\ =& \mathrm{E}\{(\mathbf{r}'_k - e^{j2\pi\xi}\tilde{\mathbf{r}}_k)(\mathbf{r}'_k - e^{j2\pi\xi}\tilde{\mathbf{r}}_k)^{\dagger}\}. \end{aligned}$$

Assume that the noise and signal are uncorrelated and the N ×1 transmitted vector xk satisfies E{xkx† k−i} = σ2 xIN δ(i), where σ2 x is the average power of the transmitted data and δ(i) is the Kronecker delta function. Then it can be shown that the autocorrelation matrix is expressed as is a set having a cardinality of 2L. The CFO can be estimated by

$$\hat{\theta} = \arg \min_{\xi \in (-0.5, 0.5]} \bar{J}(\xi).$$

The minimum can be found by taking the derivative of J¯(ξ) with respective to ξ and setting it to zero. Utilizing it can be shown that for ξ ∈ (−0.5, 0.5], the minimum is unique and an estimate of the CFO is given by

$$\hat{\theta}_{\text{coarse}} = \frac{1}{2\pi} \angle \left(\sum_{k=1}^{K-1} \sum_{i \in \mathcal{S}_1} (\tilde{r}_k(i))^* r'_k(i) \right),$$

where the phase ∠(·) of a complex number is defined in the region (−π, π]. The subscript "coarse" indicates that this gives a coarse estimate of the CFO. In the simulation, we find that the mean square error (MSE) for ˆθcoarse floors at around 10−4 for high signal-to-noise ratio (SNR). To solve this issue, we propose the following refinement on the CFO estimate.

Fine estimation Let us consider the two cost functions J1(ξ − θ) and J2(ξ − θ) . It is seen that the second order derivative (∂2J1)/(∂ξ2)|ξ=θ is larger than (∂2J2)/(∂ξ2)|ξ=θ. For the cost function J1(ξ − θ), a small estimation error ξ − θ = Δθ leads to a large increase in J1(ξ − θ). On the other hand, a small estimation error leads to a small increase in J2(ξ − θ). Therefore, when the cost functions contain a noise term, the estimate based on J1(ξ − θ) will be more accurate than the estimate based on J2(ξ − θ). For this reason, let us calculate the second order derivatives of the main diagonal of Ry(ξ). From (11), we can get

$$\left[\frac{\partial^2 \mathbf{R}_y(\xi)}{\partial \xi^2}\right]_{i,i} = \begin{cases} 8\pi^2\epsilon_1\sigma_x^2 \sum_{l=i+1}^{L} |h(l)|^2 & \text{if } 0 \le i < L, \\ 8\pi^2\epsilon_1\sigma_x^2 \sum_{l=0}^{i-N} |h(l)|^2 & \text{if } N \le i < N+L, \\ 0 & \text{otherwise,} \end{cases}$$

If |ξ − θ| < 1/4, then 1 > 0. we find that those smaller diagonal entries [Ry(ξ)]i,i have larger second order derivatives. Therefore, instead of including all 2L terms of S1 in the cost function J¯(ξ) as we include only the m (where m < 2L) smallest terms. In order to find these m smallest terms, one can use the coarse estimate ˆθcoarse to evaluate [Ry(ˆθcoarse)]i,i.

In summary, our algorithm is as follows.

1) Compute a coarse estimate ˆθcoarse using .

2) Calculate [Ry(ˆθcoarse)]i,for i ∈ S1.

3) Let the new set S2 contain those indices of the m smallest diagonal values.

4) Obtain the fine estimate as

$$\hat{\theta}_{\text{fine}} = \frac{1}{2\pi}\angle\left(\sum_{k=1}^{K-1}\sum_{i\in\mathcal{S}_2}(\tilde{r}_k(i))^* r_k'(i)\right).$$

The investigated communication system employs an OFDM physical layer with a DCR architecture. Each OFDM block has length T and is preceded by a cyclic prefix (CP) to avoid the interblock interference (IBI). We denote by N the number of available subcarriers and let 1/T be the subcarrier spacing in the frequency domain. To facilitate the frequency estimation task, a training preamble is appended in front of each data frame. In contrast to conventional pilot sequences composed by several repetitive segments -, our preamble is specifically designed to cope with I/Q imbalances and consists of two adjacent sections s0 and s1

TABLE I: Comparison of the number of CMUL

Blind CFO Estimation	CMUL in terms of N, L, and K	CMUL with the setting in Section IV
Proposed	$6LK - 4L$	896
FA [3]	$K(2N\log_2 N + 3N - 6L - 3)$	8610
Kurtosis [4]	$K(N\log_2 N + 6N)$	7680
MUSIC [6]	$QK((N\log_2 N)/3 + N - P)$	140000

NOTATION

Matrices and vectors are denoted by boldface letters, with IN and 1N being the identity matrix of order N and the N-dimensional vector with unit entries, respectively. A = diag{a(n) ; n = 1, 2,...,N} denotes an N × N diagonal matrix with entries a(n) along its main diagonal and B−1 is the inverse of a matrix B. We use E{·}, (·)∗, (·)T and (·)H for expectation, complex conjugation, transposition and Hermitian transposition, respectively. The notation · represents the Euclidean norm of the enclosed vector while e{x}, m{x}, |x| and arg{x} stand for the real and imaginary parts, the modulus and the principal argument of a complex number x. The symbol ⊗ is adopted for the continuous-time convolution and [C]k, is the (k,)th entry of C. Finally, λ˜ denotes a trial value of the unknown parameter λ.

The comparison of complexity with some existing methods is given in Table I, where Q represents the number of times that the one-dimensional search repeats in MUSIC method . In practice, the CP length L is usually much smaller than the DFT size N. Therefore, the proposed method has a much lower complexity than other methods. In the table, we also explicitly calculate the number of CMUL for the setting given in the simulation. It can be seen that the proposed method has very low complexity.

We assume that the channel does not change while CFO estimation is performed. Channel taps are generated as independent random variables with exponentially decaying powers of 2−l . The channel noise is AWGN. Two cases of the transmission symbols are considered (i) 16-QAM; (ii) QPSK.

The OFDM block size is N = 64. The length of CP is L = 16, and the set S2 has a cardinality of m = L/2=8. The number of blocks is K = 10. The MSE is defined as

$$\mathrm{MSE} = \frac{1}{R}\sum_{r=1}^{R} |\hat{\theta}^{(r)} - \theta|^2,$$

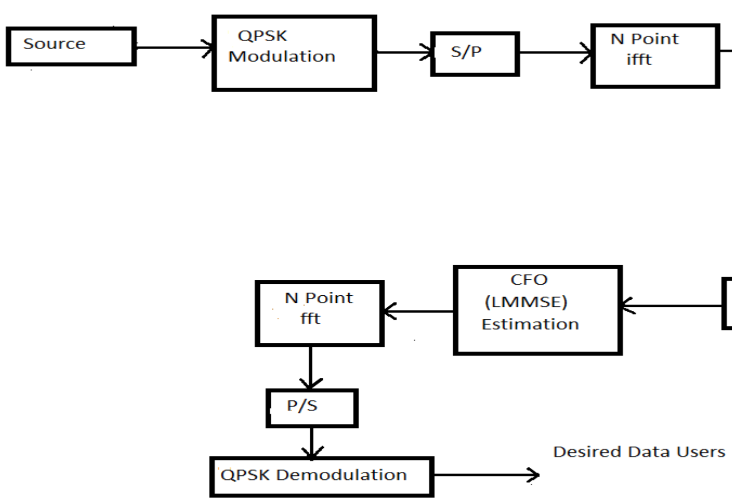
where ˆθ(r) represents the estimated CFO in the rth trial. R = 2000 denotes the total number of Monte Carlo trials. First assume that there is no virtual carriers. It can be seen that the MUSIC-like algorithm does not work since there is no VC. The FA algorithm does not perform well for 16-QAM because it is based on constant modulus modulation. For both cases of 16-QAM and QPSK, the coarse estimation has an error floor at about 10−4 at high SNR. The performance of Kurtosis method degrades for the case of 16-QAM. The proposed fine estimation method has the best performance, and its performance does not dependent on the modulation symbols.

Next we consider OFDM systems with virtual carriers.1 The numbers of data subcarriers and VCs are P = 52 and N − P = 12 respectively. The indices for VCs are {0,..., 5, 32, 59,..., 63}.

The performance of MUSIC method depends on the search resolution, and with a search resolution of 1/Q (Q = 100), its MSE floors at 2.5 × 10−5. On the other hand, the presence of VCs has little effect on the proposed method. The proposed method has the best performance in the case of VCs as well.

IMPACT OF THE NOISE

When developing the proposed method, the noise is not considered for clarity. To illustrate the impact of the noise on the FA method, the curves of |yk u(i)| 2 varies with k when N = 64, P = 8, L = 4, = 0.1, and SNR= 3, 5, 10dB. Thus the noise does not obviously affect the performance of the FA method in the higher SNR in cases we compare the proposed FA method with the CP , the MOV , the YG and the SBS methods. An OFDM system with QPSK constellation is considered with N = 64, P = 8. A five-tap Rayleigh block fading channel is used with the exponentially decaying powers set as E(|hl(i)| 2) = e−l/3/ 4 m=0 e−m/3, l = 0, 1, ··· , 4. The CFO is assumed to be uniformly distributed between (−0.5, 0.5]. All results are averaged by 5000 Monte Carlo trials. In addition, the MSEs of these methods are also evaluated in a fast time-varying channel with the classical Jakes model . The maximum Doppler frequency is 0.1 with respect to the subcarrier spacing. In the above simulations, the actual channel order is assumed to be known at the receiver. With the assumed channel order varying from 0 to 25, the performance of the FA method is evaluated and depicted . It can be seen that our FA method still performs well if the assumed channel order is (slightly) larger than the actual channel order. This implies that if the channel order is unknown, using the length of CP is a simple and reasonable choice in practice. The below figure 7.1 is the block diagram of our project



CHAPTER 8

HARDWARE AND SOFTWARE REQUIREMENTS

CHAPTER 8

8.1 HARDWARE REQUIREMENTS

- System : Pentium Dual Core.
- Hard Disk : 120 GB.
- Monitor : 15’’ LED
- Input Devices : Keyboard, Mouse
- Ram : 1 GB

8.2 SOFTWARE REQUIREMENTS

- Operating system : Windows 7.
- Software Tool : Matlab R2016

8.2.1 MATLAB

MATLAB is a high-performance language for technical computing. It integrates computation, visualization, and programming in an easy-to-use environment where problems and solutions are expressed in familiar mathematical notation. Typical uses include:

- Math and computation
- Algorithm development
- Modeling, simulation, and prototyping
- Data analysis, exploration, and visualization
- Scientific and engineering graphics

- Application development, including Graphical User Interface building

MATLAB is an interactive system whose basic data element is an array that does not require dimensioning. This allows you to solve many technical computing problems, especially those with matrix and vector formulations, in a fraction of the time it would take to write a program in a scalar noninteractive language such as C or Fortran.The name MATLAB stands for matrix laboratory. MATLAB was originally written to provide easy access to matrix software developed by the LINPACK and EISPACK projects, which together represent the state-of-the-art in software for matrix computation.MATLAB has evolved over a period of years with input from many users. In university environments, it is the standard instructional tool for introductory and advanced courses in mathematics, engineering, and science. In industry, MATLAB is the tool of choice for high-productivity research, development, and analysis.

CHAPTER 9

CODE

CHAPTER 9

CODE

```
clc
clear all;
close all;
%% Parameter declaration according to project
Nfft = 1024;
Ncp = 128;
Nsym = 6;
FreqOffset = 0.25;
SNRdb = 15;
theta = 256;
%% OFDM symbol generation
data = 2*randi([0 1],1, Nsym*Nfft)-1; % QPSK data
 Tx = zeros(1,Nsym*(Nfft+Ncp));
OFDMsym = zeros(1,Nfft);
for sym = 1:Nsym
  OFDMsym = ifft(data(Nfft*(sym-1)+1:(Nfft*sym)),Nfft)*sqrt(Nfft);
  Tx((Nfft+Ncp)*(sym-1)+1:(Nfft+Ncp)*sym) = [OFDMsym(Nfft-Ncp+1:Nfft) OFDMsym];
end
 %% AWGN channel
snr = 10^(-SNRdb/10);
noise = sqrt(snr/2)*(randn(1,Nsym*(Nfft+Ncp))+1i*randn(1,Nsym*(Nfft+Ncp)));
```

```
Rx = exp(1i*2*pi*FreqOffset*(0:length(Tx)-1)./Nfft).*Tx + noise;
 %% CFO estimation of timing and frequency offset
PHI_sum = zeros(1,Nsym*(Nfft+Ncp)-Nfft);
GM_sum = zeros(1,Nsym*(Nfft+Ncp)-Nfft);
 for n = theta:Nsym*(Nfft+Ncp)-(Nfft+Ncp)
  PHI=0;GM=0;
  for m = n:n+Ncp-1
  PHI = PHI+ (Rx(m)*conj(Rx(m)) + Rx(m+Nfft)*conj(Rx(m+Nfft)));
  GM = GM+ Rx(m)*conj(Rx(m+Nfft));
  end
  PHI_sum(n) = abs(GM)- (snr/(snr+1))*PHI;
  GM_sum(n) = -angle(GM)/(2*pi);
end
 %% Estimation results display
 clear all;
 nCP = 8;%round(Tcp/Ts);
nFFT = 64;
NT = nFFT + nCP;
F = dftmtx(nFFT)/sqrt(nFFT);
 MC = 1500;
 EsNodB = 0:5:40;
snr = 10.^(EsNodB/10);
beta = 17/9;
```

```
M = 16;
modObj = modem.qammod(M);
demodObj = modem.qamdemod(M);
L = 5;
ChEstLS = zeros(1,length(EsNodB));
ChEstMMSE = zeros(1,length(EsNodB));
TD_ChEstMMSE = zeros(1,length(EsNodB));
TDD_ChEstMMSE = zeros(1,length(EsNodB));
TDQabs_ChEstMMSE = zeros(1,length(EsNodB));
 for ii = 1:length(EsNodB)
  disp('EsN0dB is :'); disp(EsNodB(ii));tic;
  ChMSE_LS = 0;
  ChMSE_LMMSE=0;
  TDMSE_LMMSE =0;
  TDDMSE_LMMSE=0;
  TDQabsMSE_LMMSE =0;
  for mc = 1:MC
% Random channel taps
g = randn(L,1)+1i*randn(L,1);
  g = g/norm(g);
  H = fft(g,nFFT);
% generation of symbol
X = randi([0 M-1],nFFT,1); %BPSK symbols
```

```
  XD = modulate(modObj,X)/sqrt(10); % normalizing symbol power

  x = F'*XD;

  xout = [x(nFFT-nCP+1:nFFT);x];

% channel convolution and AWGN

y = conv(xout,g);

  nt =randn(nFFT+nCP+L-1,1) + 1i*randn(nFFT+nCP+L-1,1);

  No = 10^(-EsNodB(ii)/10);

  y = y + sqrt(No/2)*nt;

% Receiver processing

 y = y(nCP+1:NT);

  Y = F*y;

% frequency doimain LS channel estimation

HhatLS = Y./XD;

  ChMSE_LS = ChMSE_LS + ((H -HhatLS)'*(H-HhatLS))/nFFT;

% Frequency domain LMMSE estimation
Rhh = H*H';

  W = Rhh/(Rhh+(beta/snr(ii))*eye(nFFT));

  HhatLMMSE = W*HhatLS;

  ChMSE_LMMSE = ChMSE_LMMSE + ((H -HhatLMMSE)'*(H-HhatLMMSE))/nFFT;

  % Time domain LMMSE estimation

 ghatLS = ifft(HhatLS,nFFT);

  Rgg = g*g';

  WW = Rgg/(Rgg+(beta/snr(ii))*eye(L));

  ghat = WW*ghatLS(1:L);
```

```
  TD_HhatLMMSE = fft(ghat,nFFT);%
  TDMSE_LMMSE=TDMSE_LMMSE+((H-TD_HhatLMMSE)'*(H-TD_HhatLMMSE))/nFFT;
  % Time domain LMMSE estimation - ignoring channel covariance
ghatLS = ifft(HhatLS,nFFT);
  Rgg = diag(g.*conj(g));
  WW = Rgg/(Rgg+(beta/snr(ii))*eye(L));
  ghat = WW*ghatLS(1:L);
  TDD_HhatLMMSE = fft(ghat,nFFT);%

TDDMSE_LMMSE=TDDMSE_LMMSE+((HTDD_HhatLMMSE)'*(HTDD_HhatLMMSE))/n
FFT;
 % Time domain LMMSE estimation - ignoring smoothing matrix
ghatLS = ifft(HhatLS,nFFT);
TDQabs_HhatLMMSE = fft(ghat,nFFT);%
TDQabsMSE_LMMSE=TDQabsMSE_LMMSE+((HTDQabs_HhatLMMSE)'*(HTDQabs_Hhat
LMMSE))/nFFT;
end
ChEstLS(ii) = ChMSE_LS/MC;
  ChEstMMSE(ii)=ChMSE_LMMSE/MC;
  TD_ChEstMMSE(ii)=TDMSE_LMMSE/MC;
  TDD_ChEstMMSE(ii)=TDMSE_LMMSE/MC;
  TDQabs_ChEstMMSE(ii)=TDQabsMSE_LMMSE/MC;
  toc;
end
```

CHAPTER 10

RESULTS

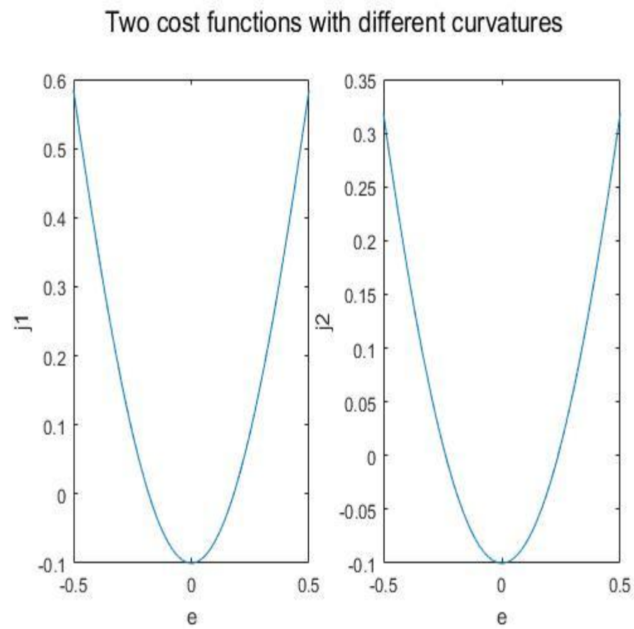


Figure 10.1

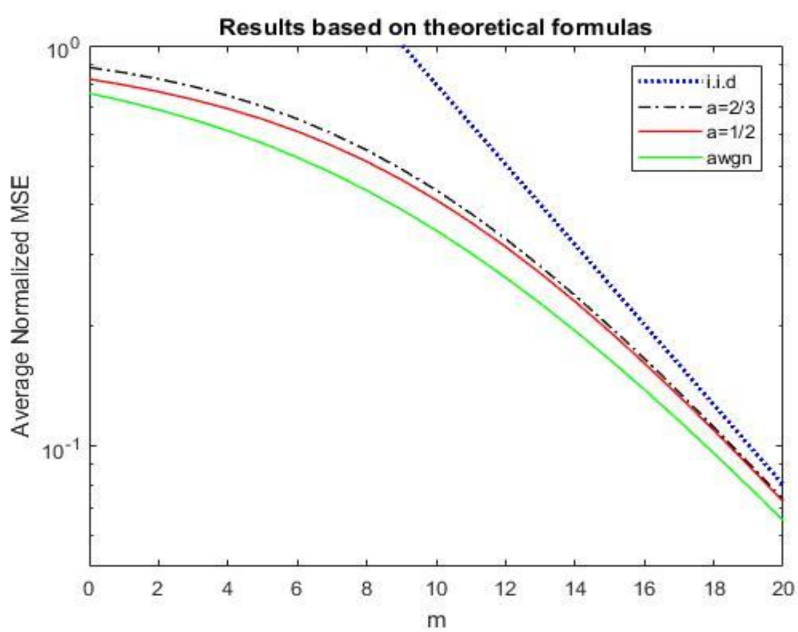


Figure 10.2

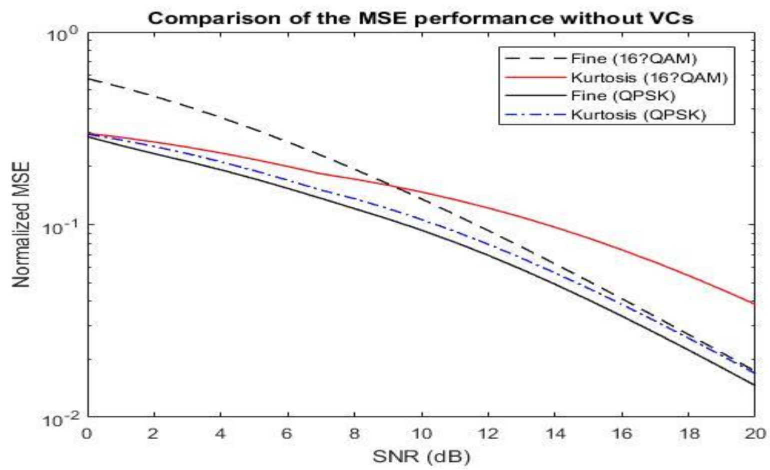


Figure 10.3

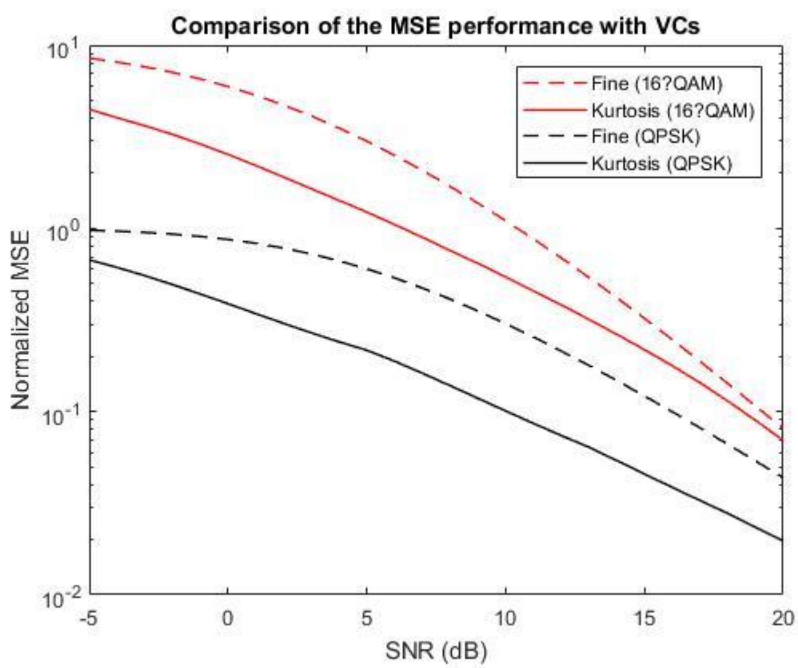


Figure 10.4

CHAPTER 11

ADVANTAGES AND APPLICATIONS

CHAPTER 11

ADVANTAGES AND APPLICATIONS

ADVANTAGES

OFDM has been used in many high data rate wireless systems because of the many advantages it provides.

- **Immunity to selective fading:** One of the main advantages of OFDM is that is more resistant to frequency selective fading than single carrier systems because it divides the overall channel into multiple narrowband signals that are affected individually as flat fading sub-channels.

- **Resilience to interference:** Interference appearing on a channel may be bandwidth limited and in this way will not affect all the sub-channels. This means that not all the data is lost.

- **Spectrum efficiency:** Using close-spaced overlapping sub-carriers, a significant OFDM advantage is that it makes efficient use of the available spectrum.

- **Resilient to ISI:** Another advantage of OFDM is that it is very resilient to inter-symbol and inter-frame interference. This results from the low data rate on each of the sub-channels.

- **Resilient to narrow-band effects:** Using adequate channel coding and interleaving it is possible to recover symbols lost due to the frequency selectivity of the channel and narrow band interference. Not all the data is lost.

- **Simpler channel equalisation:** One of the issues with CDMA systems was the complexity of the channel equalisation which had to be applied across the whole channel.

An advantage of OFDM is that using multiple sub-channels, the channel equalization becomes much simpler.

DISADVANTAGES

- Sensitive to carrier frequency offsets.
- More complex than single carrier modulation.
- Very sensitive to frequency erros.

APPLICATIONS

- 4G wireless and high capacity LAN'S
- Digital Audio Broadcasting(DAB)
- Digital Video Broadcasting(DVB)
- Wireless ATM transmission system

CHAPTER 12

CONCLUSION

CHAPTER 12

CONCLUSION

Aiming at OFDM systems, we have addressed the principles, system model and state of the art of the estimation and compensation of blind CFO's with low complexity. The results show that there has been adequate intensive study on the CFO estimation with and without virtual carriers and still need further research.

FUTURE SCOPE

The channel estimation is an area which required a lot of attention and improper channel estimation degrades the performance of system, it is assumed that channel is estimated perfectly. Hence one can evaluate the performance of work with different channel estimation method. The algorithm of frequency offset estimation can be extended for channel estimation in OFDM system.

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