Digital Methods for Reduction of IQ Imbalance in OFDM Systems

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Abstract

Orthogonal frequency division multiplexing (OFDM) based systems are susceptible to radio front end induced impairments. These impairments caused by as a consequence of imbalances in the In-phase (I) and Quadrature-phase (Q) branches. IQ imbalances are worries to be addressed for the Superhetrodyne and the zero IF architectures in analog domain.

The zero-IF architecture is a prospective solution to include multiple RF standards while making straightforward the design and dropping overall costs. However in the practical implementation of the zero-IF receiver architecture IQ imbalance has been recognized as one of the most serious concerns.

At the receiver the expected working signal to noise ratio (SNR) can be affected heavily by the presence of IQ imbalance. As a result of this imbalance the data rates and the supported constellation are also affected.

In this thesis the effect of IQ imbalance on OFDM receivers is considered and digital methods are proposed to tackle the effect of IQ imbalance and enhance the operating signal to noise ratio (SNR). The methods are the least square (LS) and the least mean algorithm (LMS). With the help of computer simulation we are going to study the performance of both reduction techniques. The bit error rate (BER) Verses signal to noise ratio (SNR) result could indicate the proposed techniques provide sufficient improved performance.

Keywords: IQ imbalance, least square (LS), least mean algorithm (LMS), OFDM

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I am thankful to my supervisor Kristian Nilsson, whose guidance from the initial to the end. And I offer my blessings to my family and my friends who supported me in any respect.

Thanks to God for making everything happen.

Daxious Mengistu
Karlskrona
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CHAPTER ONE

INTRODUCTION

1.1 Background

In the past radio frequency design has been complex and requires a lot of time to process. The growth in design of modern radio frequency integrated circuits has become much more easily automated with software tools. A trend in broadband wireless communication systems provides higher data rates for applications that require information capacity. Orthogonal frequency division multiplexing (OFDM) is a form of multi carrier modulations that provides high data rate in broadband wireless communications. The advancement of digital signal processing alleviates the barriers of OFDM construction, such as high speed memory access and huge composite multiplications are avoided.

IEEE 802.11a wireless local area network (WLAN) in the 5GHz band [14], IEEE 802.11g wireless local area network (WLAN) in the 2.4GHz band [15], and the European digital video broadcasting system (DVB-T) [34] are OFDM based. Some recently adopted IEEE P802.15.3 wireless personal area network (WPAN) [15], the IEEE 802.20 mobile broadband wireless access (MBWA) [27] and the IEEE 802.16 wireless metropolitan area networks (WMAN) [16] considered OFDM.

OFDM supports different kind of configuration to change the RF into its baseband counterpart. The Superhetrodyne configuration is a well established design that assures better reception by filtering and amplifying at each step. It changes the received signal into one or intermediate frequencies to get the required baseband signal. The drawback of this architecture is it needs a lot of apparatus to achieve fine quality resulting in need of very large surface area [33].

The zero intermediate frequency (zero-IF) provides an alternative to the superhetrodyne receiver. Compared to the superhetrodyne receivers the zero-IF provides numerous advantages most notably the single conversion without the need of intermediate frequency. Avoiding the intermediate frequency reduces the need for more components and lessens the cost of the receiver. This attractive feature revives the need of the zero-IF receiver in terms of cost, package size and power consumption [23].

OFDM based systems are susceptible to radio front end induced impairments. These impairments caused by as a consequence of imbalances in the In-phase (I) and Quadrature-phase (Q) branches. IQ imbalances are concerns to be addressed for the Superhetrodyne and the zero IF architectures in analog domain. Sensitiveness of analog circuit to component variation resulted in unavoidable errors in IQ branches due to temperature variation and process mismatches. Hence it’s a challenging matter for silicon implementations to achieve orthogonal sinusoidal waveforms at radio frequencies as high as 5.2GHz (IEEE 802.11a). To reduce these mismatches some techniques are developed in analog domain. Voltage controlled oscillator (VCO) and tunable Polyphase are used in analog designs considered to be the stoutest for component mismatches [3]. However these techniques suffer from errors in measurement, different offsets and long time calibration process [3].

1.2 Motivation

Analog domain techniques can not meet the specifications required for systems such as IEEE 802.11a without digital compensation [9]. The tradeoff caused by analog domain such as area for precision, power and speed does not exist in digital domain with the same potency. These tradeoffs make it difficult to design analog circuits efficiently in terms of power and area. As a result digital processing power takes the advantage to improve the analog domain imperfection.

To keep ortogonality between subcarriers ideal circumstances must be fulfilled. Such as the channel is time invariant over OFDM block period, no IQ imbalance and no carrier frequency offset. These conditions can not be contented in practice. In general, the motivation for digital compensation approach is as follows:

- The high complexity to attain correct IQ match in analog domain
- When high data rates are considered, higher operating SNR and higher constellation sizes are needed.
- Higher IQ imbalances resulted when the carrier frequency increased, and it’s a tough work to eliminate the imbalance [3].

Considering the above reasons it is desirable to develop digital compensation algorithms to eliminate the effect of IQ imbalance in wireless receivers. Some previous works that have been proposed to tackle the front end problems over OFDM transmission, in [1] the author assumed perfect IQ balance and proposes a training based technique for carrier frequency offset (CFO). In [2] Maximum likelihood frequency offset estimation is proposed.

1.3 Thesis objectives

The zero-IF architecture is a prospective solution to cover multiple RF standards while simplifying the design and reducing overall costs. However in the practical implementation of the zero-IF receiver architecture IQ imbalance has been recognized as one of the most serious concerns.

An IQ imbalance model is developed based on the zero-IF receivers. The developed model is used to present digital compensation algorithms. Two digital techniques are going to be analyzed. These are Least Square (LS) method and Least Mean Square (LMS) method. With the help of computer simulation performance of both compensation techniques will be examined. The bit error rate (BER) Verses signal to noise ratio (SNR) result could indicate the proposed techniques provide sufficient compensation performance.

To be more specific on the objectives:

- To develop IQ imbalance model and to investigate the OFDM systems with IQ imbalance
- Recommending Least Mean Square (LMS) and Least Square (LS) receiver methods that make up for IQ imbalance [7].
- Evaluating the proposed techniques in terms of bit error rate (BER) as a function to signal to noise ratio (SNR).

1.4 Thesis organization

In chapter two the basic principles of OFDM systems are discussed based on a given model. In addition to this the problems that make the implementation difficult are discussed and analyzed.

In chapter three the OFDM receiver with IQ imbalance impairments is considered and analysis on the basis of IQ imbalances will be given. A Model is configured based on the mathematical investigation of an IQ imbalance.

In chapter four explains the algorithms to minimize the consequence of IQ imbalance impairment in digital domain. The algorithms are Least Square (LS) and least mean square (LMS). Training method is used in both algorithms to approximate the distortion parameter that model the IQ imbalances. For some systems that do not provide training symbols other schemes like decision direction is used.

In chapter five simulation results are presented with some explanations and finally in chapter six conclusions and some futures work are mentioned.

CHAPTER TWO

INTRODUCTION TO OFDM SYSTEMS AND ITS COMPLICATION

In this chapter, the basic principles of OFDM systems are discussed based on a basic system model of OFDM modulation. In addition to this the problems that make the implementation difficult are discussed and analyzed.

2.1 Basic principles of OFDM systems

2.1.1 Introduction

OFDM is a multi channel modulating technique that makes use of Frequency Division Multiplexing (FDM) of orthogonal multi carriers being modulated by a low bit rate digital stream. In FDM, inter channel interference is diminished by the deterrence of the spectral overlapping of sub-carriers but it guides to an inadequate use of spectrum. To prevail over this obstacle OFDM uses orthogonal sub-carrier that helps an efficient use of the spectrum. This can be achieved by spacing the channels much closer to each other as shown in figure 2.1. [33]

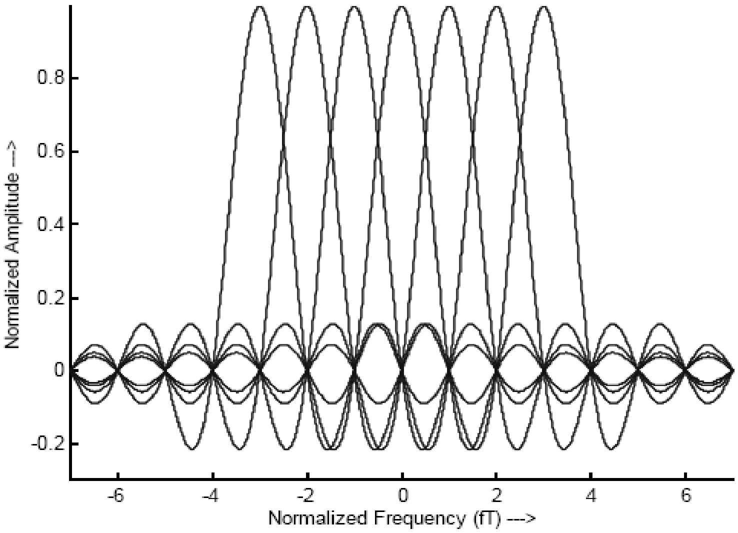

Figure2.1: Spectra of Individual Sub-Carriers

The orthogonality state shows that each sub carrier contains exactly integer number of cycles in the interval period.The Inverse Fast Fourier Transform (IFFT) and Discrete Fourier Transform (DFT) can be used to modulate and demodulate the sub-carriers hence the generation of OFDM signal.

OFDM acquires some inherent merits for wireless communication. Some of the good reasons that OFDM become more popular in the wireless industry today are:

- Emitter and receiver are efficiently implemented with FFT/IFFT
- Throughput maximization
- Effectiveness against channel distortion
- Multi-path delay spread tolerance

Even though OFDM has numerous advantages it has some disadvantages on the quality of the analogue radio frequency front end of both transmitter and receiver [13]:

- Sensitevness to carrrier frequency errors
- To maintain the orthogonality between subcarriers, the amplifiers need to be linear
- OFDM systems have high peak-to average ratio which may require large amplifier.

2.1.2 Generation of OFDM signals

To put in practice the OFDM transmission method, the message signal has to be digitally modulated. To maintain orthogonality between the sub carriers the carrier is split into lower frequency sub carriers. This is achieved by a series of digital signal processing operation [33], see figure 2.2.

2.1.3 Transmitter

Modulating schemes such as Quadrature amplitude modulation (QAM) (16QAM or 64QAM for example) or some form of phase shift keying (PSK) are used to modulate the message signal. Before passing to the IFFT block the modulated signal has to be converted to parallel signals. The IFFT operation will take place to form a single combined signal that can be transmitted. These combined signals are orthogonal subcarriers. Let the output of the IFFT to be $x(n)$ where,

$$x(n) = \mathrm{IDFT}\{X(k)\}$$

$$= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi nk / N}, \quad 0 \le n \le N-1 \qquad (2.1)$$

where $X(k)$ is the modulated signal for $k = 0,1, \dots . , N$ and N is the number of sub carriers.

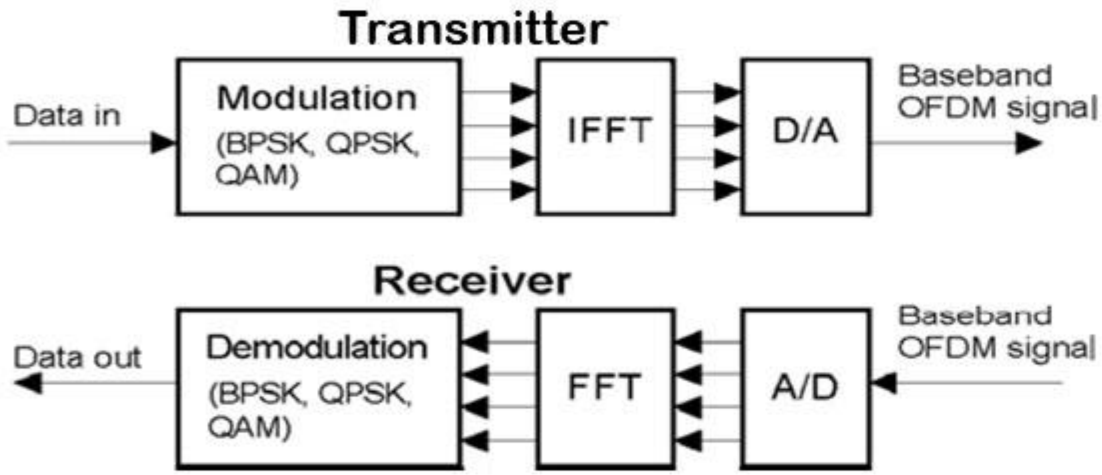

Figure 2.2 Block Diagram for OFDM modulation

To achieve a set of orthogonal signals from the sub carriers, the data is united in frames of proper size arrangement for IFFT or FFT. Next, an N point IFFT is performed as shown in figure 2.2 and the data stream is the output of the transmitter. The IFFT process helps us to evaluate the receiver signal at a frequency chosen by spacing of the subcarriers, all other signals are zero.

To attain this orthogonality the receiver and the transmitter have to be synchronized. This means both must assume exactly the same modulation scheme and the same time-scale for transmission [17].

2.1.4 Guard period

When OFDM signals are transmitted multipath delay could affect the transmission. The inter symbol interference (ISI) caused by multipath propagation can be avoided by inserting guard interval between every OFDM symbol. The insertion of the guard period between transmitted symbols is vital if the signals in subcarriers are to hold on to the orthogonality during transmission process. The guard period allows multipath components from previous symbol to die out before information from the next OFDM symbol is recorded.

Cyclic prefix is the most effective guard period attached in front of every OFDM symbol. The cyclic prefix is the copy of the last part of the OFDM symbol added in front of the transmitted symbol, provided that the length is of equal or greater than the maximum delay spread of the channel (see figure 2.3). In figure 2.4 additions of cyclic prefix and what the transmitted signal look like is presented.

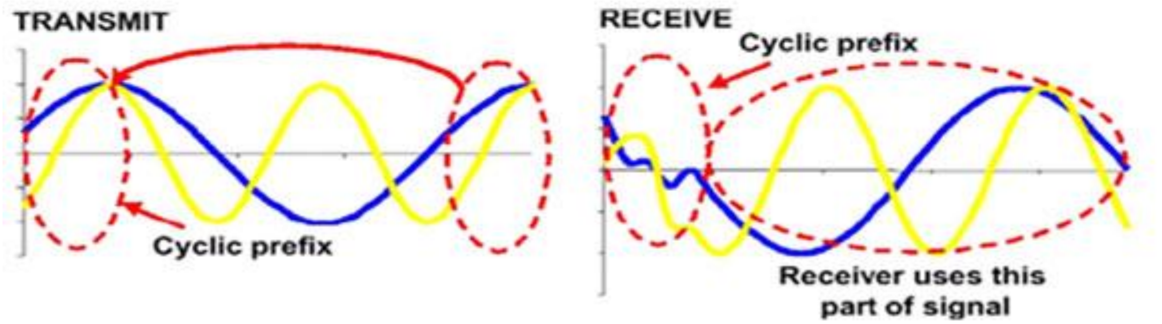

Figure 2.3 Implementation of cyclic prefix [32]

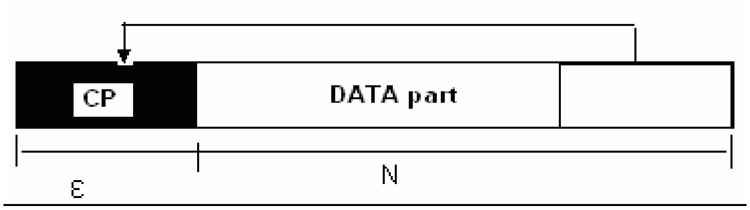

Figure 2.4 addition of cyclic prefix

2.1.5 Channel model

Transmitted signals are affected by multipath propagation. These signals are reflected, refracted or scattered by objects that come across during transmission before it reaches the intended receiver. Same signals appear at the receiver with different phase, amplitude and with different time delays. These signals are added destructively at the receiver and make the received signal to fade.Multipath fading environment and block fading channel model (the fading gain process is piecewise constant on blocks of N symbols and changes to other independent value) is assumed. Figure 2.5 shows the channel is represented by FIR system for one data packet [33].

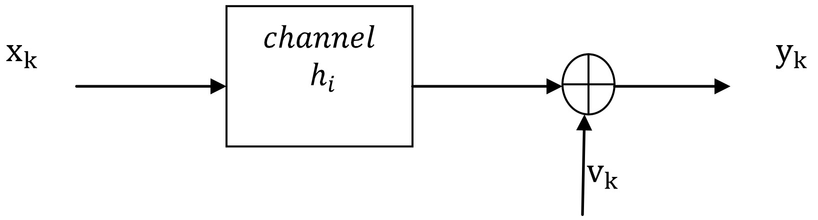

Figure 2.5 channel representation.

Therefore,

$$y_k = \sum_{i=0}^{\varepsilon} h_i \, x_{k-i} + v_k, \quad \text{for } 0 \le k \le N-1 \qquad (2.2)$$

Where: $y_k$is the received signal, $h = [h_1, h_2, \ldots, h_\varepsilon]^T$ is the channel matrix, x_k is the input symbol and v_k is the additive noise.

2.1.6 Receiver

At the receiver end the exact reverses process take place to recover the data with the help of FFT in which it converts the signal into the frequency domain it then demodulated according to the block diagram as shown in figure 2.2. The fundamental principle is that FFT can keep tones orthogonal to one another if the tones have an integer number of cycles in a symbol period [33].

The received signal can be expressed as:

$$y(t) = x'(t) * h(t) + v(t) \qquad (2.3)$$

where $x'(t)$ is the analog signal, $h(t)$ is the channel impulse response and $v(t)$ is the additive noise

Passing through the A/D converter the received signal is sampled to obtain

$$y[n] = x'[n] * h[n] + v[n], \qquad -\varepsilon \leq n \leq N-1 \qquad (2.4)$$

The N point discrete Fourier transform (DFT) is defined as

$$\begin{aligned} Y(k) &= \mathrm{DFT}\big(y(n)\big) \\ &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y(n) e^{-j(2\pi/N)kn}, \quad 0 \leq k \leq N-1 \end{aligned} \qquad (2.5)$$

The N QAM symbols obtained from the DFT has different phase shifts and amplitudes caused by frequency selective fading and analog front end imperfections. Because of these factors the received symbols can be incorrectly demodulated. To detect the QAM symbols correctly knowledge of the reference phase and amplitude of each sub carrier is needed.

2.2 IQ imbalance

OFDM systems are susceptible to analog front end imperfections; IQ imbalance is one of the impairments that cause the received symbols not to be correctly demodulated. The thesis is all about reduction of this problem in OFDM systems. Because of this front end imperfection the analog part is the expensive one in the system [30].

Down conversion is a basic phase in all radio frequency (RF) architectures. To convert RF signal into its equivalent baseband many configuration are used. The traditional OFDM system employs the Superhetrodyne receiver to convert RF signal into baseband and vice versa. This architecture use intermediate frequencies (IF) to convert RF signal down to baseband. The disadvantage of this configuration is it uses a lot of components which are expensive and external to get a good signal quality [22].

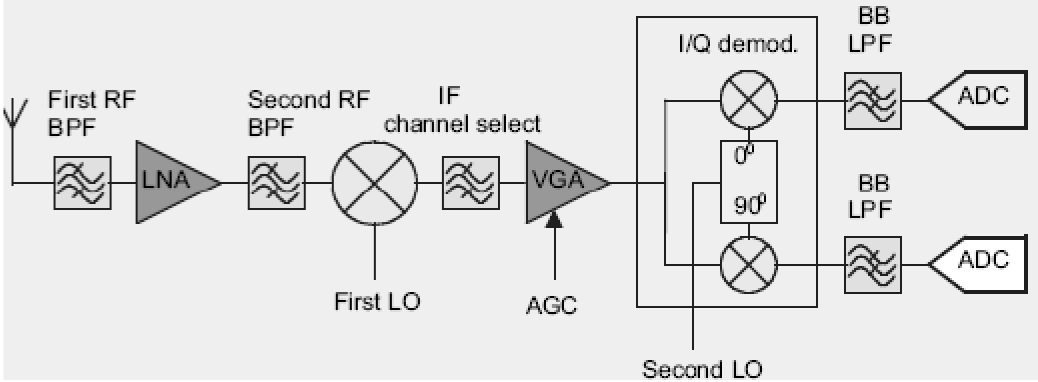

Figure 2.6 Block Diagram of Superhetrodyne receiver

Alternative to the Superhetrodyne receiver is the Zero-IF architecture which avoids the costly IF components. It converts the RF signal directly to its baseband equivalent without the need of intermediate frequencies. However Zero-IF architectures bring in additional severe front end alteration such as IQ imbalance [3].

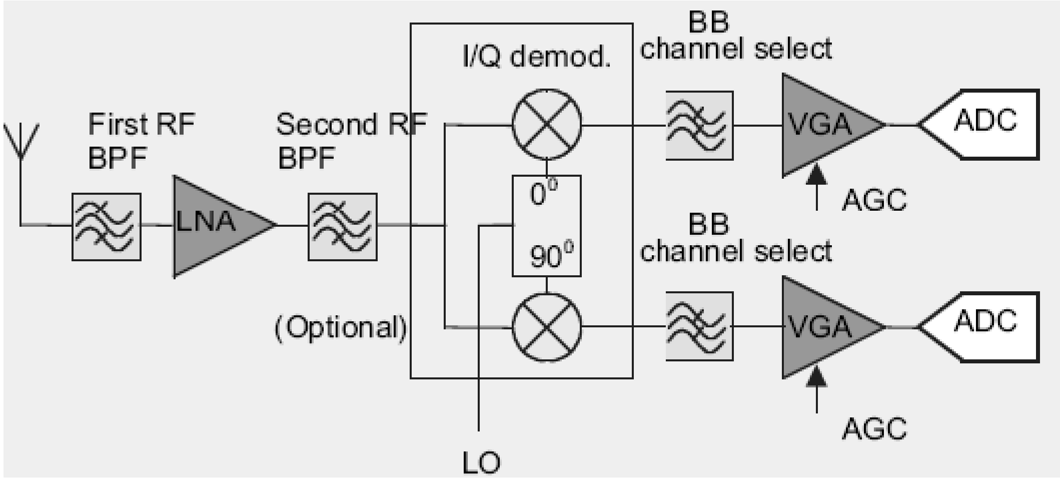

Figure2.7 Zero-IF architecture

In direct conversion (zero-IF receiver) analog IQ separation is carried out and IQ imbalance is almost inevitable. This is because the receiver is unable to generate correct quadrature phases with high-quality amplitude at higher carrier frequency [23].

IQ imbalance cause harsh deficit of demodulation performance [7].The consequence of IQ imbalance on OFDM receivers and several compensating methods were proposed [8-

11]. Reference [9] proposes non-adaptive time scheme for reduction of IQ imbalance in OFDM-WLAN receivers. Reference [10] points out the IQ imbalance can be eliminated based on one OFDM symbol and carries out well in the occurrence of phase noise. In [11] frequency domain is used and aims DVB with adaptive algorithm.

In this paper digital algorithms like least square and least mean square schemes are used for IQ compensation.

CHAPTER THREE

OFDM SYSTEMS WITH IQ IMBALANCE

3.1 Introduction

In chapter three the OFDM receiver with IQ imbalance impairments is considered and analysis on the basis of IQ imbalances will be given. A Model is configured based on the mathematical investigation of an IQ imbalance. In sub section 3.3 OFDM receivers with IQ imbalance is examined in a way to drive the compensation algorithms in the following chapters needed to tackle the impairments discussed.

3.2 The Model

The objective of any receiver is to convert the radio frequency to its baseband equivalent [3]. In the complex down conversion the RF signal is multiplied by the complex waveform $e^{-j2\pi fLo}$ where f_{LO} is the local oscillator at the receiver [4]. The received signal is split in I and Q. The I component is multiplied by the waveform at carrier frequency and the Q signal is multiplied by the same waveform with 90 degree phase shift. The mixture of I and Q signal provides the required baseband signal. The cosine and sine waveforms received has to be orthogonal i.e. they should have same amplitude and 90 degree phase difference. Any disparity that brings error in the 90 degree phase difference and on the amplitudes of I and Q component of the received signal will alter the signal constellation. The imbalance on the amplitudes of the IQ channels and the phase difference is termed as IQ imbalance [6]. To develop the mathematical model two cases are considered.

For reference purpose it is good to consider the ideal case which is without any IQ imbalance. The received signal is given by

$$y(t) = L_p(r(t)L_r(t)) \tag{3.1}$$

Where Lr(t)= $e^{-j2\pi fLot}$ denotes the complex waveform and LP{.} the low pass filter. In this thesis direct conversion architecture is used, so y(t) equals the received OFDM signal.

Achieving perfect IQ matching is impossible in practice [28]. IQ imbalance is generated by the components used in the phase and quadrature division along with analog to digital converter (ADC).

Let a_1 represent the amplitude of the I branch and a_2 represent the amplitude of the Q branch. The phase of I branch is defined to be $\theta_1$and the phase of Q branch is defined to be θ_2. Lr'(t) represent the imbalanced signal and can be expressed as:

$$L'_r(t) = a_1\cos(2\pi fct + \theta_1) - ja_2\sin(2\pi f_c\, t + \theta_2) \qquad (3.2)$$

From disparity point of view the above equation can be rearranged in more suitable way as follows

$$L'_r(t) = a_1[\cos(2\pi f_c t + \theta_1) - j(a_2/a_1)\sin(2\pi f_c\, t + \theta_1 + (\theta_2 - \theta_1))] \qquad (3.3)$$

For further simplification

Let $a_1=1$ and $a_2=a$; $\theta_1=0$ and $\theta_2= \theta$

Substituting these values in equation (3.3) and expressing the imbalanced local oscillator signal as:

$$L'_r(t) = \cos(2\pi f_c t) - ja\sin(2\pi f_c t + \theta) \qquad (3.4)$$

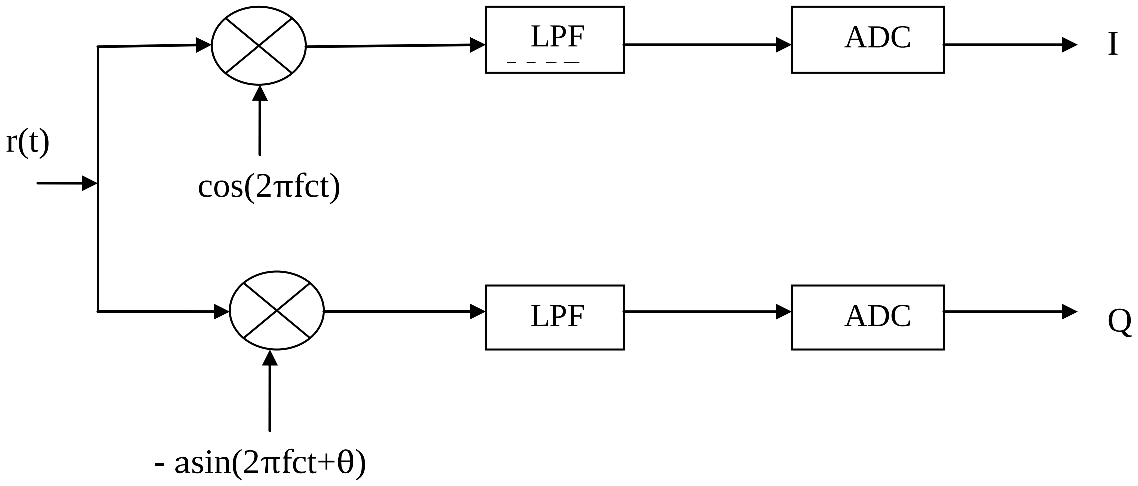

Figure 3.1 Imbalance Model

Applying Euler's formula for sine and cosine equation 3.4 can be expressed in a more descriptive way to understand the mismatch concept:

$$L'_r(t) = [\frac{e^{j2\pi fct}+e^{-j2\pi fct}}{2}] - ja[\frac{e^{j(2\pi fct+\theta)}-e^{-j(2\pi fct+\theta)}}{2j}] \qquad (3.5)$$

$$= \frac{(1+ae^{-j\theta})}{2}e^{-j2\pi fct} + \frac{(1-ae^{-j\theta})}{2}e^{j2\pi fct}$$

Let $\alpha = \frac{(1+ae^{-j\theta})}{2}$ and $\beta = \frac{(1-ae^{-j\theta})}{2}$

$$= \alpha e^{-j2\pi fct}+\beta e^{j2\pi fct} \qquad (3.6)$$

In this case the down conversion yields

$$y'(t) = L_p(r(t)L'_r(t))$$
$$= L_p\{[y(t)e^{j2\pi fct}+ y^*(t)e^{-j2\pi fct}][\alpha e^{-j2\pi fct}+\beta e^{j2\pi fct}]$$
$$y'(t) = \alpha y(t)+ \beta y^*(t) \qquad (3.7)$$

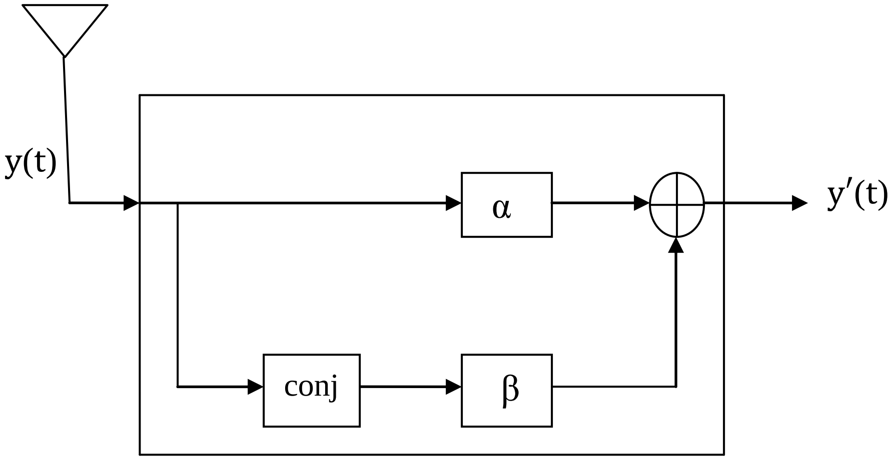

Figure3.2 Analog imbalance model with distortion parameters α and β

From figure 3.2 and equation 3.7 the received signal with IQ imbalance y'(t), is the sum of the required signal y(t) and its conjugate $y^*(t)$. The distortion parameters represent the amplitude and phase imbalance in the sine (Q) and cosine (I) branches.

3.3 Analysis of OFDM receivers with IQ imbalance

Quadrature Amplitude modulator is used in an OFDM transmitter to modulate its input signal. The resulting complex symbols $X[0], X[1],\ldots, X[N-1]$ passed through a set of serial to parallel converter to obtain a set of N QAM symbols. Using IFFT the discrete frequency components are converted into time samples by performing an inverse DFT of these symbols.

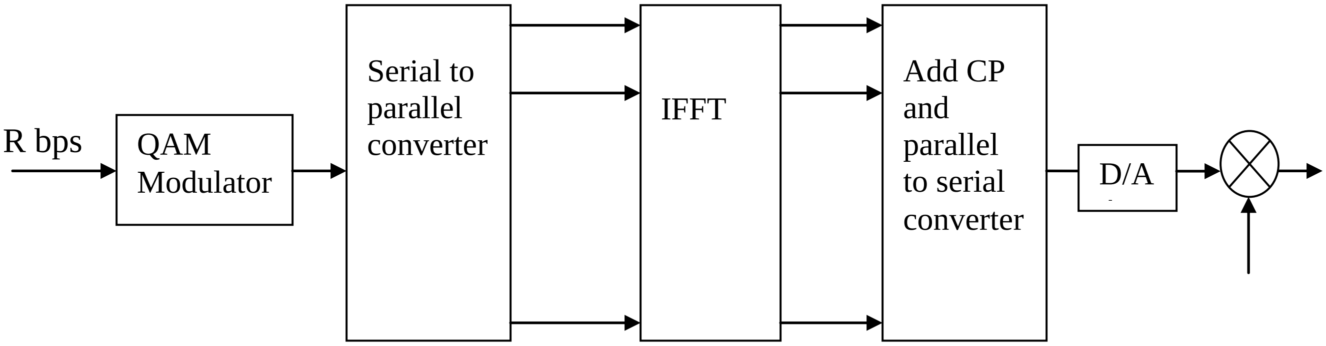

Figure 3.3 Considered OFDM transmitter

In Figure 3.3 the IFFT yields OFDM symbols consisting of the sequence $x[n] = x[0],\ldots, x[N-1]$ of length N:

$$\text{Where } x[n] = \frac{1}{\sqrt{N}}\sum_{i=0}^{N-1} X[i]e^{j2\pi ni/N} \quad 0 \leq n \leq N-1, \qquad (3.8)$$

Linearly modulated subchannels are included in the sequence of the multicarrier signal samples.Equation 3.8 can be represented by the matrix application [27]:

$$X = Mx \qquad (3.9)$$

Where: $X = (X[0], \ldots, X[N-1])^T$, $x = (x[0], \ldots, x[N-1])^T$, and
$M =$ an NxN matrix given by [18]

$$M = \frac{1}{\sqrt{N}}\begin{bmatrix} 1 & 1 & 1 & \cdots & 1 \\ 1 & W_N & W_{N^2} & \cdots & W_N^{N-1} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 1 & W_N^{N-1} & W_N^{2(N-1)} & \cdots & W_N^{(N-1)^2} \end{bmatrix} \qquad (3.10)$$

where $W_N = e^{-j2\pi/N}$ (3.11)

The resulting samples $x'[n] = x[-\varepsilon], \dots, x'[N-1] = x'[N-\varepsilon\,], \dots, x[0], \dots, x[N-1]$ are passed through parallel to serial converter and through digital to analog converter to obtain the baseband OFDM signal $x'(t)$ and up converted to frequency $f_c$and transmitted as $s(t)$. In the beginning no IQ imbalance is considered in which the transmitted signal pass through $h(t)$ and degraded by additive noise.

$$y(t) = x'(t) \otimes h(t) + v(t),\ \text{'}\otimes\text{' is convolution operation} \qquad (3.12)$$

The high frequency components are removed when the signal is down converted to baseband. The analog to digital converter samples the signals to obtain equation

$$y(t) = x'[n] \otimes h[n] + v[n] \qquad -\varepsilon \le n \le N-1 \qquad (3.13)$$

Now let us consider the discrete channel with FIR as described in section 2.1.5

$$h[n] = h[0] \dots, h[\varepsilon] \text{ of length } \varepsilon + 1 = {T_m}/{T_s} \qquad (3.14)$$

Where $T_m$the channel delay spread and $T_s$is the sampling interval. The n^{th} element of these sequences is indicated by $h_n = h[n],\ x'_n = x[n], v_n = v[n].$ the channel output can be written as equation 2.1 with these notations and can be expressed in matrix form as follows:

$$\begin{bmatrix} y_{N-1} \\ y_{N-2} \\ \vdots \\ y_0 \end{bmatrix} = \begin{bmatrix} h_0 & h_1 & \dots & h_\varepsilon & 0 & \dots & 0_{N+\varepsilon} \\ 0 & h_0 & \dots & h_{\varepsilon-1} & h_\varepsilon & \dots & 0 \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \dots & 0 & h_0 & \dots & h_{\varepsilon-1} & h_\varepsilon \end{bmatrix} \begin{bmatrix} x_{N-1} \\ \vdots \\ x_0 \\ x'_{-1} \\ \vdots \\ x'_{-\varepsilon} \end{bmatrix} + \begin{bmatrix} v_{N-1} \\ v_{N-2} \\ \vdots \\ v_0 \end{bmatrix} \qquad (3.15)$$

The symbols that are affected by the IS are removed since they have no impact in the recovery of the input signal. After the removal the last ε symbols of $x[n]$ the above matrix representation can be written as [25]:

$$\begin{bmatrix} y_{N-1} \\ y_{N-2} \\ \vdots \\ \vdots \\ y_0 \end{bmatrix} = \begin{bmatrix} h_0 & h_1 & \dots & h_\varepsilon & 0 & \dots & 0_{N+\varepsilon} \\ 0 & h_0 & \dots & h_{\varepsilon-1} & h_\varepsilon & \dots & 0 \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \dots & 0 & h_0 & \dots & h_{\varepsilon-1} & h_\varepsilon \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \vdots \\ h_2 & h_3 & \dots & h_{\varepsilon-2} & \dots & h_0 & h_1 \\ h_1 & h_2 & \dots & h_{\varepsilon-1} & \dots & 0 & h_0 \end{bmatrix} \begin{bmatrix} x_{N-1} \\ x_{N-2} \\ \vdots \\ \vdots \\ x_0 \end{bmatrix} + \begin{bmatrix} v_{N-1} \\ v_{N-2} \\ \vdots \\ \vdots \\ v_0 \end{bmatrix} \qquad (3.16)$$

Equation 3.16 can be written as in more general form as:

$$y = Hx + v \tag{3.17}$$

H is an $N \times N$ matrix which has an eigen value decomposition [25],

$$H = M^{*} \wedge M \tag{3.18}$$

Where $\wedge$ is a diagonal matrix that represents eigenvalues of H and M is a unitary matrix that the rows encompass the eigen vectors of H. That is$\lambda_i m_i = \ Hm_i$ for $i = 0,1,\dots,N-1$ where m_i is the i^{th} row of M. Hence

$$H = M^{*}\mathrm{diag}(\lambda)M \tag{3.19}$$

λ is associated to channel tap, $h' = [h_1,\ h_2,\dots,h_\varepsilon\ ]T$ via [24] [26],

$$\lambda = \sqrt{N}M^{*}\begin{bmatrix} h' \\ 0_{(N-(\varepsilon+1)\times 1)} \end{bmatrix} \tag{3.20}$$

Inserting equation 3.19 into 3.17 and rewriting equation 3.17 as:

$$y = M^{*}\mathrm{diag}(\lambda)Mx + v \tag{3.21}$$

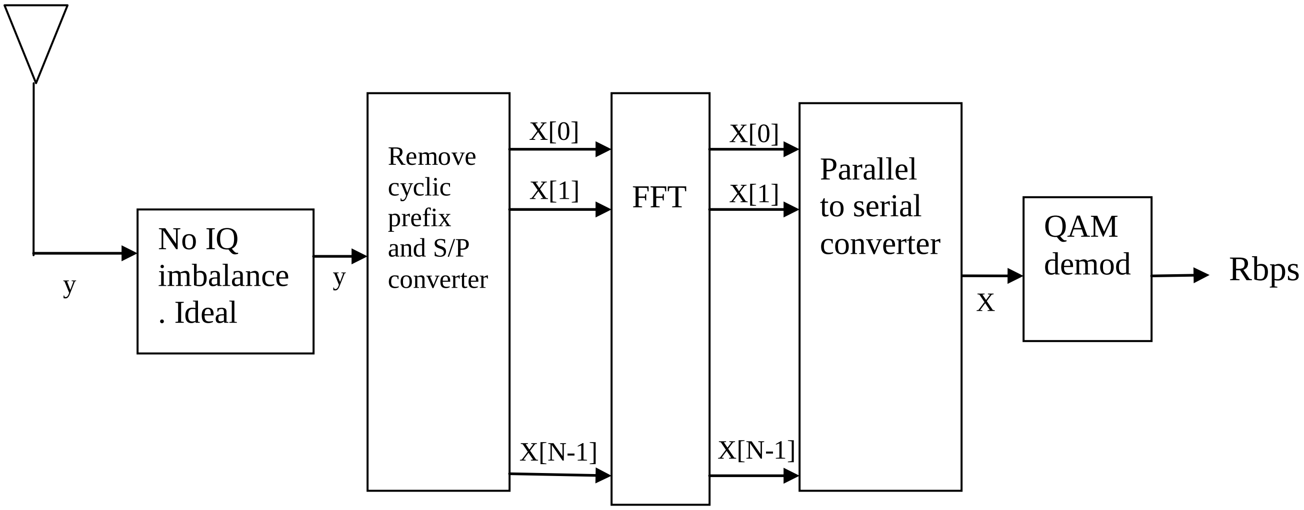

Figure 3.4 Ideal cases of OFDM receiver

In figure 3.4 it is indicated the signal y is passed through serial to parallel converter before it is demodulated by FFT.Now let us consider the case with IQ imbalance in which this imbalance distorts the received signal.

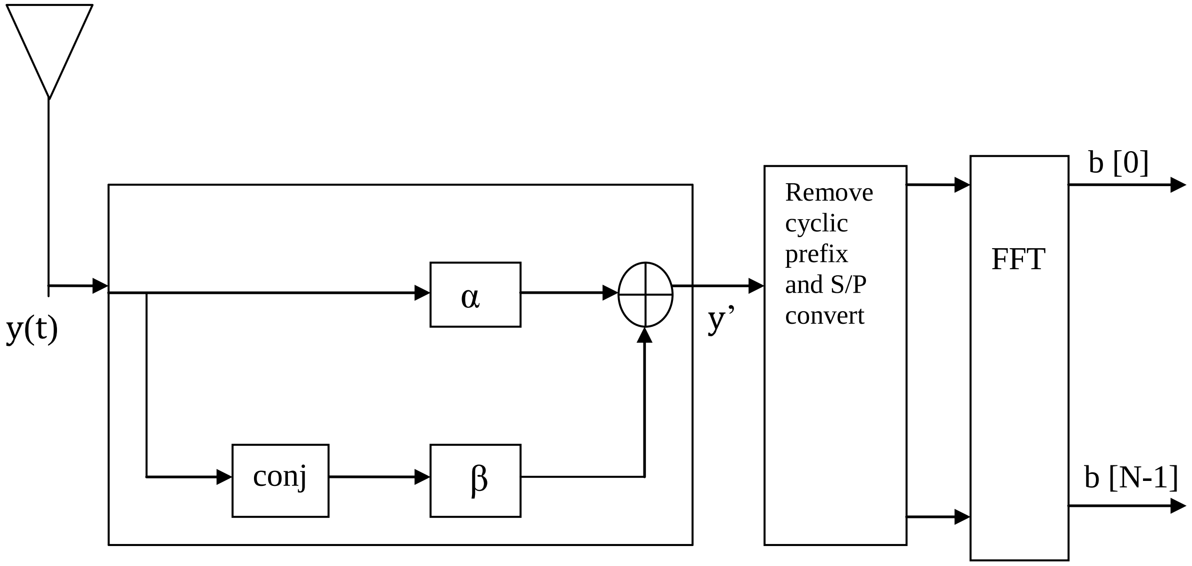

Figure 3.5 IQ imbalances in OFDM receivers

After the distortion the received signal y' is given by:

$$y' = \alpha y + \beta \mathrm{conj}(y) \qquad (3.22)$$

However,

$$\mathrm{conj}(y) = \mathrm{conj}(Hx + v) = \mathrm{conj}(H)\mathrm{conj}(x) + \mathrm{conj}(v) \qquad (3.23)$$

From [24] it is defined as

$$\mathrm{conj}(H) = \mathrm{conj}(M)\mathrm{diag}(\lambda')M \qquad (3.24)$$

where

$$\lambda' = \sqrt{N}M^{*}\begin{bmatrix} \mathrm{conj}(h') \\ 0_{(N-(\varepsilon+1)\times 1)} \end{bmatrix} \qquad (3.25)$$

Therefore

$$\mathrm{conj}(y) = M^{*}\mathrm{diag}(\lambda')M\mathrm{conj}(x) + \mathrm{conj}(v) \qquad (3.26)$$

Substituting equation 3.21 and equation 3.26 in equation 3.22 :

$$y' = \alpha[M^* \mathrm{diag}(\lambda) Mx + v] + \beta[M^* \mathrm{diag}(\lambda') M \mathrm{conj}(x) + \mathrm{conj}(v)] \quad (3.27)$$

As it can be seen from figure 3.5 the distorted signal passes through serial parallel converter and demodulated by FFT and it becomes

$$b = \alpha \mathrm{diag}(\lambda) X + \beta \mathrm{diag}(\lambda') X' + v' \quad (3.28)$$

where b, X, X' represent $My', Mx, M\mathrm{conj}(x)$ respectively and v' is the transformation of the original noise. From the properties of Discrete Fourier transform (DFT) [25] [31], for $1 \leq n \leq N$ and $1 \leq k \leq N$:

$$\text{If } X = Mx = \begin{bmatrix} X(1) \\ X(2) \\ \vdots \\ X\left(\frac{N}{2}\right) \\ X\left(\frac{N}{2}+1\right) \\ X\left(\frac{N}{2}+2\right) \\ \vdots \\ X(N) \end{bmatrix} \text{ then } X' = M\mathrm{conj}(x) = \begin{bmatrix} X^*(1) \\ X^*(N) \\ \vdots \\ X^*\left(\frac{N}{2}+2\right) \\ X^*\left(\frac{N}{2}+1\right) \\ X^*\left(\frac{N}{2}\right) \\ \vdots \\ X^*(2) \end{bmatrix} \quad (3.29)$$

Insert equation 3.29 into 3.28 and discard tones 1 and $\frac{N}{2} + 1$ for ease of representation. The reasons to remove the two samples 1 and $\frac{N}{2} + 1$ is because their alteration returns the same indices but the other tones are reflected around the tone $\frac{N}{2} + 1$ and conjugated [27].

Now let us define two vectors after discarding the unwanted tones
$X'' = [X(2), \ldots\ldots, X(N/2), X^*(N/2+2), \ldots., X^*(N)]^T$ And,

$$B = [b(2), \ldots\ldots, b(N/2), b(N/2+2), \ldots., b^*(N)]^T \quad (3.30)$$

Using equation 3.30, equation 3.28 can be written in more general form us

$$B = QX'' + V'' \quad (3.31)$$

Where Q is given by:

$$Q = \begin{bmatrix} \alpha\lambda(2) & & & & & & & \beta\lambda^*(N) \\ & \ddots & & & & & \cdot & \\ & & \alpha\lambda(\frac{N}{2}-1) & 0 & 0 & \beta\lambda^*(\frac{N}{2}+3) & & \\ & & & \alpha\lambda(\frac{N}{2}) & \beta\lambda^*(\frac{N}{2}+2) & & & \\ & & & \beta^*\lambda(\frac{N}{2}) & \alpha^*\lambda^*(\frac{N}{2}+2) & & & \\ & & \beta^*\lambda(\frac{N}{2}-1) & 0 & 0 & \alpha^*\lambda^*(\frac{N}{2}+2) & & \\ & \cdot & & & & & \ddots & \\ \beta^*\lambda(2) & & & & & & & \alpha^*\lambda^*(N) \end{bmatrix} \tag{3.32}$$

The main task is to extract the signal X” from the received signal B as described in equation 3.31. For $i = 2, \ldots, \frac{N}{2}$, equation 3.31 can be re written as follows:

$$\begin{bmatrix} b(i) \\ b^*(N-i+2) \end{bmatrix} = \begin{bmatrix} \alpha\lambda(i) & \beta\lambda^*(N-i+2) \\ \beta^*\lambda(i) & \alpha^*\lambda^*(N-i+2) \end{bmatrix} \begin{bmatrix} X(i) \\ X^*(N-i+2) \end{bmatrix} = \begin{bmatrix} v(i) \\ v^*(N-i+2) \end{bmatrix} \tag{3.33}$$

In more general form:

$$b_i = A_i X_i + v_i \tag{3.34}$$

Where

$$b_i = \begin{bmatrix} b(i) \\ b^*(N-i+2) \end{bmatrix}, A_i = \begin{bmatrix} \alpha\lambda(2) & \beta\lambda^*(N-i+2) \\ \beta^*\lambda(i) & \alpha^*\lambda^*(N-i+2) \end{bmatrix},$$

$$X_i = \begin{bmatrix} X(i) \\ X^*(N-i+2) \end{bmatrix}, \text{and } v_i = \begin{bmatrix} v(i) \\ v^*(N-i+2) \end{bmatrix}$$

The main objective is to extract $X(i)$ and $X^*(N-i+2)$ from the received signal b_i.

CHAPTER FOUR

DIGITAL METHODS FOR REDUCTION OF IQ IMBALANCE

4.1 Introduction

The previous chapter shows IQ imbalance in OFDM systems. Subsequently this chapter talks about the algorithms to minimize the consequence of IQ imbalance impairment in digital domain. The algorithms are Least Square (LS) and least mean square (LMS). Training method is used in both algorithms to approximate the distortion parameter that model the IQ imbalances. For some systems that do not offer training symbols other methods like decision direction is used [29]. The decision directed algorithm is carried out according to the status of the data to be poised for use in digital communication receiver. Two decision areas are created having an I-axis and Q-axis. Then a calculation is performed to determine a sign pair of error estimation data. The algorithm is then executed to update the equalization coefficients. The method is performed again to reduce the number of decision areas formed by the I and Q axis in which the symbol data is arranged [29].

4.2 Least Square Algorithms

In the previous chapter equation 3.34 the vector $\mathrm{b_i}$ and the noise component $\mathrm{v_i}$ are not in the column space of matrix $\mathrm{A_i}$. Least square algorithms determine the vector in the column space of $\mathrm{A_i}$ closest to $\mathrm{b_i}$ [27]. From definition of least square the variable that minimizes the distance between $\mathrm{b_i}$ and $\mathrm{A_iX_i}$ is determined by:

$$\min_{\mathrm{X_i}} \left\| \mathrm{b_i} - \mathrm{A_i}\vec{\mathrm{X}}_\mathrm{i} \right\|^2 \tag{4.1}$$

The solution of equation 4.1 $\vec{\mathrm{X}}_\mathrm{i}$ is called the least square solution and it provides the estimate for the unknown $\mathrm{X_i}$. It means ortogonality condition must be satisfied by the closest element. Hence,

$$\mathrm{A^T}.\left(\mathrm{b_i} - \mathrm{A_i}\vec{\mathrm{X}}_\mathrm{i}\right) = 0 \tag{4.2}$$

For $\mathrm{i} = 2, \dots, \frac{\mathrm{N}}{2}$ the least square estimate of $\mathrm{X(i)}$ and $\mathrm{X^*(N - i + 2)}$ designated by $\vec{\mathrm{X}}_\mathrm{i}$ and $\vec{\mathrm{X}}_\mathrm{i}^{\,*}(\mathrm{N - i + 2})$ are given by

$$\vec{\mathrm{X}}_\mathrm{i} = (\mathrm{A_i^T}.\mathrm{A_i})^{-1}.\mathrm{A_i^T}.\mathrm{b_i} \tag{4.3}$$

Implementing equation 4.3 would require the alteration parameters described in the previous chapter (α and β) with the channel information λ. Training symbols are needed to allow the receiver to guess those values. From equation 3.33 for channel evaluation and by re writing it as:

$$\begin{bmatrix} \mathrm{b}(\mathrm{i}) \\ \mathrm{b}^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix} = \begin{bmatrix} \mathrm{X}(\mathrm{i}) & 0 & \mathrm{X}^*(\mathrm{N}-\mathrm{i}+2) & 0 \\ 0 & \mathrm{X}(\mathrm{i}) & 0 & \mathrm{X}^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix} . \begin{bmatrix} \alpha\lambda(\mathrm{i}) \\ \beta^*\lambda(\mathrm{i}) \\ \beta\lambda^*(\mathrm{N}-\mathrm{i}+2 \\ \alpha^*\lambda^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix} + \begin{bmatrix} \mathrm{v}(\mathrm{i}) \\ \mathrm{v}^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix} \qquad (4.4)$$

Let $\varphi_i = \begin{bmatrix} \mathrm{X}(\mathrm{i}) & 0 & \mathrm{X}^*(\mathrm{N}-\mathrm{i}+2) & 0 \\ 0 & \mathrm{X}(\mathrm{i}) & 0 & \mathrm{X}^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix}$ and $\emptyset_i = \begin{bmatrix} \alpha\lambda(\mathrm{i}) \\ \beta^*\lambda(\mathrm{i}) \\ \beta\lambda^*(\mathrm{N}-\mathrm{i}+2 \\ \alpha^*\lambda^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix}$

And $v_i = \begin{bmatrix} \mathrm{v}(\mathrm{i}) \\ \mathrm{v}^*(\mathrm{N}-\mathrm{i}+2) \end{bmatrix}$

So equation 4.4 can be written as:

$$\mathrm{b}(\mathrm{i}) = \varphi_i \emptyset_i + v_i \qquad (4.5)$$

The least square estimation of channel taps and distortion IQ parameters calculated from eq. 4.5 as

$$\overrightarrow{\emptyset_i} = \varphi_i^{T}(\varphi_i . \varphi_i^{T})^{-1} . b_i \qquad (4.6)$$

Implementing equation 4.6 for n OFDM symbols used for training can be pulled collectively to approximate A_i using least square algorithm. The receiver structure with the least square algorithm looks like:

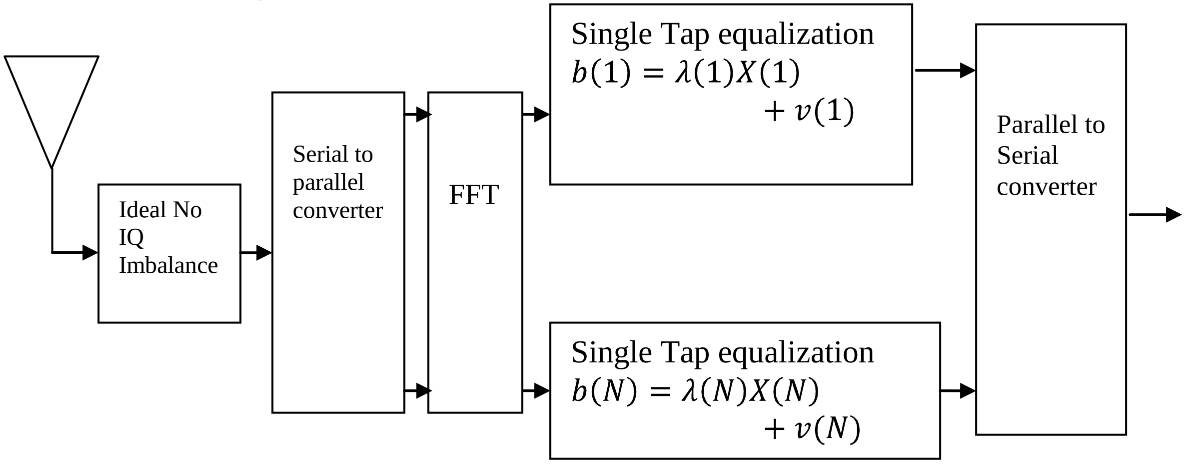

Figure 4.1 OFDM receivers with No IQ imbalance

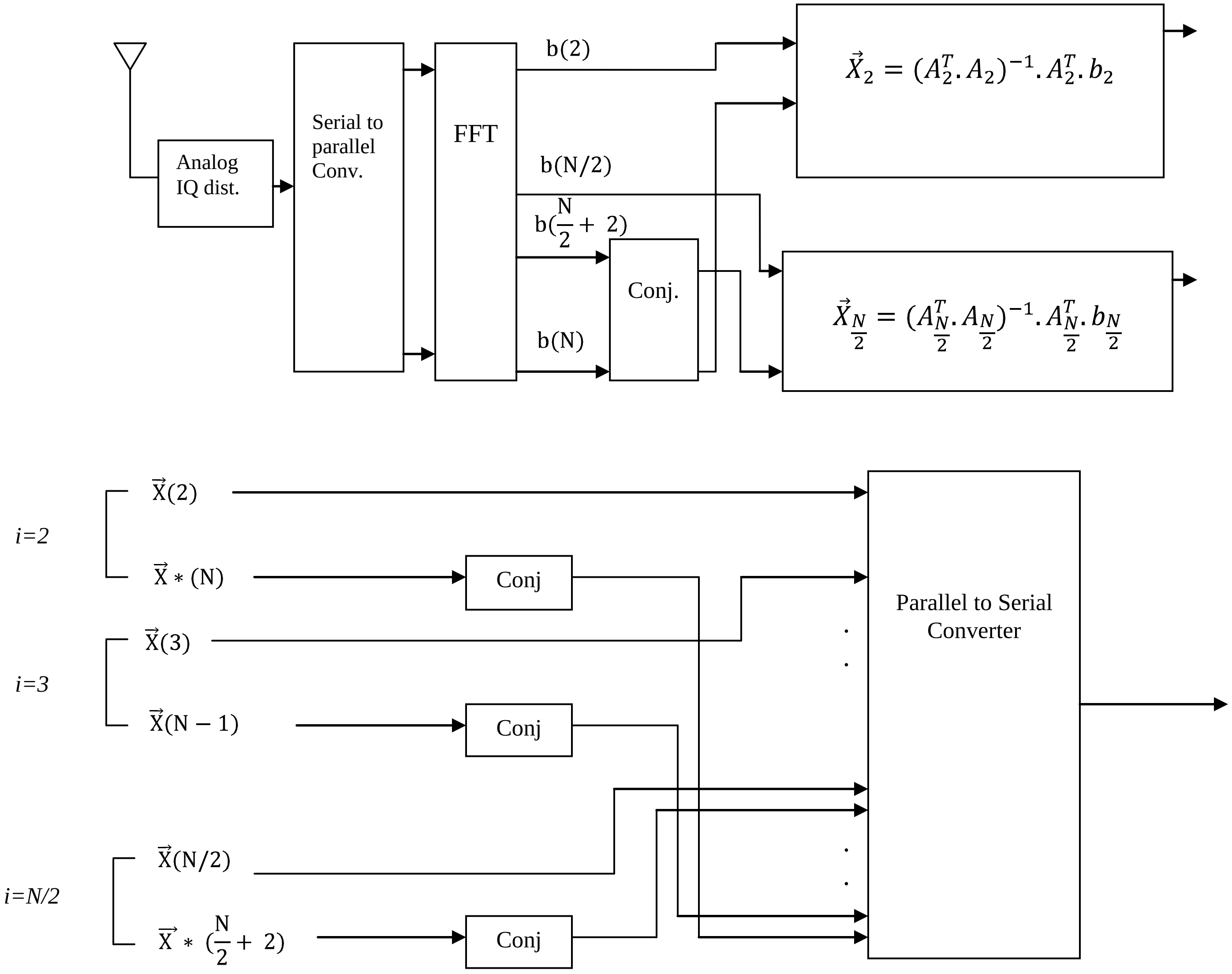

Figure 4.2 OFDM receivers with least square reduction of IQ imbalance

4.3 Least Mean Algorithm (LMS)

The least mean approximation of $\mathrm{X(i)}$ and $\mathrm{X^*(N-i+2)}$ can be accomplished as follows:

$$\vec{\mathrm{X}}(\mathrm{i}) = \mathrm{w_i}.\begin{bmatrix}\mathrm{b(i)} \\ \mathrm{b^*(N-i+2)}\end{bmatrix} \tag{4.7}$$

$$\vec{\mathrm{X}}^*(\mathrm{N-i+2}) = \mathrm{w_{N-i+2}}.\begin{bmatrix}\mathrm{b(i)} \\ \mathrm{b^*(N-i+2)}\end{bmatrix} \tag{4.8}$$

Where $\mathrm{w_i}$ and $\mathrm{w_{N-i+2}}$ are equalization vectors

To emphasize the equation more iteration index k is introduced. Hence at time instant k, $\mathrm{w_i^k}$ and $\mathrm{w_{N-i+2}^k}$ correspond to the equalization expressions. And with the same k the terms $\mathrm{b(i)}$ and $\mathrm{b^*(N-i+2)}$ are represented by $\mathrm{b^k(i)}$ and $\mathrm{b^{*(k)}(N-i+2)}$. The coefficients for $\mathrm{i} = 2, \dots, \frac{\mathrm{N}}{2}$ are given by:

$$\mathrm{w_i^{(k+1)}} = \mathrm{w_i^{(k)}} + \mathrm{\mu_{LMS}}.\begin{bmatrix}\mathrm{b^k(i)} \\ \mathrm{b^{*(k)}(N-i+2)}\end{bmatrix}^{\mathrm{H}}.\mathrm{e_i^{(k)}} \tag{4.9}$$

$$\mathrm{w_{N-i+2}^{(k+1)}} = \mathrm{w_{N-i+2}^{(k)}} + \mathrm{\mu_{LMS}}.\begin{bmatrix}\mathrm{b^k(i)} \\ \mathrm{b^{*(k)}(N-i+2)}\end{bmatrix}^{\mathrm{H}}.\mathrm{e_{N-i+2}^{(k)}} \tag{4.10}$$

For $0 \le \mathrm{\mu_{LMS}} \le \frac{2}{\mathrm{\lambda_{max}}}$ the algorithm converges to its mean. Where $\mathrm{\mu_{LMS}}$ is the step size parameter and $\mathrm{\lambda_{max}}$ the maximum Eigen value

Defining a training symbol $\mathrm{d_i^{(k)}}$ at iteration index k for the creation of the error signal and can be expressed as:

$$\mathrm{e_i^{(k)}} = \mathrm{d_i^{(k)}} - \mathrm{w_i^{(k)}}.\begin{bmatrix}\mathrm{b^k(i)} \\ \mathrm{b^{*(k)}(N-i+2)}\end{bmatrix} \tag{4.11}$$

$$\mathrm{e_{(N-i+2)}^{(k)}} = \mathrm{d_{(N-i+2)}^{(k)}} - \mathrm{w_{(N-i+2)}^{(k)}}.\begin{bmatrix}\mathrm{b^k(i)} \\ \mathrm{b^{*(k)}(N-i+2)}\end{bmatrix} \tag{4.12}$$

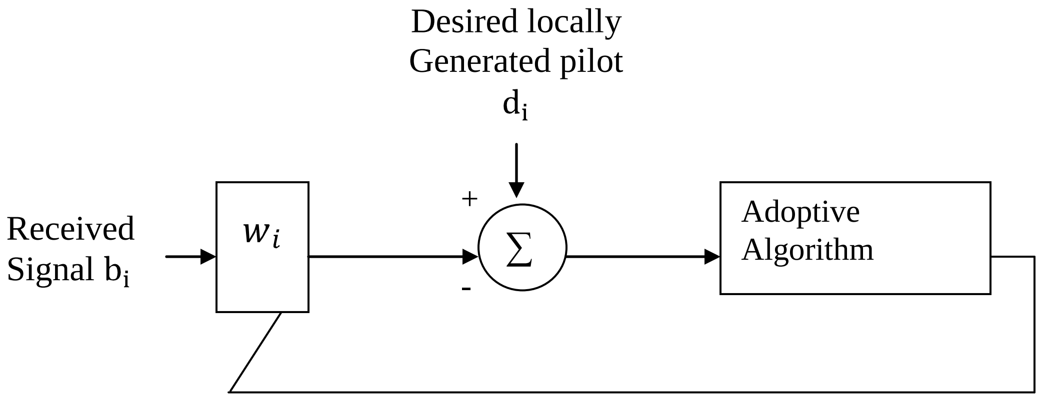

Figure 4.3 Least Mean Algorithm techniques

The slow convergence rate is the problem with least mean square algorithm. Current OFDM systems apply short length for training symbols in order to minimize training overhaul in packet delivery. For ideal IQ branches short training length is acceptable to achieve good estimation. However the existence of IQ imbalance brings cross coupling between every tone and its mirrored equivalent. This slows down the convergence rate. In the above equations 4.11 and 4.12 the initials values of the coefficients usually set to zero, to improve the convergence new calculated values are assigned to the coefficients by assuming ideal I and Q branches.

To calculate the initial values the distortion parameters from equation 3.33 α and β are set to 1 and 0 , and the equation is reduced to:

$$\begin{bmatrix} b(i) \\ b^*(N-i+2) \end{bmatrix} = \begin{bmatrix} \alpha\lambda(i) & 0 \\ 0 & \lambda^*(N-i+2) \end{bmatrix} \begin{bmatrix} X(i) \\ X^*(N-i+2) \end{bmatrix} = \begin{bmatrix} v(i) \\ v^*(N-i+2) \end{bmatrix} \tag{4.13}$$

The least square solution of $X(i)$ and $X^*(N-i+2)$ for n OFDM symbols used for training is given by

$$\vec{\lambda}(i) = \frac{\sum_{k=1}^{n} X_k^*(i) b_k(i)}{\sum_{k=1}^{n} X_k^*(i) X_k(i)} \tag{4.14}$$

Similarly for $N-i+2$

$$\vec{\lambda}(N-i+2) = \frac{\sum_{k=1}^{n} X_k^*(N-i+2) b_k(N-i+2)}{\sum_{k=1}^{n} X_k^*(N-i+2) X_k(N-i+2)} \tag{4.15}$$

The received and the transmitted i^{th} tones at time instant k are represented by $b_k(i)$ and $X_k^*(i)$ respectively

Therefore using the above equation 4.14 and 4.15 for channel parameter and initializing the equalization vectors in eq. 4.9 and 4.10 as:

$$\mathrm{w}_{\mathrm{i}}^{(0)} = \begin{bmatrix} \frac{1}{\vec{\lambda}(\mathrm{i})} & 0 \end{bmatrix} \text{ And } \mathrm{w}_{(\mathrm{N}-\mathrm{i}+2\,)}^{(0)} = \begin{bmatrix} 0 & \frac{1}{\vec{\lambda}(\mathrm{N}-\mathrm{i}+2\,)} \end{bmatrix} \qquad (4.16)$$

Using equation 4.16 as initial values for estimation and calculating the LMS solution that gives us fast converged result than using zero as initial value.

CHAPTER FIVE

SIMULATION RESULTS

A distinctive OFDM system is simulated based on some given parameters to assess the performance parameters of the estimation methods in contrast with perfect OFDM receivers with no IQ imbalance. The parameters considered are:

- Guard time assumed to be 4 times the delay spread around 800ns
- OFDM symbol duration assumed to be 4 times the guard time which is 4µs
- Bit rate to be 20Mps,
- Band width ≤20MHZ
- 16QAM and 64QAM are used to determine the required number of subcarriers
- OFDM symbol length i.e number of FFT points 64
- 16 is chosen for the cyclic prefix
- Channel length is assumed to be $\varepsilon + 4$

The BER versus SNR for the proposed digital methods are depicted below in figures 5.1 to 5.4. No IQ imbalance is referred to as Ideal IQ, IQ imbalance without compensation refers to the existence of IQ imbalance and no compensation technique is used. Besides these graphs IQ constellation for the transmitted signal is shown for SNR 25dB. From the last figure it is shown that for large phase imbalance increment from 0-5degrees, SNR 20dB and amplitude imbalance at 1dB the schemes employed brought an acceptable result. In the constellation diagrams scheme a refers to least square (LS) algorithm and scheme b refers to the least mean algorithm (LMS)

The selection of different OFDM parameters is a tradeoff between different needs. Usually, the three important parameters to start with are: Bandwidth, bit rate, and delay spread. The guard time directly stated by the delay spread. Commonly, the guard time is selected about two to four times delay spread [31]. The guard time depends on the type of coding and QAM modulation.

Inter carrier interference (ICI) and inter symbol interference (ISI) affect Higher order QAM (like 64-QAM) than QPSK while heavier coding evidently trim down the sensitivity to such interference [31]. After the guard time is configured the symbol duration can be set. The guard time causes the signal to noise ratio (SNR) loss. It is important to make the symbol duration much larger than the guard time. However, large symbol duration brings subcarriers having small spacing. Because of this implementation complexity is inevitable and is easily affected by phase noise and frequency offset which result in high peak to average power ratio. Hence to avoid all these problems the symbol duration has to be five times the guard time.

After setting the guard time and symbol duration the subcarriers can be determined by taking the ratio of the required bit rate the symbol duration to the bit rate per sub carrier. The modulation type, symbol rate and coding rate define the bit rate per sub carrier.

From the above figures 5.1-5.4 we can see saturation in BER due to the existence of IQ imbalance. The compensation mechanisms are required because improving the BER values by increasing the operating SNR could not bring a significant improvement. Figures 5.5-5.7 illustrates how the transmitted constellation, the received constellation and the extraction of the transmitted signal from the received signals.

It is seen that the digital algorithms we tried to compensate the IQ imbalance achieve close to the ideal case. From the figure 5.8 it is shown that for large phase imbalance the schemes we employ brought an acceptable result.

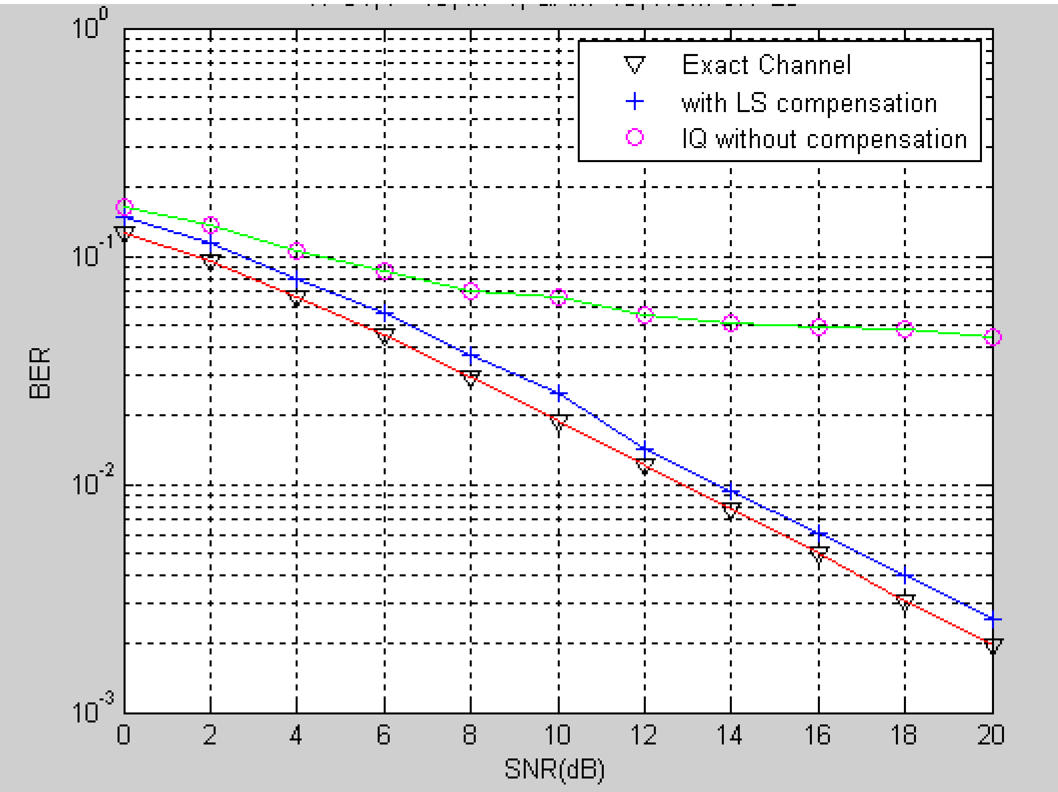

Figure 5.1 BER versus SNR for 16QAM

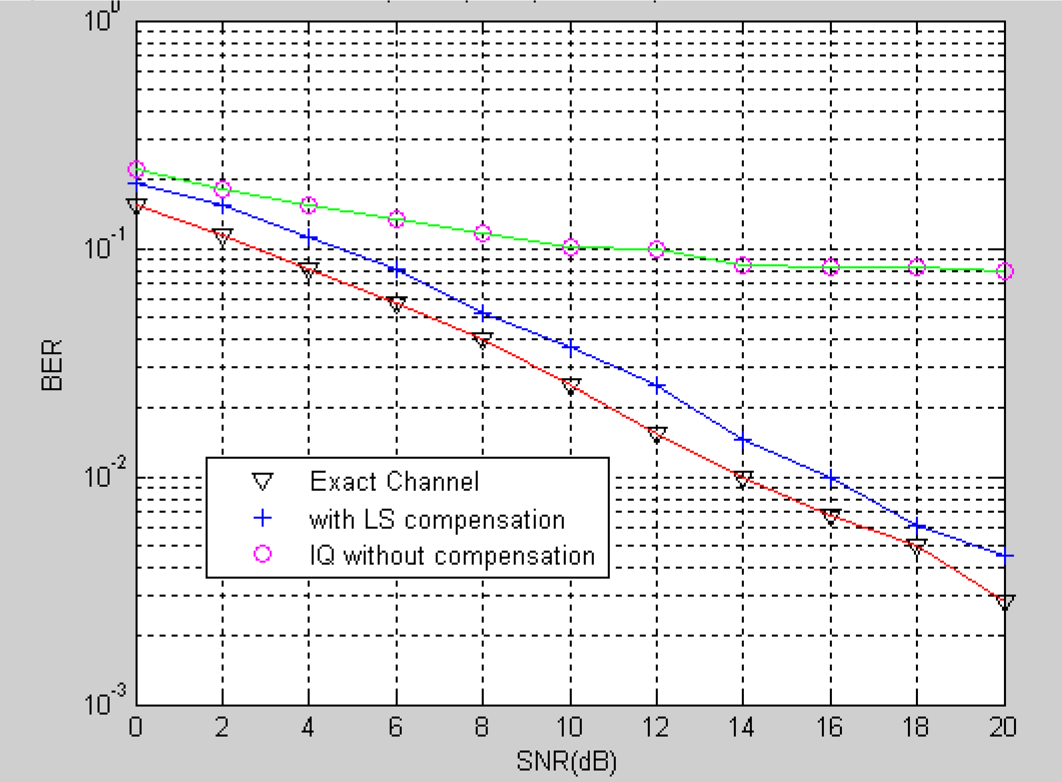

Figure 5.2 BER versus SNR for 64 QAM

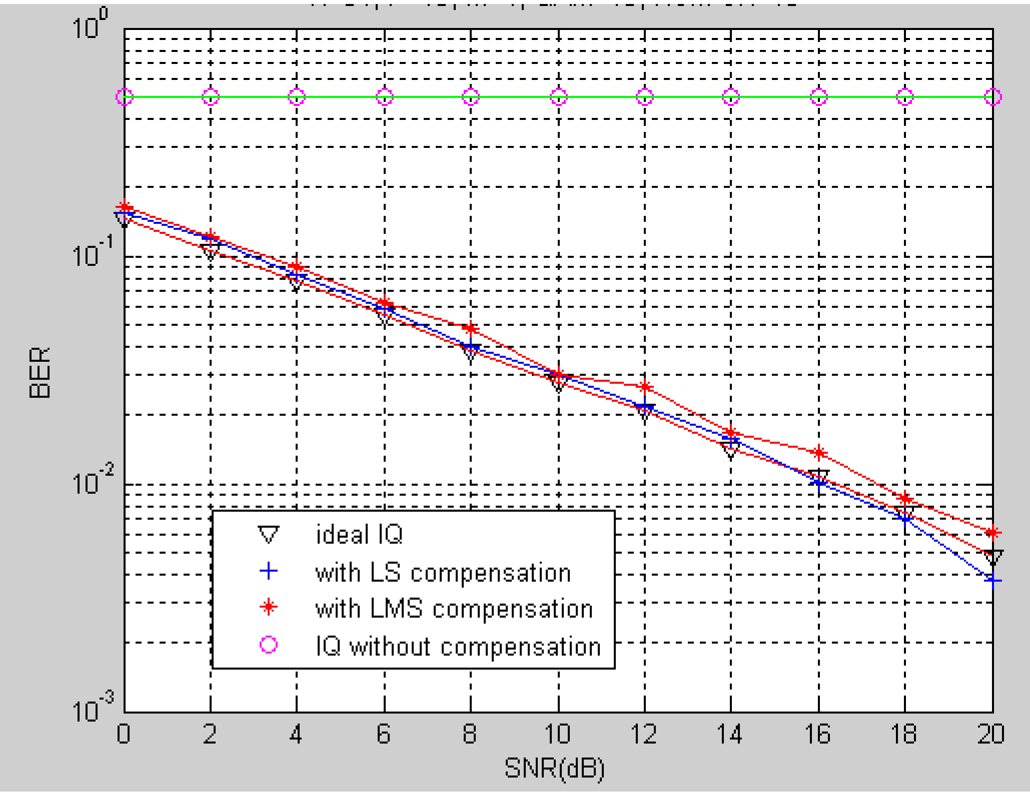

Figure 5.3 BER versus SNR for 16 QAM

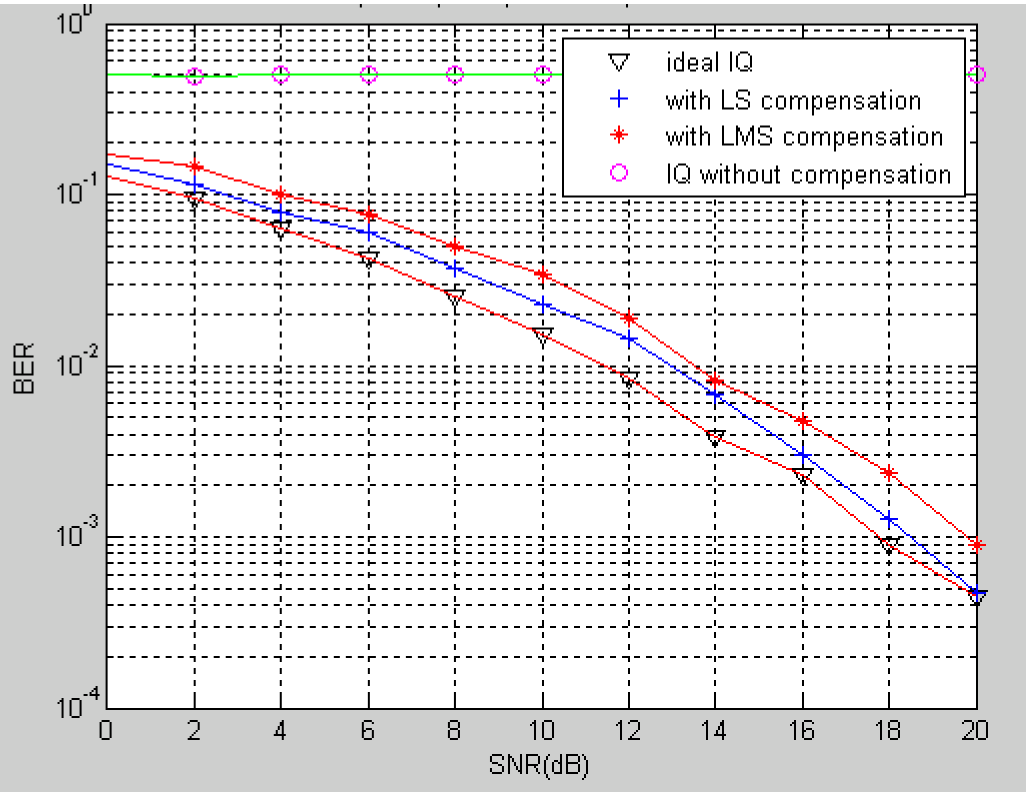

Figure 5.4 BER versus SNR for 64 QAM

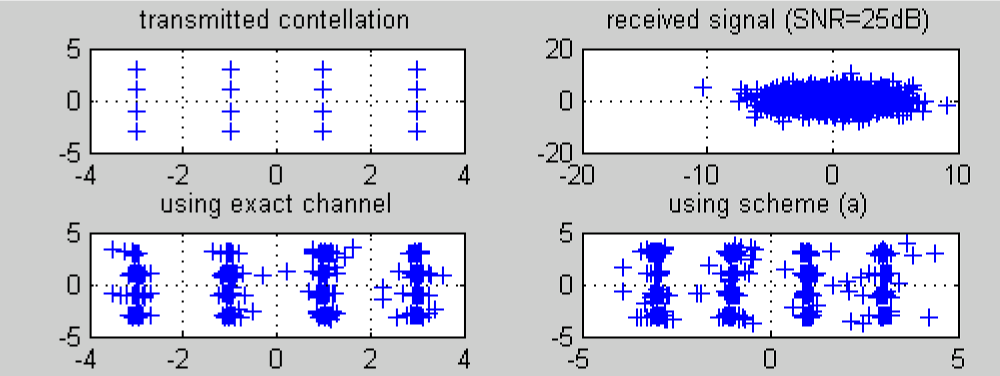

Figure 5.5 Constellation diagram for 16QAM

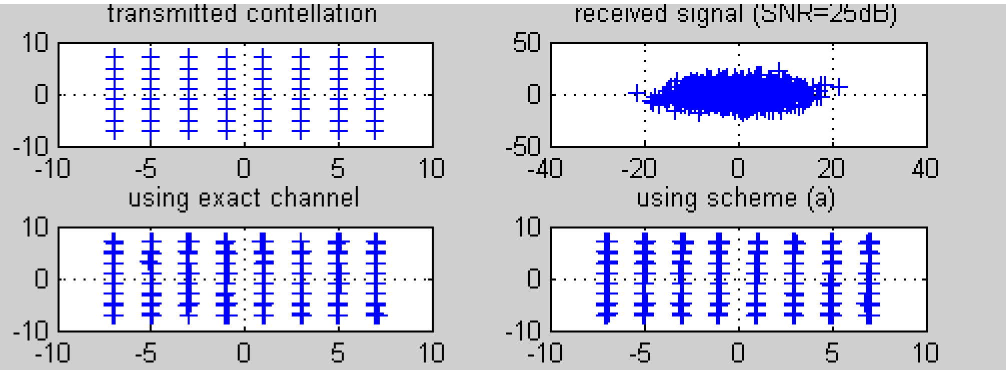

Figure 5.6 Constellation diagram for 64QAM

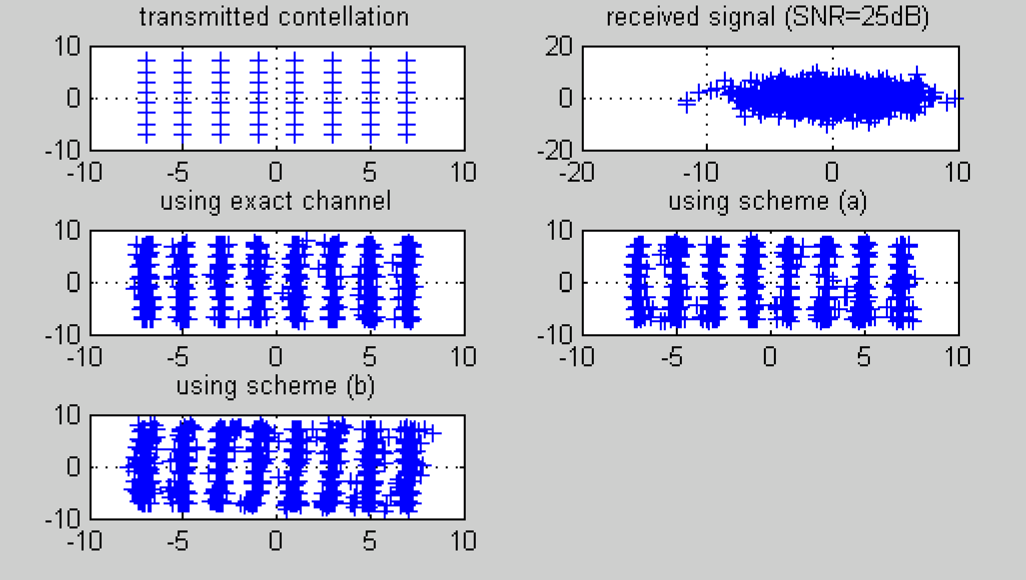

Figure 5.7 Constellation diagram for 64 QAM.

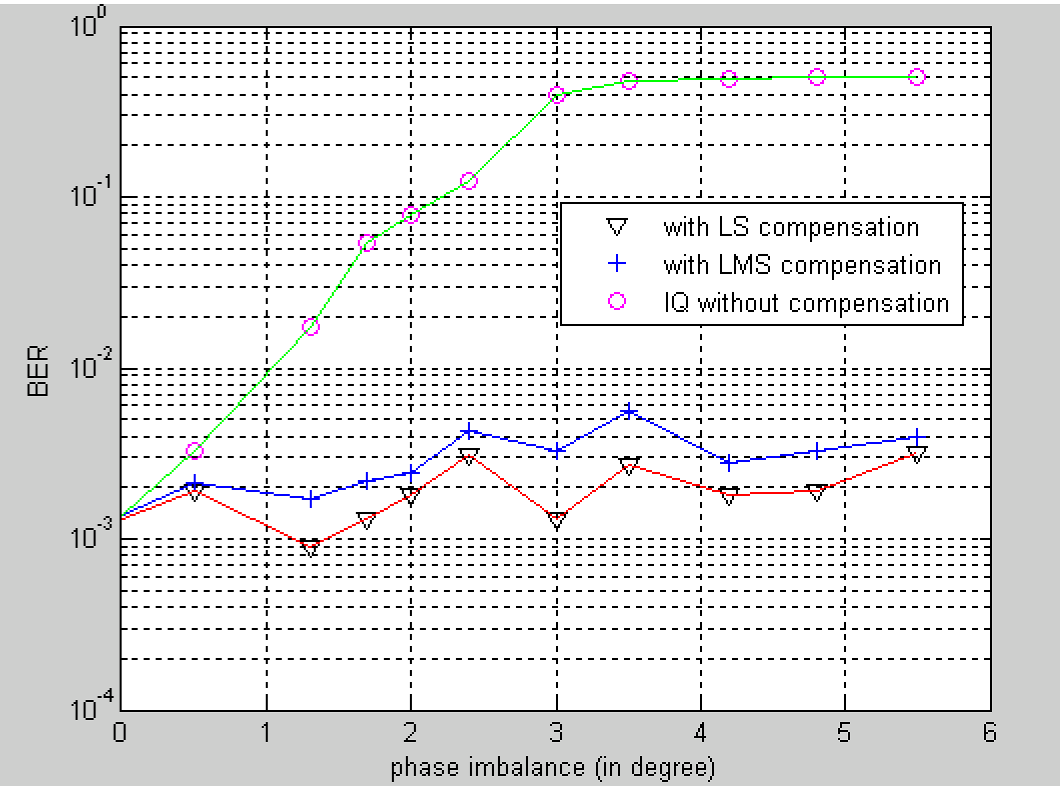

Figure 5.8 phase imbalance variation from 0 to 5 degree for 16QAM constellations

CHAPTER SIX

CONCLUSION AND FUTURE WORK

6.1 Conclusion

The recent development in wireless design is to reduce the analog front-end processing problems and apply most of the basic functions using digital signal processing. This is mainly inspired by the need as to multiple access wireless standards and services using simplified and cheap equipments. As a result the customary superheterodyne receiver is not the most suitable option and new receiver structures like direct conversion receiver has to be considered. In reality, IQ imbalances are inevitable in the analog front end, which results in limited and frequently inadequate rejection of the image frequency band. This brings the image signal to come out as intervention on the top of the desired signals.

Direct conversion receivers are used with significant IQ imbalances to down convert RF signal to baseband. After down conversion the algorithms are tested on the down converted signals. To illustrate the success of the algorithms a comparison is made on the performance with and without the compensation methods. It was demonstrated for OFDM systems having significant IQ imbalances the BER values may be intolerable. A notable outcome of IQ imbalances is that the attainable BER saturates as the SNR increases, signifying the occurrence of the IQ imbalance restrict the systems performance at high SNR. The performance deteriorates and becomes harsher at high SNR values and high density constellations.

In chapter four the figures shows the effectiveness of the digital compensation methods. As discussed in chapter five the BER saturates as the SNR increases, showing that the presence of IQ imbalance hamper the system performances at high SNR. For high SNR values the degradation becomes high.

The methods employed in this thesis take up digital signal processing techniques to reduce the impairments in the digital domain. In this thesis the main contribution is stated as, least square (LS) and least mean algorithm (LMS) methods are used to compensate the analog impairments (IQ imbalance). Several advantages are linked with the digital methods, with the main advantage looms in making use of the digital processing influence to ease the analog imperfections.

Zero IF receivers design normally gives an IQ imbalance in order of phase imbalance of one to two degrees and amplitude one to two percent [19]. In chapter five it is verified the need of compensation techniques to facilitate high data rate communications.

6.2 Future Work

The maximum transmission speed in IEEE 802.11a WLAN is 54Mbps, but it is reduced to 20~30Mbps due to multipath and faulty hardware design such as IQ imbalance from direct conversion receivers. The need of compensation algorithms is needed because IQ imbalance is a serious interference in higher order QAM modulation.

In MIMO systems higher data throughput can be achieved by allowing higher constellation sizes. The system offers higher SNR but at higher SNR IQ imbalance hampers the BER performance. Thus studding MIMO OFDM systems and tackling such impairments to achieve higher data throughput is an attractive topic to look into.

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