ADAPTIVE CHANNEL ESTIMATION FOR LTE UPLINK

A THESIS SUBMITTED IN PARTIAL FULFILLMENT
OF THE REQUIREMENTS FOR THE DEGREE OF

Master of Technology
In
Communication and Signal Processing
By

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Department of Electronics & Communication Engineering
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CERTIFICATE

This is to certify that the thesis entitled, "**ADAPTIVE CHANNEL ESTIMATION FOR LTE UPLINK** " submitted by DIVYA VIJAYAN (211EC4112) in partial fulfillment of the requirements for the award of Master of Technology degree in Electronics and Communication Engineering with specialization in "Communication and Signal Processing" at National Institute of Technology, Rourkela (Deemed University) and is an authentic work by her under my supervision and guidance. To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other university/institute for the award of any Degree or Diploma.

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Divya Vijayan

ABSTRACT

Third generation partnership project (3GPP) long term evolution (LTE) uses single carrier frequency division multiple access (SC-FDMA) in uplink transmission and orthogonal frequency division multiple access (OFDMA) scheme for the downlink. A variable step size based least mean squares (LMS) algorithm is formulated for a single carrier frequency division multiple access (SC-FDMA) system, in its channel estimation (CE). The weighting coefficients on the channel condition can be updated using this unbiased CE method. Channel and noise statistics information are not essential. Rather, it uses a phase weighting scheme to eliminate the signal fluctuations due to noise and decision errors. The convergence towards the true channel coefficient is guaranteed. The proposed algorithm is compared with the existing algorithm for BER and MSE performance in different channel environments.

ACRONYM

1G	First Generation
2G	Second Generation
3G	Third Generation
3GPP	Third Generation Partnership Project
4G	Fourth Generation
AMPS	Advanced Mobile Phone System
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CE	Channel Estimation
CDMA	Code Division Multiple Access
WCDMA	Wideband Code Division Multiple Access
DFDMA	Distributed Frequency Division Multiple Access
DFT	Discrete Fourier Transform
DVB	Digital Video Broadcasting
EDGE	Enhanced Data Rates for GSM Evolution
FDD	Frequency Division Duplex
FDMA	Frequency Division Frequency Division
GPRS	General Packet Radio Service
GSM	Global System for Mobile
IDFT	Inverse Discrete Fourier Transform

IFDMA	Interleaved Frequency Division Multiple Access
ICI	Inter Channel Interference
IP	Internet Protocol
ISI	Inter Symbol Interference
LFDMA	Localized Frequency Division Multiple Access
LMS	Least Mean Square
LTE	Long Term Evolution
MMSE	Minimum Mean Square Error
MSE	Mean Square Error
OFDMA	Orthogonal Frequency Division Multiple Access
PAPR	Peak to Average Power Ratio
QPSK	Quaternary Phase Shift Keying
RLS	Recursive Least Square
SCFDMA	Single Carrier Frequency Division Multiple Access
SMS	Short Message Service
TACS	Total Access Communication Systems
TDD	Time Division Duplex
TDMA	Time Division Multiple Access
VSS	Variable Step Size
WIMAX	World Wide Interoperability for Microwave Access
ZF	Zero Force

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CHAPTER-1

INTRODUCTION

1.1 Evolution Of Wireless Network

First Generation (1G) network focused on mobilizing landline telephony. The outcome networks, were Advanced Mobile Phone Systems (AMPS) in US and Total Access Communication Systems (TACS) in Europe. 1G is circuit switched that aided analog voice transmission over air. It employed Frequency Division Multiple Access (FDMA). The main drawbacks of 1G was its high sensitivity in dynamic environment and degraded quality.

Second Generation (2G) network supported data transmission along with enhancement in voice transmission. It was based on digital transmission via circuit switching. Multiple users could use the same channel by means of multiple access. 2G introduced Short Message Service(SMS). 2G comprised of Global System for Mobile (GSM) communication and Interim Standard 95 (IS- 95), which is known as (cdma one) commercially. GSM relied mostly on TDMA and cdma-one utilized Code Division Multiple Access(CDMA).Universal frequency reuse plan is used in cdma-one, in which same frequency can be reused in every cell, since the voice channels are distinguished by unique codes; it ensures better network capacity than CDMA system.

2.5G wireless communication network consists of GPRS, where each mobile network is assigned an IP address, which could be static, determined by cellular operator or dynamic, dependent on per connection basis. GPRS (General Packet Radio Service) is a packet-based radio technology for GSM networks. It enables omnipresent wireless Internet and other high-speed data communications such as SMS, MMS, email, games.

Generation	Year	Network	Technology	Data
1G	Early 1980s	Circuit Switched	TACS,AMPS	Analog voice
2G	Early 1990s	Circuit Switched	D-AMPS,GSM,CDMA	Digital voice
2.5G	1996	Circuit Switched/Packet Switched	GPRS,EDGE,EVDO, EVDV	Digital voice+ Data
3G	2000	Non-IP Packet Switched/Circuit Switched	CDMA 2000, WCDMA	Digital voice+ High speed data+ Video
4G	2012	IP based, Packet switched core network	WiMAX , LTE	Digital voice+ High speed data+ multimedia security

Table1.1: Comparison of different wireless cellular networks

3G networks are designed to deliver services with transmission rates beyond 2.5G systems that can support multimedia, data & video along with voice. 3G is the current generation of mobile telecommunication standards. It offers data rates of up to 2 Mbps. There are a bunch of technologies that fall under 3G, like WCDMA, EV-DO, and HSPA and others. UMTS, cdma 2000, and EDGE support 3G services.

4G wireless communication network supports for all prior 2G/3G features, in addition to that it facilitates high quality streaming video, high quality videoconferencing, high quality Voice-over-IP (VoIP). Wi-MAX (Worldwide Interoperability for Microwave Access), UMB (Ultra Mobile Broadband) and LTE are the outcomes of 4G.

1.2 Literature Review

To mitigate inter-channel interference (ICI) in the uplink direction of 3GPP LTE system, several channel estimation techniques have been proposed. Minimization of the mean square error (MSE) between the output of the adaptive filter and noisy received signal is the factor on which most of the channel estimation techniques have been targeted at [2]. Some estimation method needs accurate knowledge of second order channel statistics, such as Wiener filtering based iterative CE. Intense computational complexity and information of channel correlations are required in these schemes [3]. The simplest and most popular CE algorithm, adaptive in nature is the least mean squares (LMS) algorithm which has low computational complexity, memory load, and simplicity of practical implementation [4]. Moreover, its performance and fast convergence speed are inversely related through a single parameter, step size. For large values of step size, the convergence of the LMS filter coefficients is very fast, but the steady state MSE is large and vice versa. In order to ensure the algorithm to be convergent, the range of step size is specified but the choice of optimal step size has not been properly addressed. Therefore, the existing LMS CE algorithm is not possible to obtain fast convergence and small steady state MSE at the same time. One of the important concerns in all practical realistic situations is to develop algorithms which give fast convergence of the filter coefficients and good MSE performance [5].

In order to increase MSE performance as well as fast convergence, normalized LMS (NLMS) CE algorithm is proposed which takes into account the variation in the signal level at the filter input and selects a normalized step size parameter. A drawback of NLMS algorithms is a higher computational complexity and misadjustment i.e., the mismatch between the true and estimated coefficients [6]. When a constant scalar step size is used in the LMS/NLMS algorithm, there is a trade off between the steady state error and the convergence speed, which prevents a fast convergence when the step size is chosen to be small for small output estimation error.

In order to deal with this problem, one important idea is to use varying step size during adaptation. Variable step size (VSS) methods are commonly sought after to provide steady state MSE performance. This method uses larger step size at the start of the iteration to speed up the convergence rate of the algorithm, and smaller step size when the algorithm is convergent [7]. Several VSS-LMS type CE techniques have been proposed in the literature [8], [9], [10]. But these algorithms are not adaptive to track the optimum step size parameter in a nonstationary environment. The existing VSS-LMS CE algorithms [11] cannot provide the minimum MSE in the tracking problem, since they cannot acquire and track the optimum step size. They may even cause worse steady state results, when the algorithm parameters are not appropriately adjusted. In [12], an adaptive time-varying step size LMS is proposed where the step size is adjusted using the energy of the instantaneous error. But due to the presence of the estimation error and measurement noise, the step size update is not an accurate reflection of the state of adaptation before or after convergence. This degrades the significant performance of these adaptive approaches. Furthermore, in [13], [14] proposed a time-varying step size LMS method that gives improved performance compared with standard LMS and NLMS algorithm. But when the channel is fast time-varying then this algorithm cannot accurately measure the autocorrelation between estimation error to control step size update. Therefore, the performance is reduced significantly.

To combat the channel dynamics, the recursive least squares (RLS) based CE algorithm is frequently used for rapid convergence and improved MSE performance [15]. But it requires optimum forgetting factor such that the estimator error is minimized. Although a lot of modified CE algorithms have been studied on employing adaptive forgetting factor and parallel forgetting factor, the CE performance is severely graded in highly dynamic fading channel even when the forgetting factor is well optimized [16]. However, this scheme also requires high computational complexity that is the major obstacle for practical base station (BS) as well as tiny mobile terminal implementation. Therefore, an efficient CE algorithm superior to the existing methods is necessary which provides both rapid convergence to the true channel coefficient and smallest steady state MSE.

1.3 Objective

The objective of the work is to estimate the LTE channel in uplink direction. SCFDMA signal is used in uplink. The estimation is carried out under the following set up conditions.

- An SCFDMA system is modelled for LTE uplink.
- Signal that reaches the receiver is subjected to multipath interference and Doppler effect.
- Estimation of the channel is accomplished by using an adaptive algorithm . It is also compared with existing algorithms like LMS, NLMS, VSS and RLS.

1.4 Thesis Outline

Chapter 1 describes the evolution of wireless cellular network. A brief overview of 1G, 2G, 3G and 4G cellular network is presented. It is followed by literature review.

Chapter 2 comprises OFDMA, SCFDMA signal description and different subcarrier mapping schemes, PAPR analysis of SCFDMA and OFDMA signal.

Chapter 3 gives a basic awareness of Adaptive Filters and its different type like LMS, NMS, RLS and Variable Step Size Filters. An LMS based Adaptive channel estimation is described. Existing algorithms are compared with the proposed one in terms of BER and MSE in fast and slow fading scenario.

Chapter 4 illustrates inference of the thesis along with a discussion on the scope of the work for future.

CHAPTER-2

Single Carrier Frequency Division Multiple Access

2.1 Single Carrier Frequency Division Multiple Access

The significant expansion seen in mobile and cellular technologies over the last two decades is a direct result of the increasing demand for high-data-rate transmissions over bandwidth and power limited wireless channels. This requirement for high data rates results in significant inter symbol interference (ISI) for single carrier systems, and therefore requires the use of robust coding and powerful signal processing techniques in order to overcome the time and frequency selective natures of the propagation channel. In recent years orthogonal frequency division multiplexing (OFDM) has been proposed as an efficient high data rate solution for wireless applications. Particular examples include the physical layer of high-performance wireless local area networks (WLANs), such as the 802.11a/g/n, DVB-T/H, and 802.16 WiMAX standards. This trend has occurred since OFDM offers excellent performance in highly dispersive channels with low terminal complexity. The Third Generation Partnership Project (3GPP) Long Term Evolution (LTE) radio access standard is based on shared channel access providing peak data rates of 75 Mb/s on the uplink and 300 Mb/s on the downlink. A working assumption in the LTE standard is the use of orthogonal frequency-division multiple access (OFDMA) on the downlink. This supports different carrier bandwidths (1.25–20 MHz) in both frequency-division duplex (FDD) and time-division duplex (TDD) modes [2]. OFDMA is an OFDM-based multiple access scheme [1] that provides each user with a unique fraction of the system bandwidth. OFDMA is highly suitable for broadband wireless access networks (particularly the downlink) since it combines scalability, multipath robustness, and multiple-input multiple-output (MIMO) compatibility [1]. OFDMA is sensitive to frequency offset and phase noise, and thus requires accurate frequency and phase synchronization. In addition, OFDMA is characterized by a high transmit PAPR, and for a given peak-power-limited amplifier this results in a lower mean transmit level. For these reasons, OFDMA is not well suited to the uplink transmission. Single carrier FDMA (SC-FDMA), also known as discrete Fourier transform (DFT) precoded OFDMA, has been proposed in the LTE standard for the uplink. PAPR reduction is motivated by a desire to increase the mean transmit power, improve the power amplifier efficiency, increase the data rate, and reduce the bit error rate (BER). This comes at the expense of cost, complexity, and efficiency.

OFDM has become a most favored technique for broadband wireless system due to susceptibility to signal dispersion under multipath conditions. OFDM can also be viewed as a multi-carrier narrowband system where the whole system bandwidth is split into multiple smaller subcarriers with simultaneous transmission. Simultaneous data transmission and reception over these subcarriers are handled almost independently. Each subcarrier is usually narrow enough that multipath channel response is flat over the individual subcarrier frequency range, i.e. frequency non-selective. Another way to look at is that an OFDM symbol time is much larger than the typical channel dispersion. Hence OFDM is inherently susceptible to channel dispersion due to multipath propagation. One major difference between an OFDM and the TDMA or CDMA techniques is important to note. In traditional systems the symbol detection is on the samples at either symbol or chip rate, and it cares about the carrier-to-interference level only at the sampling points. But, OFDM symbol detection requires that the entire symbol duration be free of interference from its previous symbols, inter-symbol interference. Even though OFDM symbol duration is much larger than channel dispersion, even a small amount of channel dispersion causes some spilling of each OFDM symbol to the next symbol, thus it causes some

ISI. However this ISI spill-over is limited to only the initial part of the neighboring symbol. Hence this ISI spill-over at the beginning of each symbol can easily be removed by adding a cyclic prefix to each transmit symbol. Cyclic prefix is the process of extending each symbol by duplicating a portion of the signal at the symbol ends, which is thrown away at the receiver. The amount of symbol extension, i.e. length of cyclic prefixes, is a system design parameter, and it is based on the expected signal dispersion in the environment of system operation.

Single Carrier Frequency Division Multiple Access (SC-FDMA) is a promising technique for high data rate uplink communication and has been adopted by 3GPP for its next generation cellular system, called Long-Term Evolution (LTE). SC-FDMA is a modified form of OFDM with similar throughput performance and complexity. This is often viewed as DFT-coded OFDM where time-domain data symbols are transformed to frequency-domain by a discrete Fourier transform (DFT) before going through the standard OFDM modulation. Thus, SC-FDMA inherits all the advantages of OFDM over other well-known techniques such as TDMA and CDMA. The major problem in extending GSM TDMA and wideband CDMA to broadband systems is the increase in complexity with the multipath signal reception. The main advantage of OFDM, as is for SC-FDMA, is its robustness against multipath signal propagation, which makes it suitable for broadband systems. SC-FDMA brings additional benefit of low peak-to-average power ratio (PAPR) compared to OFDM making it suitable for uplink transmission by user-terminals.

LTE is a next generation mobile system from the 3GPP with a focus on wireless broadband. LTE is based on Orthogonal Frequency Division Multiplexing (OFDM) with cyclic prefix (CP) in the downlink, and on Single-Carrier Frequency Division Multiple Access (SC-FDMA) with cyclic prefix in the uplink.

2.2 Single Carrier Modulation

Based on SC-FDMA's structure, the reasons for some of its names, such as DFT-precoded OFDM or DFT-spread OFDM, are clear. But for the use of 'Single Carrier' in its name, SCFDMA, is not as obvious and is often the reason why is not explained. Unlike the standard OFDM where the each data symbol is carried by the individual subcarriers, the SC-FDMA transmitter carries data symbols over a group of subcarriers transmitted simultaneously. In other words, the group of subcarriers that carry each data symbol can be viewed as one frequency band carrying data sequentially in a standard FDMA.

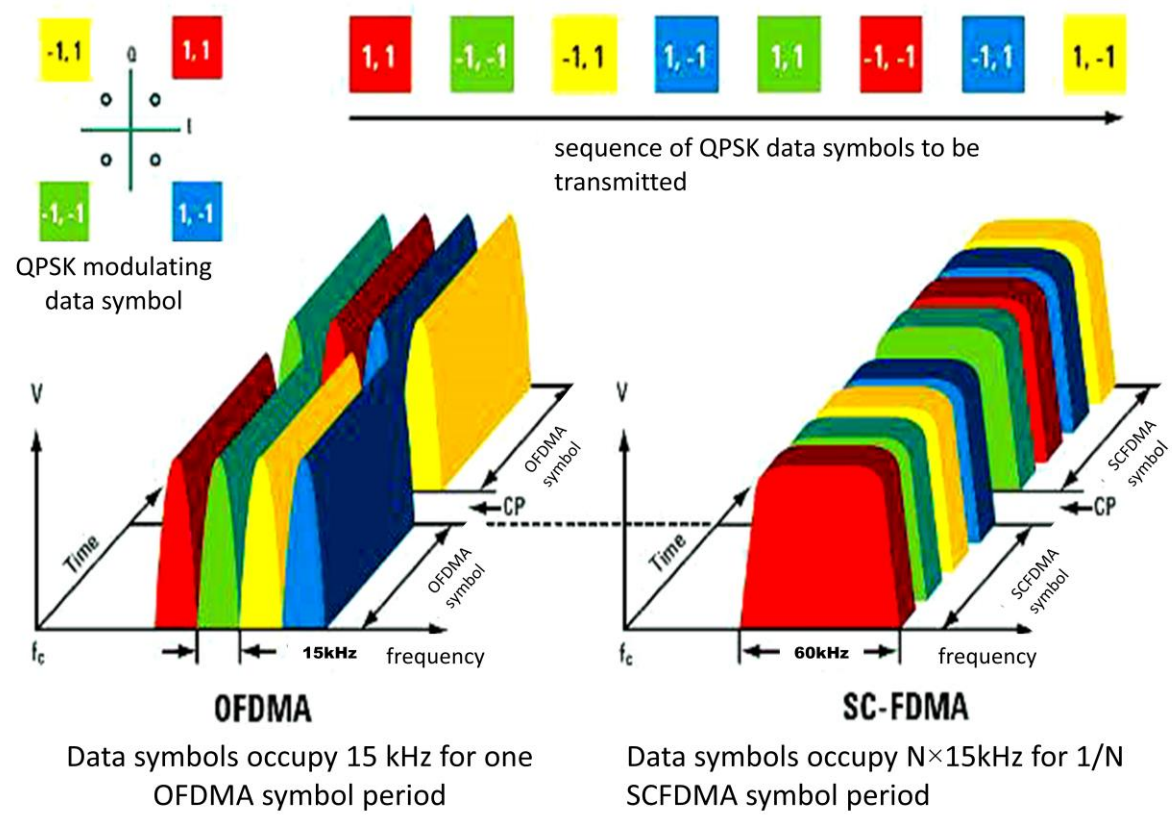

Fig.2. 1 : Comparison of SCFDMA and OFDMA symbol

Fig 2.1: shows how a series of QPSK symbols are mapped into time and frequency by the two different modulation schemes. OFDMA is simply an elaboration of OFDM used by LTE and other systems that increases system flexibility by multiplexing multiple users onto the same subcarriers. This can benefit the efficient trunking of many low-rate users onto a shared channel as well as enable per-user frequency hopping to mitigate the effects of narrowband fading. For clarity, the example here uses only four (N) subcarriers over two symbol periods with the payload data represented by QPSK modulation. On the left side of Fig 2.1, N adjacent 15 kHz subcarriers — already positioned at the desired place. In this simple four-subcarrier example, four symbols are taken in parallel. These are QPSK data symbols so only the phase of each subcarrier is modulated and the subcarrier power remains constant between symbols. After one OFDMA symbol period has elapsed, the CP is inserted and the next four symbols are transmitted in parallel. For visual clarity, the CP is shown as a gap; however, it is actually filled with a copy of the end of the next symbol, meaning the transmission power is continuous but has a phase discontinuity at the symbol boundary. To create the transmitted signal, an inverse FFT is performed on each subcarrier to create N time-domain signals that are vector summed to create the final time-domain waveform used for transmission. SC-FDMA signal generation begins with a special precoding process but then continues as with OFDMA. Before outlining the generation process it is helpful to first describe the en result as shown on the right side of Fig 2.1. The most obvious difference between the two schemes is that OFDMA transmits the four QPSK data symbols in parallel, one per subcarrier, while SC-FDMA transmits the four QPSK data symbols in series at four times the rate, with each data symbol occupying N x 15 kHz bandwidth. Visually, the OFDMA signal is clearly multi-carrier and the SC-FDMA signal looks more like single-carrier, which explains the “SC” in its name. Note that OFDMA and SC-FDMA symbol lengths are the same at 66.7 μs; however, the SC-FDMA symbol contains N “sub-symbols” that represent the modulating data.

It is the parallel transmission of multiple symbols that creates the undesirable high PAR of OFDMA. By transmitting the N data symbols in series at N times the rate, the SC-FDMA occupied bandwidth is the same as multi-carrier OFDMA but crucially, the PAPR is the same as that used for the original data symbols. This should make heuristic sense without delving into the mathematics: Adding together many narrowband QPSK waveforms in OFDMA will always create higher peaks than would be seen in the wider-bandwidth single-carrier QPSK waveform of SC-FDMA. As the number of subcarriers N increases, the PAR of OFDMA with random modulating data approaches Gaussian noise statistics but, regardless of the value of N, the SC-FDMA PAR remains the same as that used for the original data symbols in the channel bandwidth .

2.3 SCFDMA Signal Processing

An SCFDMA transmitter is shown in Fig.2.2, which sends one block of data to receiver. The input of the transmitter and output of receiver are complex modulation symbols. The N-point discrete Fourier transform (DFT) produces N frequency domain symbols, that modulate N out of M orthogonal subcarriers spread over a bandwidth,

$$W_{channel=} M f_0 [\text{Hz}]$$

Where f_0 Hz is subcarrier spacing. The channel transmission rate is

$$R_{channel=} \frac{M}{N} R_{source} [\text{symbols/second}]$$

If Q denotes bandwidth spreading factor, i.e. ,

$$Q = \frac{R_{channel}}{R_{source}} = \frac{M}{N}$$

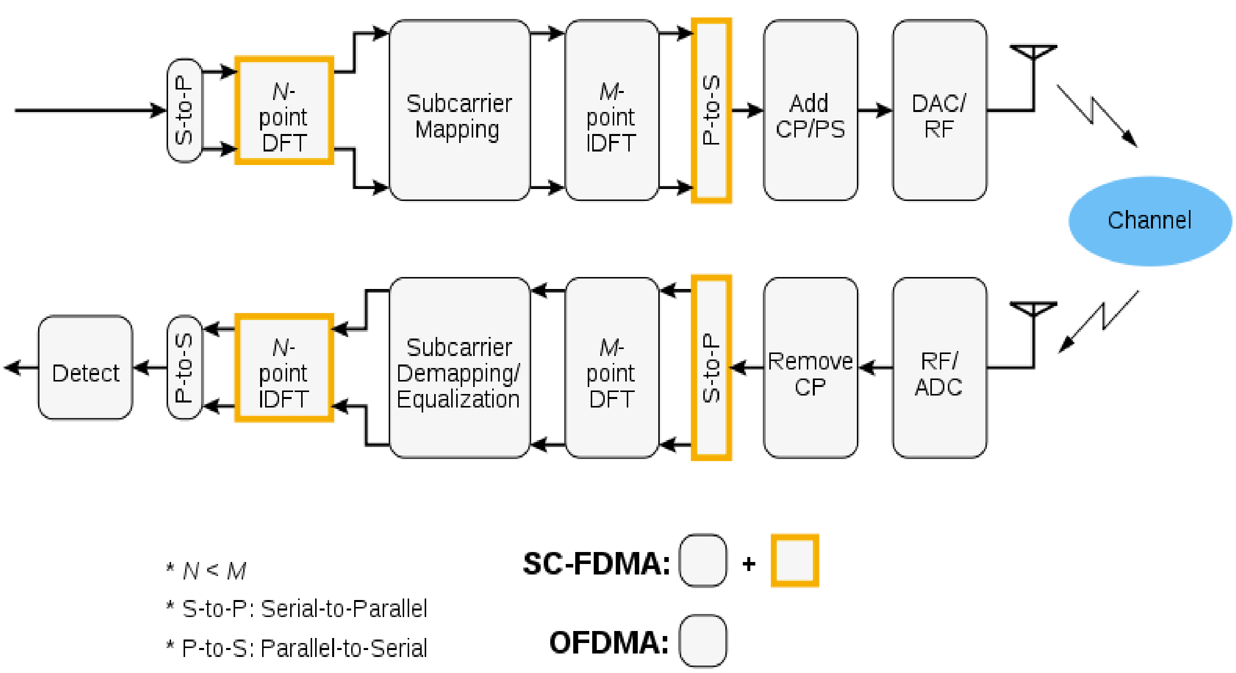

Fig. 2.2: Transmitter and receiver structure of SCFDMA and OFDMA systems

A Subcarrier mapping scheme follows DFT, which assign frequency domain modulation symbols to subcarriers. The mapping process is sometimes referred to as scheduling. The IDFT creates a time domain representation of M subcarrier symbols .The parallel to serial converter places the time domain symbols in a sequence suitable for modulating a radio frequency carrier.

The transmitter performs two other signal processing operation prior to transmission .First one, is insertion of CP inorder to prevent IBI due to multipath propagation. CP, is a copy of last part of the block, it provides a guard time between the blocks .Length of CP is ensured longer than maximum delay spread to prevent IBI. The removal of channel distortion comprise of dividing the DFT of received signal by DFT of channel impulse response. Raised Cosine filter is one of the commonly used pulse shaping filter [1].

At the receiver, the DFT transforms, received signal to frequency domain to recover M subcarriers N frequency domain samples of each source signal is isolated by demapping. ISI caused by single carrier modulation is cancelled by frequency domain equalization. The equalized symbols are converted to time domain by IDFT M modulation symbols are generated by a detector.

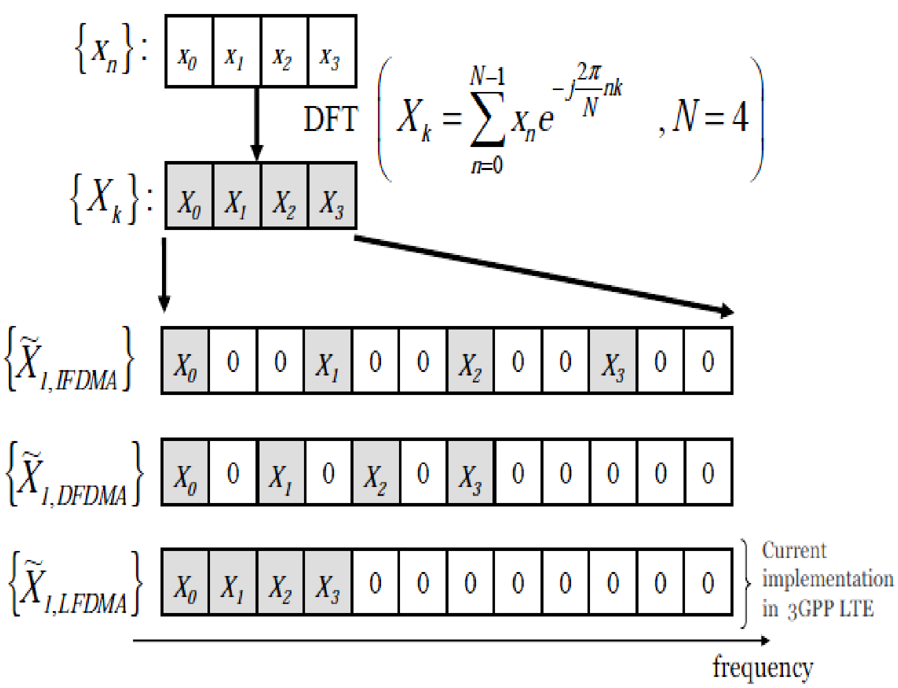

Fig. 2.3: An example of different subcarrier mapping scheme for N=4, Q=3 and M=12

DFT output of the data symbols is mapped to a subset of subcarriers, a process called subcarrier mapping. The subcarrier mapping assigns DFT output complex values as the amplitudes of some of the selected subcarriers. Subcarrier mapping can be classified into two types: localized mapping and distributed mapping. In localized mapping, the DFT outputs are mapped to a subset of consecutive sub-carriers thereby confining them to only a fraction of the system bandwidth. In distributed mapping, the DFT outputs of the input data are assigned to subcarriers over the entire bandwidth non-continuously, resulting in zero amplitude for the remaining subcarriers. A special case of distributed SC-FDMA is called interleaved SC-FDMA, where the occupied subcarriers are equally spaced over the entire bandwidth. Fig 2.4 is a general picture of localized and distributed mapping.

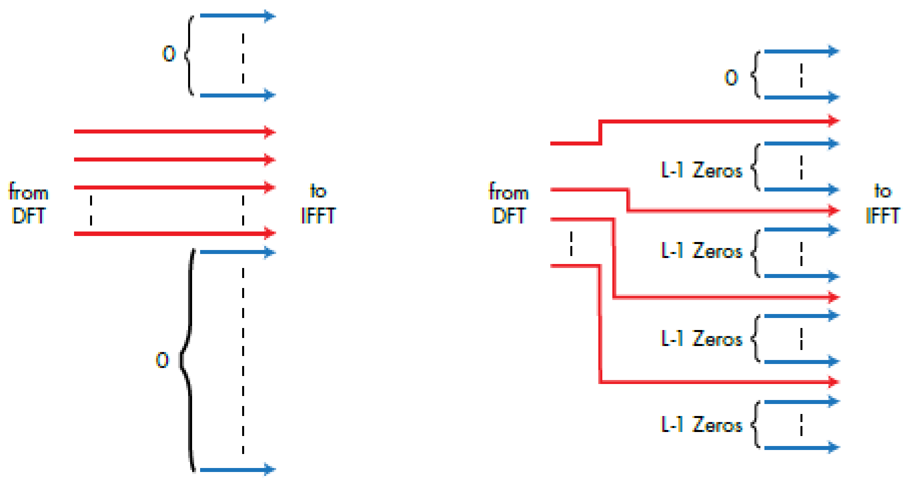

Fig. 2.4: Localized Mapping Verses Distributed Mapping

An example of subcarrier mapping is shown in Fig 2.3. This example assumes three users sharing 12 subcarriers. Each user has a block of four data symbols to transmit at a time. The DFT output of the data block has four complex frequency domain samples, which are mapped over 12 subcarriers using different mapping schemes. SC-FDMA inherently offers frequency diversity gain over the standard OFDM, as all information data is spread over multiple subcarriers by the DFT mapper. However, the distributed SC-FDMA is more robust with respect to frequency selective fading and offers additional frequency diversity gain, since the information is spread across the entire system bandwidth. Localized SC-FDMA in combination with channel-dependant scheduling can potentially offer multi-user diversity in frequency selective channel conditions.

Parameters	Assumptions
Modulation	BPSK
FFT size	16
Subcarrier mapping	IFDMA,LFDMA
IFFT size	64
Cyclic prefix	20
Channel	3GPP Pedestrian A
Equalization	MMSE

Table 2.1: Simulation parameters for SCFDMA .

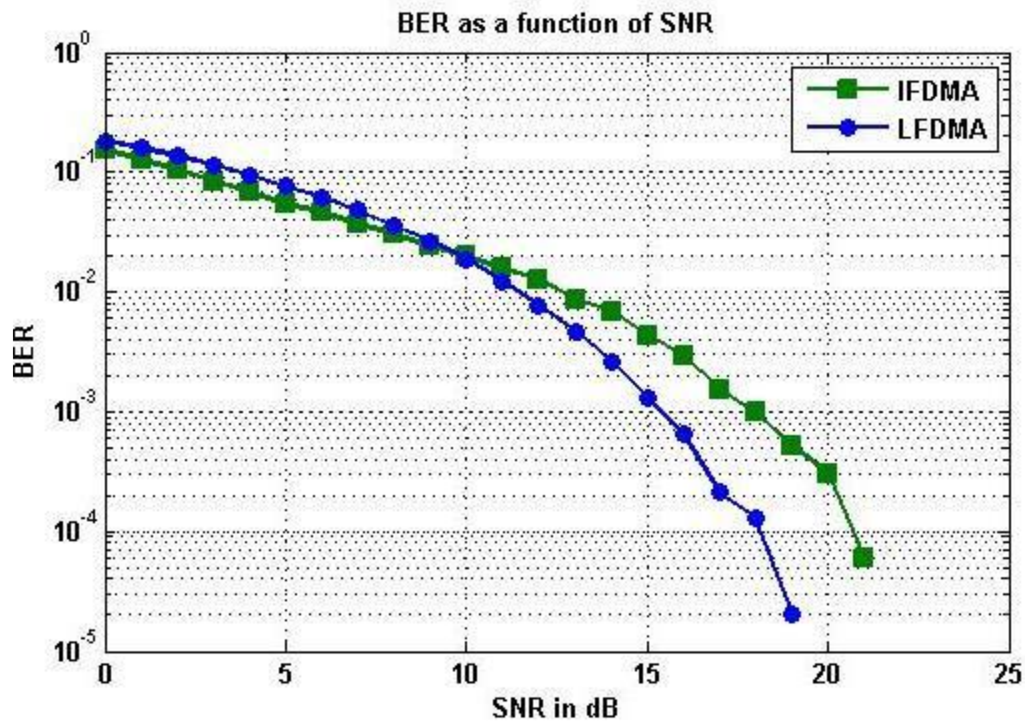

Fig 2.5: BER performance of SCFDMA in different subcarrier mapping schemes, in Pedestrial A channel using MMSE

It is observed from the simulated result that, an LFDMA signal gives better BER performance than an IFDMA signal in a four path Pedestrian A multipath channel for MMSE estimation.

2.4 PAPR Analysis

SC-FDMA offers similar performance and complexity as OFDM. However, the main advantage of SC-FDMA is the low PAPR (peak-average-power ratio) of the transmit signal. PAPR is defined as the ratio of the peak power to average power of the transmit signal. As PAPR is a major concern at the user terminals, low PAPR makes the SC-FDMA the preferred technology for the uplink transmission. PAPR relates to the power amplifier efficiency at the transmitter, and the maximum power efficiency is achieved when the amplifier operates at the saturation point. Lower PAPR allows operation of the power amplifier close to saturation resulting in higher efficiency. With higher PAPR signal, the power amplifier operating point has to be backed off to lower the signal distortion, and thereby lowering amplifier efficiency. As SC-FDMA modulated signal can be viewed as a single carrier signal, a pulse shaping filter can be applied to transmit signal to further improve PAPR. Interleaved SC-FDMA is a preferred modulation technique for lower PAPR. Pulse shape filtering of SC-FDMA in fact degrades the PAPR level of interleaved SC-FDMA whereas it shows no effect with localized SC-FDMA [24].

2.4.1 PAPR OF SCFDMA AND OFDMA SIGNAL

The PAPR of SCFDMA and OFDMA signals have been compared in Fig 2.6.It is observed that an SCFDMA signal gives low PAPR than an OFDMA signal, which makes SCFDMA suitable for use in uplink.

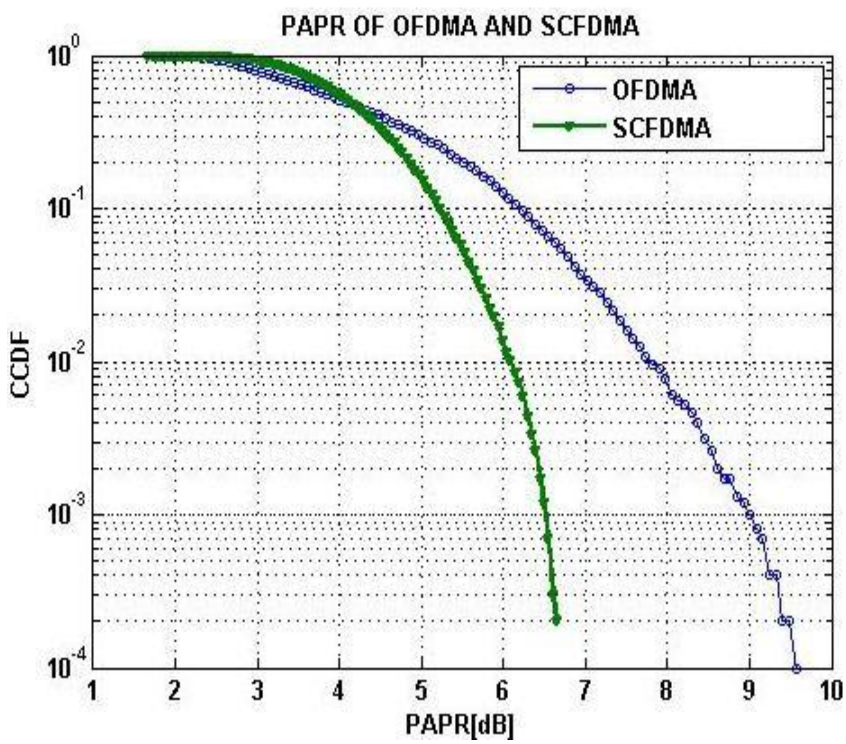

Fig 2.6: PAPR of OFDMA and SCFDMA signal

2.5 Comparison of Different Subcarrier Mapping Methods

The different versions of SC-FDMA with different subcarrier allocation methods vary in their properties such as: power efficiency, performance in frequency selective channels, and system throughput. The time domain samples in IFDMA consist of the actual input symbols only, whereas in LFDMA they also include the complex weighted sum of all the input symbols in the block. Therefore, the transmitted waveforms in LFDMA have more amplitude fluctuations than in IFDMA. As a result, LFDMA has much higher PAPR compared to IFDMA. In frequency selective channels, where the channel gain is not constant over the entire bandwidth, LFDMA has worse performance than IFDMA. Since in IFDMA the data is distributed throughout the whole bandwidth, it is not affected by the channel gain. The error performance will be the same for all users. But in LFDMA, each user utilizes a block of subcarriers located at a particular area of the total bandwidth, so the bit error rate will vary from one user to another depending on where the block of the subcarriers is located. To improve the performance of LFDMA schemes in frequency selective channels, channel-dependent subcarrier allocation (CDS) instead of static (round robin) scheduling can be used. Channel dependent scheduling is a form of subcarrier mapping, where the transmission of each terminal is mapped to a set of subcarriers with favorable transmission characteristics. Myung and Goodman in [14], showed that when CDS is applied, there is a significant improvement in the average throughput for both IFDMA and LFDMA. But compared to IFDMA, the capacity gain from CDS is much higher in LFDMA. Therefore, as discussed in [14], when power efficiency is considered, IFDMA is more desirable than LFDMA, but in terms of system throughput, LFDMA outperforms IFDMA when CDS is applied.

Chapter 3

LMS Based Adaptive Channel Estimation

3.1.1 Adaptive Filters

An adaptive filter may be understood as a self-modifying digital filter that adjusts its coefficients in order to minimize an error function. This error function, also referred to as the cost function, is a distance measurement between the *reference* or *desired* signal and the output of the adaptive filter. Application area include noise and echo canceling, channel equalization, signal prediction, adaptive arrays as well as many others. In order to compare the wide variety of algorithms available in the literature of adaptive filtering, the following aspects must be taken into account

• **Filter structure:** The input–output relationship of the adaptive filter depends on its transfer function implementation. Due to its simplicity and efficacy, the most widely employed adaptive structure is by far the transversal filter (or tapped delay line) associated to standard finite-duration impulse response (FIR) filters.

Other structures comprise FIR lattice and infinite-duration impulse response (IIR) filters. This aspect greatly influences the computational complexity of a given adaptive algorithm and the overall speed of the adaptation process.

• **Rate of convergence, misadjustment, and tracking**: In a noiseless (no measurement or modeling noise) situation, the coefficients of an adaptive filter can be made to converge fast or slowly to he optimum solution. In practice, the adaptive coefficients do not reach the optimum values but stay close to the optimum. Misadjustment is a measure of excess error associated to how close these coefficients (the estimated and the optimum) are to each other in steady-state. It can be taken as a general rule that, for a given algorithm, a faster convergence yields a higher misadjustment. In non-stationary environments, the algorithm convergence speed is also associated to the tracking ability of the adaptive filter.

• **Computational aspects:** Due to the desired real-time characteristic, the adaptive filter performance must take into account practical levels of computational complexity and limited-precision representation of associated signals and coefficients. The effort in obtaining fast versions of more complex algorithms results from the desire of reducing the computational requirements to a minimal number of operations, as well as reducing the size of memory necessary to run these algorithms in practical applications. On the other hand, a limited-precision environment generates quantization errors which drive the attention of designers to numerical stability, numerical accuracy, and convergence robustness of the algorithm.

The basic configuration of an adaptive filter, operating in the discrete-time domain k, is illustrated in Fig 3.1. In such a scheme, the input signal is denoted by $x(k)$, the reference signal $d(k)$ represents the desired output signal (that usually includes some noise component), $y(k)$ is the output of the adaptive filter, and the error signal is defined as

$$e(k) = d(k) - y(k). \tag{3.1}$$

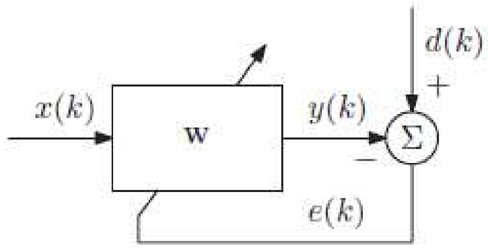

Fig 3.1 : Basic block diagram of an adaptive filter.

The error signal is used by the adaptation algorithm to update the adaptive filter coefficient vector $\mathbf{w}(k)$ according to some performance criterion. In general, the whole adaptation process aims at minimizing some metric of the error signal, forcing the adaptive filter output signal to approximate the reference signal in a statistical sense. In practical communications systems, a transmitted signal can be heavily distorted by the transmission channel. One may attempt to recover the original signal by employing an adaptive filter in the channel equalization configuration, as depicted in Fig 2.4. In such a framework, a training sequence $s(k)$ known by the receiver is sent via a given channel generating a distorted signal. The same sequence $s(k)$, after a proper time shift to compensate for transmission delays, is used as a reference signal in the receiver for the adaptive filter, whose input is the distorted signal. When the error function approximates zero, the output signal $y(k)$ resembles the transmitted signal $s(k)$, indicating that the adaptive filter is compensating for the channel distortions. After this training process, the desired information can be sent through the channel, which is properly equalized by the adaptive filter.

Adaptation of the filter coefficients follows a minimization procedure of a particular objective or cost function. This function is commonly defined as a norm of the error signal $e(k)$.

3.1.2 The mean-square error

The MSE is defined as

$$\xi(\mathrm{k}) = \mathrm{E}[e^2(\mathrm{k})] = \mathrm{E}[\ |\mathrm{d(k)} - \mathrm{y(k)}|\ ^2] \tag{3.2}$$

Writing the output signal *y*(*k*) $= w^T \mathrm{x(k)}$,

$$\xi(\mathrm{k}) = \mathrm{E}[\ \big| \ \mathrm{d(k)} - w^T \mathrm{x(k)} \ \big| \ ^2] \tag{3.3}$$

$$= E[d^2(\mathrm{k})] - 2w^T \mathrm{E}[\mathrm{d(k)x(k)}] + w^T \mathrm{E}[\mathrm{x(k)}x^T(\mathrm{k})]\mathrm{w} \tag{3.4}$$

$$= E[d^2(\mathrm{k})] - 2w^T \mathrm{p} + w^T \mathrm{Rw} \tag{3.5}$$

where **R** and **p** are the input-signal correlation matrix and the cross-correlation vector between the reference signal and the input signal, respectively, and are defined as

$$\mathrm{R} = \mathrm{E}[\mathrm{x(k)}x^T(\mathrm{k})] \tag{3.6}$$

$$\mathrm{p} = \mathrm{E}[\mathrm{d(k)}x^T(\mathrm{k})] \tag{3.7}$$

Note, from the above equations, that **R** and **p** are not represented as a function of the iteration *k* or not time-varying, due to the assumed stationarity of the input and reference signals. From Equation (3.5), the gradient vector of the MSE function with respect to the adaptive filter coefficient vector is given by

$$\Delta_{\mathrm{w}} \xi(\mathrm{k}) = -2\mathrm{p} + 2\mathrm{Rw} \tag{3.8}$$

The so-called Wiener solution w_0, that minimizes the MSE cost function, is obtained by equating the gradient vector in Equation (2.8) to zero. Assuming that **R** is non-singular, we have,

$$w_0 = R^{-1}\,\mathrm{p} \tag{3.9}$$

3.2 .1 LMS Algorithm

Determining the Wiener solution for the MSE problem requires inversion of matrix **R**, which makes Equation (3.9) hard to implement in real time. One can then estimate the Wiener solution, in a computationally efficient manner, iteratively adjusting the coefficient vector **w** at each time instant *k*, in such a manner that the resulting sequence **w**(*k*) converges to the desired w_0 solution, possibly in a sufficiently small number of iterations. The so-called steepest-descent scheme searches for the minimum of a given function following the opposite direction of the associated gradient vector. A factor μ/2, where μ is the so-called convergence factor, adjusts the step size between consecutive coefficient vector estimates, yielding the following updating procedure:

$$e(m) = S^T(\mathrm{m})\mathrm{w}(\mathrm{m}) + \mathrm{z}(\mathrm{m}) - S^T(\mathrm{m})\mathrm{h}(\mathrm{m})$$
$$w(m+1) = w(m) + \eta S(m)e(m) \tag{3.10}$$

where η is step size, S(m) is the transmitted diagonal matrix at sampling time m, h(m) is the adaptive filter coefficient, and e(m) is the estimation error. The filter coefficients are updated using an estimate of the cost function gradient, [ηS(m)e(m)]. In all practical applications, the signals involved might be corrupted by noise. When the noise is present in the received sequence, interference will also in the coefficients adaption process through the term [ηS(m)e(m)]. As a result, where the distribution of the noise is highly impulsive, the LMS scheme might have low convergence and lower steady state MSE performance. The step size parameter, η determines the convergence rate of the algorithm and higher value provides faster convergence. However, if η exceeds certain bound then the algorithm will diverge. As the bound on η is not known a priori and is dependent on the various statistics. In practice, a somewhat conservative scalar value of η is used. Also a higher value of η results in higher variations in the tap weight vector estimate after the initial convergence phase. Such variations result in increased distortion in the combiner output which in turn results in an increased MSE and BER [9], [13].

3.2.2 Normalized LMS (NLMS) Algorithm

The main problem of the LMS CE algorithm is that it is sensitive to the scaling of its input signals. This makes it very hard to choose η that guarantees stability of the algorithm. The NLMS is a variant of the LMS algorithm that solves this problem by normalizing with the power of the input signal. The NLMS algorithm can be summarized as [29]:

$$e(m) = S^T(\mathrm{m})\mathrm{w}(\mathrm{m}) + \mathrm{z}(\mathrm{m}) - S^T(\mathrm{m})\mathrm{h}(\mathrm{m})$$
$$h(m+1) = h(m) + \eta e(m)[S^T(\mathrm{m})\mathrm{S}(\mathrm{m})]^{-1}\mathrm{S}(\mathrm{m}) \qquad (3.11)$$

when a constant scalar step size is employed in the LMS/NLMS algorithm, there is a trade off among the steady state error-convergence towards the true channel coefficients, which avoids a fast convergence when the step size is preferred to be small for small output estimation error. In order to guarantee the algorithm to be convergent, the range of step size is specified but the choice of optimal learning step size has not been appropriately addressed. In order to deal with these troubles, one key idea is to exploit varying step size during adaptation.

3.2.3 Variable Step Size Algorithm

The VSS-LMS algorithm involves one additional step size update equation compared with the standard LMS algorithm. The VSS algorithm is [30], [19]

$$\eta(m+1) = \alpha\eta(m) + \gamma\ \ p^2(m$$
$$p(m) = \beta p(m) + (1-\beta)\ e^{\mathrm{T}}(m)e(m-1) \qquad (3.12)$$

where $0 < \alpha < 1$, $0 < \beta < 1$, and $\gamma > 0$ When the channel is fast time-varying then algorithm cannot accurately mea- sure the autocorrelation between estimation error to control step size update. So, this CE algorithm cannot provide the minimum MSE in the tracking problem, since it cannot acquire and track the optimum step size. It may even cause worse steady state results, when the algorithm parameters are not appropriately adjusted. In addition, control parameters α and β need to be adjusted for a better performance. As can be seen here, a general characteristic of these VSS CE methods is that predetermined control parameters are necessary to improve the performance. Though, in most of them, rules to choose control parameters are not specified. Those parameters are always selected from extensive simulations, or from experience. It is clear that the choice of parameters would significantly influence the performance of these schemes.

3.2.4 RLS Algorithm

To combat the channel dynamics, the RLS based CE algorithm is frequently used for rapid convergence and improved MSE performance [9]. The standard RLS algorithm is

$$\begin{aligned} e(m) &= S^T(\mathrm{m})\mathrm{w}(\mathrm{m}) + \mathrm{z}(\mathrm{m}) - S^T(\mathrm{m})\mathrm{h}(\mathrm{m}) \\ R(m) &= B(m-1)S(m)[\lambda + S^T(\mathrm{m})\mathrm{B}(\mathrm{m}-1)\mathrm{S}(\mathrm{m})]^{-1} \\ B(m) &= \lambda^{-1}B(m-1) - \lambda^{-1}R(m)S^T(\mathrm{m})\mathrm{R}(\mathrm{m}-1) \\ h(m+1) &= h(m) + S(m)e(m)R(m) \end{aligned} \tag{3.13}$$

where λ is the exponential forgetting factor with $0 < \lambda < 1$. The smaller value of λ leads to faster convergence rate as well as larger fluctuations in the weight signal after the initial convergence. On the other hand, too small λ value makes this algorithm unstable. Subsequently, it requires best possible forgetting factor such that the estimator error is de- creased. Although a lot of modified CE algorithm has been studied on employing adaptive forgetting factor and parallel forgetting factor, the CE performance is severely degraded in highly dynamic fading channel even when the forgetting factor is well optimized [23]. However, this scheme also has computational complexity-performance trade off problem that is the major obstacle for practical mobile terminal as well as base station (BS) implementation [30]. Consequently, an efficient CE algorithm better than existing algorithms is required which gives both fast convergence and minimum steady state MSE.

3.3 Adaptive LMS Algorithm

The signal s(m) is transmitted via a time-varying channel w(m), and corrupted by an observation noise z(m) before being detected in a receiver. The block diagram of proposed CE algorithm in LTE SC-FDMA system is illustrated in Fig 3.2 .The signal received at time index m is

$$r(m) = s_1(\mathrm{m}-1)w_1(\mathrm{m}) + \ldots \ldots \quad \ldots + s_l(\mathrm{m}-\mathrm{l})w_l(\mathrm{m}) + \mathrm{z(m)}$$

$$= \sum_{j=1}^{l} s_{\mathrm{j}}(m-j)w_j(\mathrm{m}) + \mathrm{z(m)}$$

$$= S^T(\mathrm{m})\mathrm{w(m)} + \mathrm{z(m)} \qquad (3.14)$$

Where $s_j(\mathrm{m}-\mathrm{j})]$, $\mathrm{j} = 1,2 \ldots .\mathrm{l}$ are transmitted signal vectors at time m, l is the distinct paths from transmitter to the receiver, w(m) is the channel coefficients at time m, and z(m) is the noise with zero mean and variance σ^2

After processing some intermediate steps (synchronization, remove CP, DFT, and demapping), the decision block reconstructs the detected signal to an approximate modulated signal and its phase. The output y(m) of the adaptive filter is expressed as

$$y(m) = d_1(\mathrm{m}-1)h_1(\mathrm{m}) + \ldots \ldots \ldots + d_l(\mathrm{m}-\mathrm{l})h_l(\mathrm{m})$$

$$= \sum_{j=1}^{l} d_{\mathrm{j}}(m-j)h_j(\mathrm{m})$$

$$= D^T(\mathrm{m})\mathrm{h(m)}$$

where $d_j(\mathrm{m}-\mathrm{j})]$, $\mathrm{j} = 1,2 \ldots . \mathrm{l}$, are detected signal vectors at time m,.. $D(m) = diag[d_1(\mathrm{m}-1), d_2(\mathrm{m}-2), \ldots . d_l(\mathrm{m}-\mathrm{l})]$. In this problem formulation, the ideal adaptation procedure would adjust $w_j(\mathrm{m})$ such that $w_j(\mathrm{m}) = h_j(\mathrm{m})$ as m → ∞ . In practice, the adaptive filter can only adjust w(m) such that y(m) closely approximates desired signal over time. Therefore, the instantaneous estimated error signal needed to update the weights of the adaptive filter is

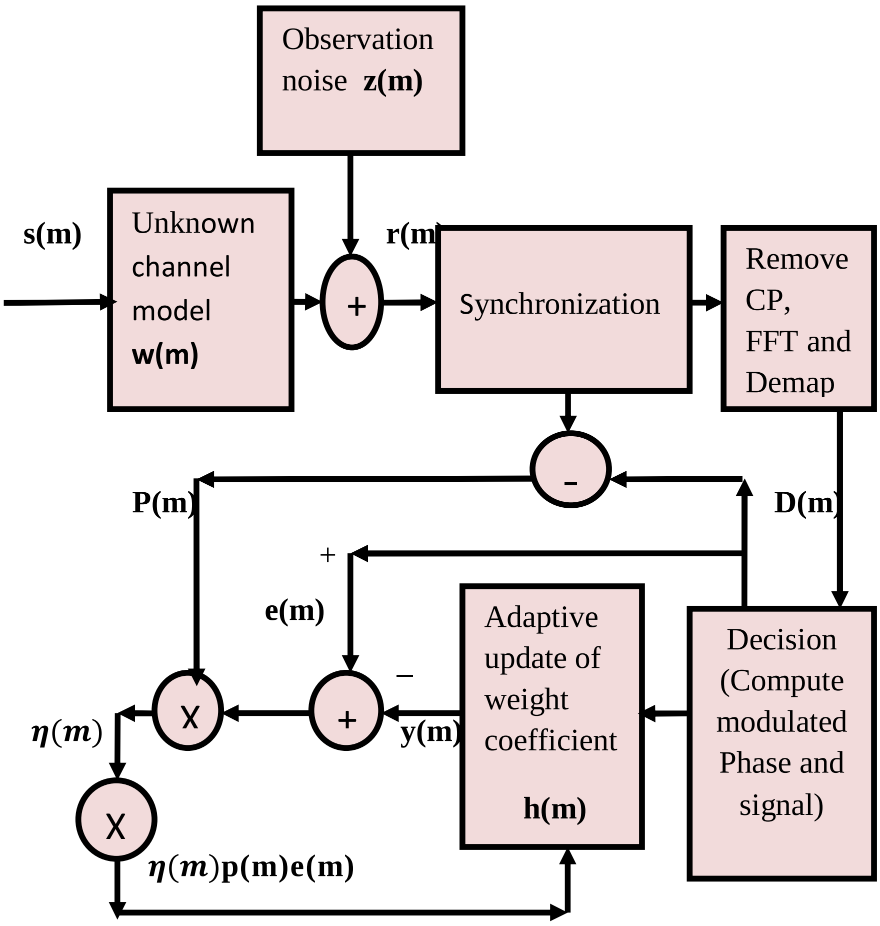

Fig. 3.2: Block diagram of adaptive algorithm for a dynamic system

$$j(m) = p(m)e^T(\mathrm{m})\mathrm{e}(\mathrm{m})$$

$$\mathrm{e}(\mathrm{m}) = \mathrm{r}(\mathrm{m}) - \mathrm{y}(\mathrm{m})$$

$$= r(m) - D^T(\mathrm{m})\mathrm{h}(\mathrm{m}) \tag{3.14}$$

This priori error signal, e(m) is used to minimize the estimator error by adaptive updation of filter weights..

The proposed cost function j(m) for the adaptive filter, minimizes the square distance between the received signal and its estimate. A phase discriminate weighting sequence p(m) is included, so that algorithm is less vulnerable to signals fluctuations subjected to noise as or estimation errors. The weighting sequence p(m) is the distance between the initial modulated carrier phase α and carrier synchronization phase β i.e.,

$$p(m) = \frac{\min \left|\alpha(\mathrm{m}) - \beta(\mathrm{m})\right|}{\pi/M}$$

$$p(m) = p_1(m)p_1(m-1) \ldots \ldots p_1(m-l) \tag{3.15}$$

where M is the alphabet size i.e. M = 2 for BPSK, M = 16 for 16-QAM etc, (π/M) is the normalized factor. So, the proposed cost function j(m) is

$$j(m) = p(m)e^T(\mathrm{m})\mathrm{e}(\mathrm{m})$$

$$= p(m)[r^T(\mathrm{m})\mathrm{r}(\mathrm{m}) - r^T(\mathrm{m})D^T(\mathrm{m})\mathrm{h}(\mathrm{m}) - \mathrm{r}(\mathrm{m})h^T(\mathrm{m})\mathrm{D}(\mathrm{m}) + D^T(\mathrm{m})D(m)h^T(\mathrm{m})\mathrm{h}(\mathrm{m})] \tag{3.16}$$

So as to minimize the cost function in (3.16), the gradient with respect to filter coefficient results,

$$\Delta_{\mathrm{h}}\, \mathrm{j(m)} = p(m)[-2\mathrm{r(m)D(m)} + 2\mathrm{D(m)D}^{T}\mathrm{(m)h(m)}] \qquad (3.17)$$

The steepest descent method is used to adjust adaptive parameters in order to search the quadratic MSE performance function for its minimum. According to this method, a sequence of change is made to the weight vector along the direction of the negative gradient.

Hence, next weight vector, h(m + 1), is the sum of ,present weight vector, h(m) and a change proportional to the negative gradient at the m^{th} iteration, ie.

$$\begin{aligned} h(m+1) &= h(m) - 1/2\eta(m)\Delta_{\mathrm{h}}\, \mathrm{j(m)} \\ &= h(m) + p(m)\eta(m)\mathrm{D(m)}[r\mathrm{(m)} - D^{T}\mathrm{(m)h(m)}] \\ &= h(m) + p(m)\eta(m)\mathrm{D(m)e(m)} \qquad (3.18) \end{aligned}$$

where η(m) is the time-varying step size parameter which is related to the convergence rate . The term [η(m) p(m)D(m)e(m)] is the updating factor. Its observed that coefficients of the adaptive filter are updated using an estimate cost function gradient, priori error e(m), phase discriminate weighting sequence p(m), and time-varying step size parameter η(m).

For obtaining time- varying step size for the proposed LMS algorithm, the gradient in (3.16) with respect to η(m) is calculated as

$$\begin{aligned} \Delta\eta\, j(m) &= p(m)\left[\frac{\partial e^{T}\mathrm{(m)}}{\partial\eta(m)}e(m) + \frac{\partial e(m)}{\partial\eta(m)}e^{T}(m)\right] \\ &= -p(m)[D^{T}\mathrm{(m)c(m)e(m)} \\ &\quad + D^{T}\mathrm{(m)c(m)e(m)}] \\ &= -2p(m)D^{T}\mathrm{(m)c(m)e(m)} \end{aligned}$$

It is assumed that, $\mathrm{c(m)} = \frac{\partial \mathrm{e(m)}}{\partial \eta\mathrm{(m)}}$. By differentiating h(m) with respect to η(m), we obtain,

$$\begin{aligned} c(m+1) &= c(m) + b(m)\frac{\partial e(m)}{\partial\eta(m)} + \eta\mathrm{(m)p(m)D(m)}\frac{\partial \mathrm{e(m)}}{\partial\eta\mathrm{(m)}} \\ &\approx b(m) + \xi c(m) \end{aligned}$$

Where $b(m) = \mathrm{p(m)D(m)e(m)}$,

$\xi = I - \eta p(m) D^T(\mathrm{m})\mathrm{D}(\mathrm{m})$, ξ is taken as a scalar positive constant nearly equal to unity; $\mathrm{c}(\mathrm{m})$ is initial zero vector.

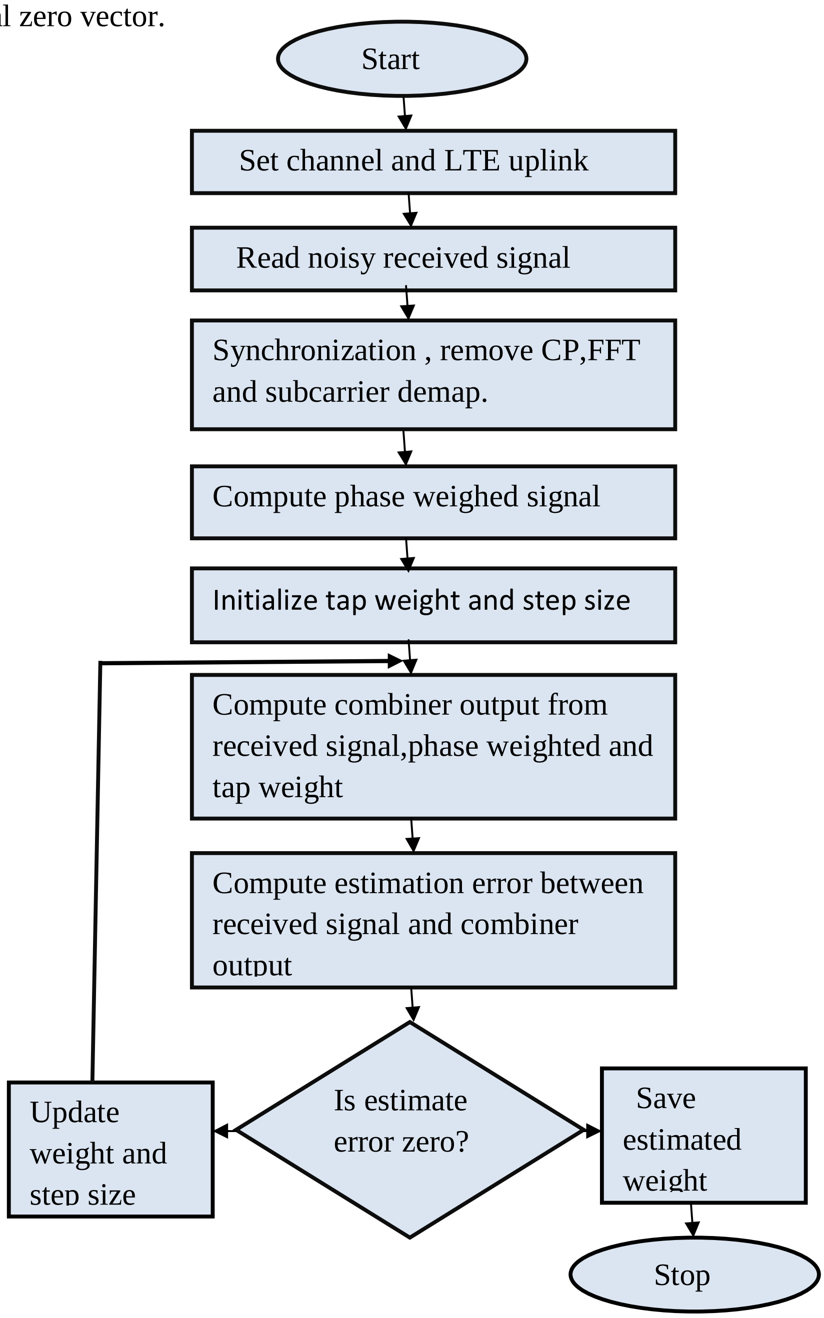

Fig. 3.3: Flowchart of Adaptive LMS for LTE uplink

Updating equation for step size is,

$$\eta(m+1) = \eta(\mathrm{m}) - 1/2\varphi\Delta_{\eta}\ \mathrm{j}(m)$$
$$= \eta(\mathrm{m}) + \varphi(\mathrm{m})D^{T}(\mathrm{m})\mathrm{e}(\mathrm{m})\mathrm{c}(\mathrm{m}) \qquad (3.19)$$

where φ is the learning rate parameter. This time-varying step size is re-selected at each iteration to minimize the sum of the squares of the prior estimation errors up to that recent time point. So, this algorithm is able to sense the convergence rate at which the best possible tap weight coefficients are changing. At the beginning of estimation an initial CIR and step size is given to commence the iteration process. The algorithm is kept on iterative until the channel estimator converges towards the true channel vectors.

3.5 SIMULATION RESULT

The performance of the proposed CE algorithm is compared with the fixed step size LMS algorithm, NLMS algorithm, VSS-LMS algorithm, and RLS algorithm subjected to a Rayleigh fading environment. The simulation parameters are listed in Table 3.1. The BER is a significant performance parameter for quality measurement of recovered data in wireless communication system. The perameters considered for the simulation is enlisted in the Table 3.1. IFDMA signal is used to evaluate the system performance. The effect of the proposed CE in terms of BER performance is compared with existing estimators. It is evident that the proposed CE algorithm outperforms the existing algorithms. The performance degrades with the increased Doppler frequency, ie, when f_d increases from 100 Hz to 1000 Hz.

PARAMETERS	ASSUMPTIONS
Modulation	BPSK
FFT size	16
Subcarrier mapping	IFDMA
IFFT size	64
Cyclic prefix	20
Equalization	Zero force
Doppler frequency	100,1000Hz

Table 3.1 Parameters considered for simulation

3.5.1 Analysis of BER performance in AWGN CHANNEL

The simulation results when an SCFDMA signal is passed through an AWGN channel is shown in Fig 3.4. It is evident that the adaptive algorithm outperforms the existing algorithms. The updating of filter coefficient as well as the step size, using the phase discriminate weighing sequence parameter, as described in Sec. 3.5 have resulted in a superior performance of the proposed algorithm relative to the current algorithms.

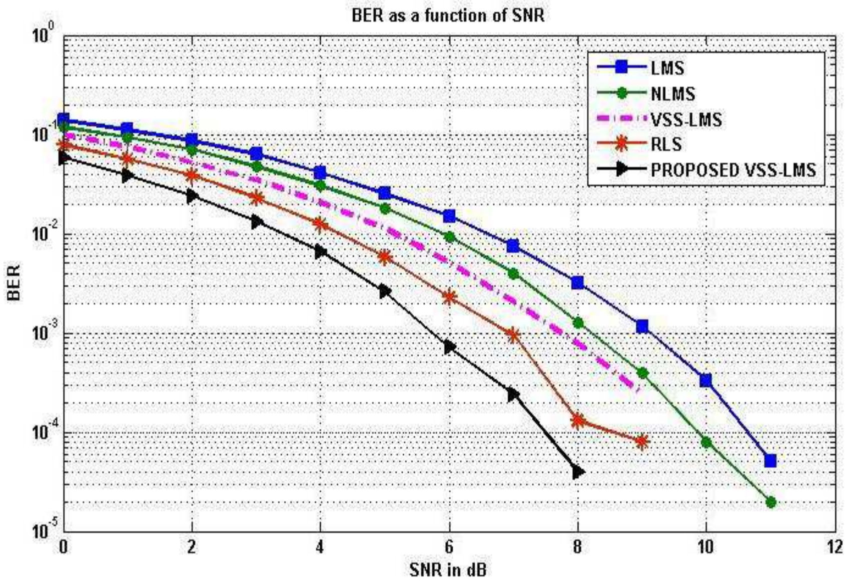

Fig. 3.4: BER performance of five algorithms as a function of SNR in AWGN channel for BPSK modulation.

The simulated results reveal that in an AWGN channel, for BPSK modulation, the proposed Variable Step Size LMS algorithm gives remarkable BER performance than the existing algorithm. At a BER of 10^{-4} the proposed algorithm outperforms the LMS algorithm by 3dB. The RLS algorithm is outperformed by the propose algorithm by 1.5dB at BER of 10^{-4} .

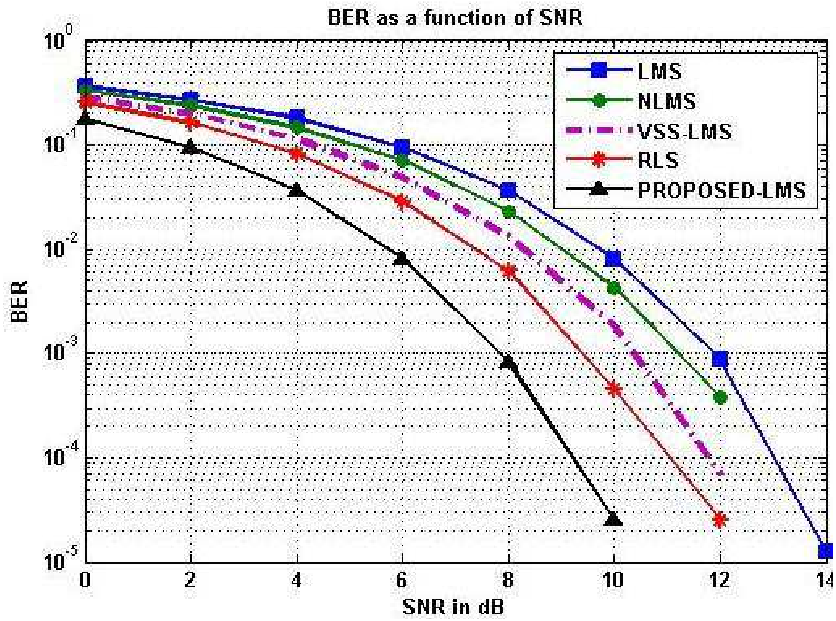

Fig. 3.5: BER performance of five algorithms as a function of SNR in AWGN channel for QPSK modulation.

For QPSK modulation, a BER performance is slightly degraded than in BPSK. The proposed algorithm exhibits an SNR of 7.5dB for BPSK modulation and an SNR of 9 dB for QPSK modulation. The proposed algorithm for BPSK modulation outperforms QPSK by 1.5dB at a BER of 10^{-4} .

3.5.2 Analysis of BER performance in Rayleigh fading channel

The estimation of channel is performed in a Rayleigh fading environment by taking into consideration effect of Doppler spread on the system. Slow and fast fading scenarios are tested. Fig.3 and Fig.4 shows the BER verses SNR plot for different algorithms at Doppler frequencies of 100Hz and 1000Hz respectively. The aforementioned algorithm outperforms the current algorithms in each of the results. It is noticed that the performance degrades with increase in Doppler frequency, ie, when doppler frequency increases from 100 Hz to 1000 Hz

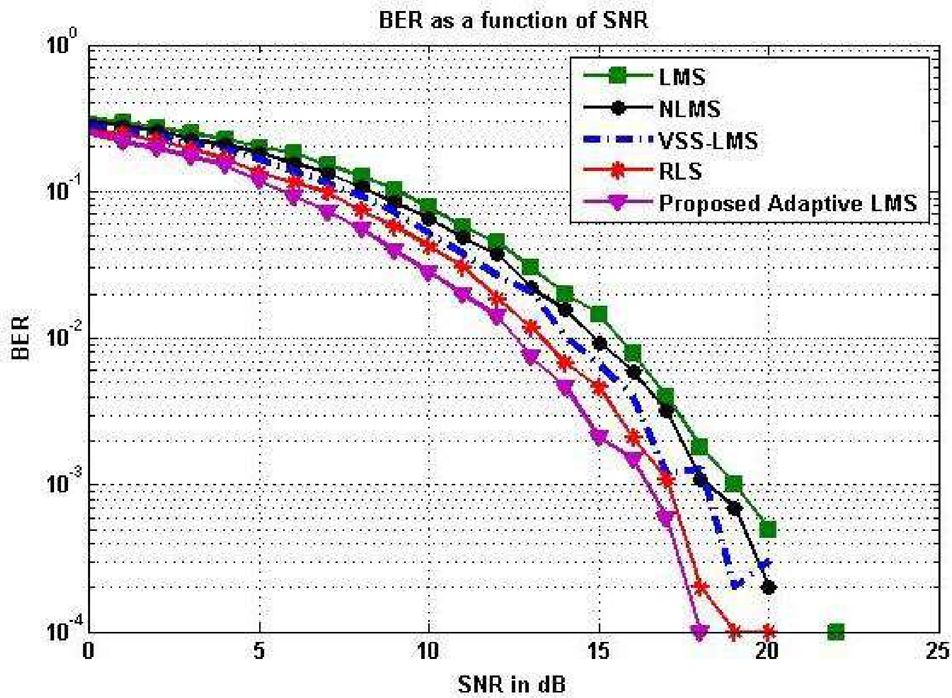

Fig. 3.6: BER of five algorithms as a function of SNR in Rayleigh fading channel at a doppler frequency of 100 Hz.

At a Doppler frequency of 100 Hz, the system behaves as slow fading, characterized by a BER performance that is worse than the one in AWGN channel. A degradation in performance is perceived in this case.

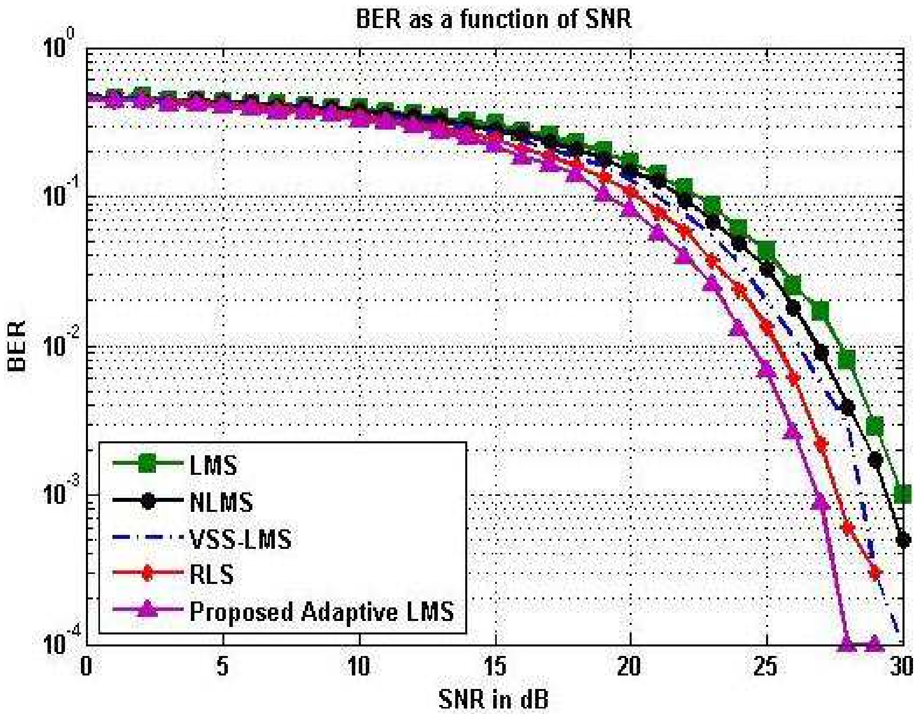

Fig. 3.7: BER performance of five algorithms as a function of SNR in Rayleigh fading channel at a Doppler frequency of 1000Hz.

A fast fading channel is simulated in Fig. 3.6 at a high doppler frequency of 1000 Hz. It is obvious that the bit error degrades to a large extent than in a slow fading channel, which was simulated at a Doppler frequency of 100 Hz.

3.5.3 Analysis of MSE performance in AWGN channel

The error function performance of the channel relative to SNR in AWGN channel is shown in Fig. It reveals better performance of the adaptive algorithm compared to the existing algorithms.

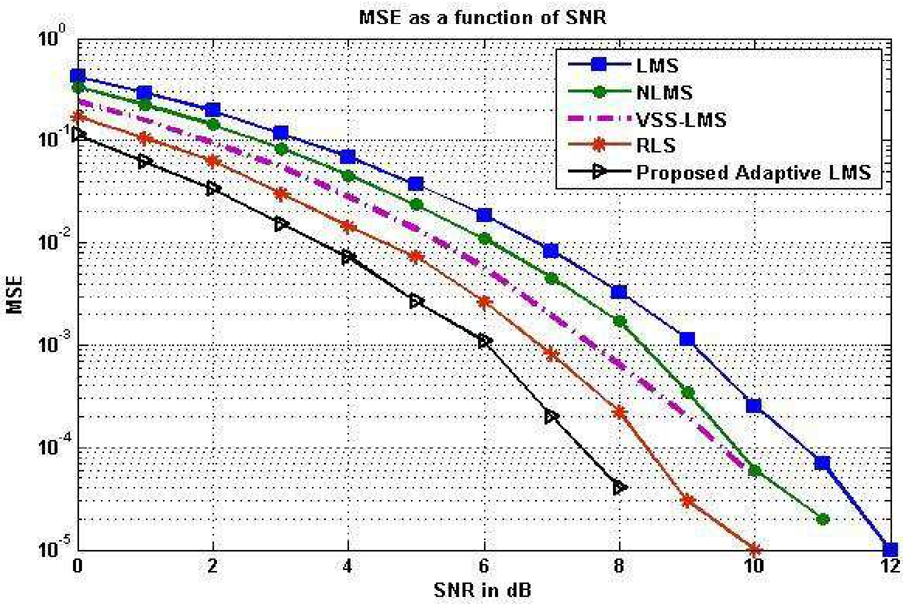

Fig. 3.8: MSE of five algorithms as a function of SNR in AWGN channel.

The MSE performance is analysed in AWGN channel, the proposed Adaptive LMS algorithm, gives good MSE performance than the existing algorithm.

3.5.3 Analysis of MSE performance in Rayleigh fading channel

Fig. 3.9 shows the performance of the error function with respect to SNR in a Rayleigh fading channel. MSE of the proposed algorithm yield remarkably good result than other algorithms. An increase in doppler frequency causes a fast fading of the signal, received at the receiver. High corruption in the received signal occurs as a result of high Doppler spread, thereby degrading the system performance.

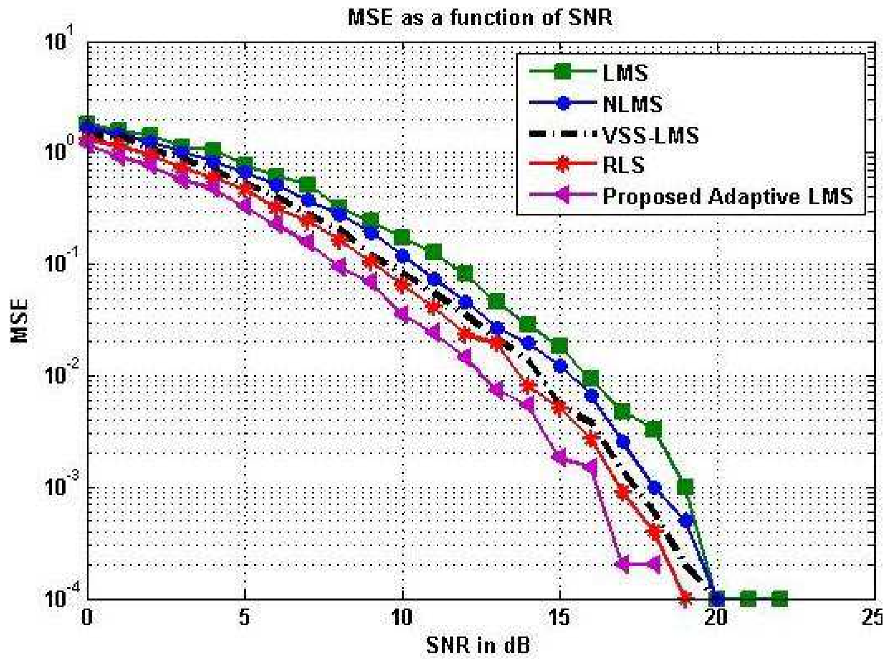

Fig. 3.9: MSE of five algorithms as a function of SNR in Rayleigh fading channel at a doppler frequency of 100 Hz.

I n a slow fading Rayleigh channel, at a Doppler frequency of 100 Hz, the five algorithms have been compared for its MSE performance. The proposed algorithm gives a low error performance than the existing algorithms.

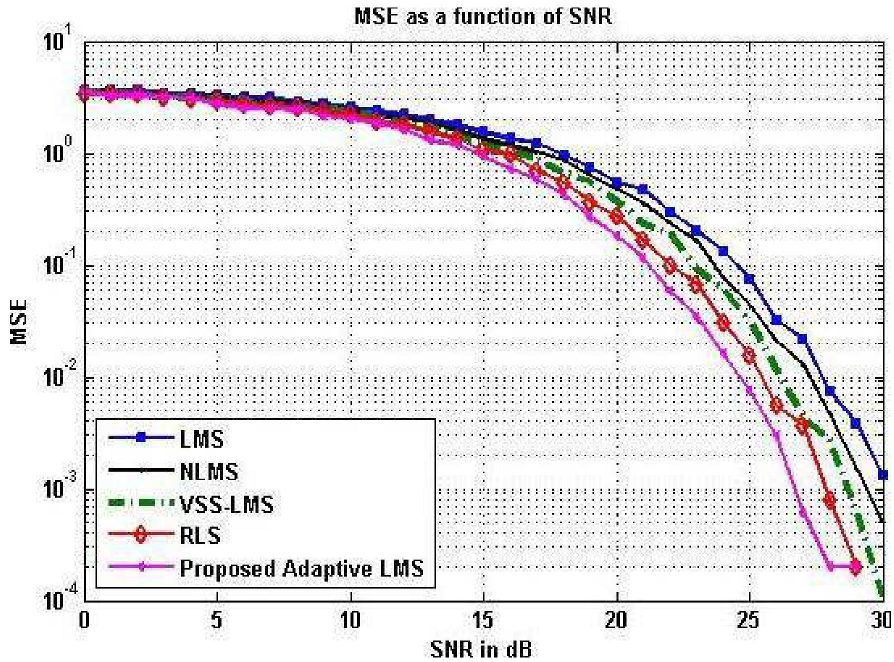

Fig. 3.10: MSE of five algorithms as a function of SNR in Rayleigh fading channel at a doppler frequency of 1000 Hz.

In a fast fading channel, ie, at a doppler frequency of 1000Hz, when the SCFDMA uplink system was simulated, the whole algorithm shows a high degrading performance, than in a slow fading channel.

3.6 Summary of the adaptive estimation technique

The following advantages are gained by using this adaptive uplink channel estimation for LTE.

- This CE algorithm uses phase weighting scheme such that the algorithm is less vulnerable to signal variations owing to noise as well as estimation errors, such an approach guarantees convergence towards the true channel vector.

- Time-varying step size parameter such that larger step size at the beginning of the iteration to accelerate the convergence rate of the algorithm, and uses smaller step size when the algorithm is convergent. Hence, this CE algorithm does not have the convergence speed toward the true channel coefficients-MSE trade off problem.

- Thirdly, the proposed estimate h(m) is an unbiased estimate of the tap weight vector w(m). An unbiased estimate indicates that its mean value is identical to the true parameter value. Consequently, as the number of observation increases, the estimate is assured to converge to the true parameter. Ideally, we would like our estimator to be unbiased and to have the smallest possible error variance.

- Measurements of the relevant channel correlation functions, nor does it require matrix inversion is required. Finally, the proposed scheme outperforms conventional methods with respect to the MSE and bit error rate (BER) of the estimated channel .

CHAPTER 4

CONCLUSIONS AND FUTURE WORK

4.1 Conclusions

A time-varying step size LMS channel estimation scheme is proposed so as to combat channel dynamics and support broadband multimedia access. The weighting coefficients are updated automatically, despite the unavailability of channel information. Besides, signals fluctuations due to noise decision errors can be nullified by the phase weighting scheme. Thus, the algorithm guarantees convergence towards accurate channel coefficient. Even though, the proposed CE technique requires little bit high computational complexity, the advantage in the performance of MSE, convergence towards true channel coefficient as well as BER performance could be of relevant use in future mobile communications which allow broadband multimedia access, anywhere, and anytime wireless communication.

4.2Future Work

The proposed work may be applied to a pilot inserted SCFDMA system so that more accurate estimation of the channel in uplink direction can be attained.

The proposed adaptive algorithm can be compared with the estimation using single and multilayer perceptron to evaluate its performance.

PUBLICATION

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