Joint Detection and Channel Estimation of LTE Downlink System using Unique Iterative Decoding Technique

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Abstract: - Long Term Evolution (LTE) is emerging as a strong option towards wireless broad band. Inclusion of Orthogonal Frequency Multiplexing (OFDM) and Multiple Input Multiple Output (MIMO) has exploited the efficient spectrum utilization, realizing high bandwidth potential and latencies in the current hostile wireless communication. LTE Advanced (LTE-A) is shaping up to be a pervasive technology with solutions that not only meet the ever-increasing data demand of traditional mobile broadband services, but also open up opportunities to transform new industries. In this paper we propose to improve the performance of channel estimation and detection for LTE downlink systems under the effect of increased channel length. Focus is on to the joint channel estimation and mitigation of inter-symbol Interference (ISI). Cyclic prefix (CP) is used to compensate for the ISI. Proposed iterative technique gives improved detection in terms of mean square error (MSE) as well as bit error rate (BER) for the given CP and channel length.

In this paper Iterative channel estimation using Least Square (LS) and Minimum Mean Square Error (MMSE) methods are explored to improve the channel estimation. With the use of iterative algorithms, it is shown that within adequate number of iterations channel estimation can be improved.

Key-Words: - MIMO, OFDM, Channel estimation, Least Square, LMMSE.

1 Introduction

Development of 4G mobile-broadband systems based on 3GPP LTE RAT (Radio Access Technology) is now progressing on a large scale [1], with 55 million users as of November 2012 and close to 1.6 billion users anticipated in 2018 [2]. The current commercial LTE deployments are based on 3GPP Release 8 and Release 9 – that is the first releases of the LTE technical specifications[3][4].

The first major step in the evolution of LTE – sometimes referred to as LTE-Advanced (LTE-A) occurred as part of 3GPP Release 10, which was finalized in 2010. Release 10 extended and enhanced LTE RAT in several dimensions. For example, the possibility was created for transmission bandwidth beyond 20 MHz and improved spectrum flexibility through carrier aggregation and enhanced multi-antenna transmission based on an extended and more flexible reference-signal structure. Advanced antenna techniques leverage more antennas on the device to offer higher data rates and capacity. LTE-A is shaping up to be a pervasive technology with solutions that not only meet the ever-increasing data demand of traditional mobile broadband services, but also open up opportunities to transform new industries. The evolution of LTE is a well thought-out roadmap with releases providing successively improving capabilities that result in higher and more consistent data rates for users, higher capacity, and a better overall user experience.

Multiple Input Multiple Output(MIMO) is the most attractive feature that is introduced in LTE and further advancement. It was first introduced by Alamouti for 2x2 antennas achieving full diversity gain. [5] Multiple antennas at both transmitter and receiver provide diversity and capacity gain. [6] This allows LTE-A system designer to achieve the required specifications laid in the standard. [7] Advancement in space time coding were introduced by Tarokh et all [8]. It has been found that the space time code and space frequency coding are the best suitable codes for MIMO systems. Orthogonal Frequency Division Multiplexing (OFDM) is a multi-carrier modulation technique which consists of dividing the frequency-selective fading channel into parallel flat-fading sub-channels. Introduction of guard interval, called cyclic prefix (CP) plays

important role in reducing ISI and inter-carrier interference (ICI). The inserted CP is assumed to be equal to or larger than the maximum channel delay[9].
One popular combination of MIMO and OFDM is the Space Time Block Coded OFDM (STBC-OFDM) which was first proposed in [10]. In addition to spatial and temporal diversity, the combination of MIMO-OFDM offers a third dimension of coding which achieves frequency diversity. This coding scheme known as Space-Frequency Block Coding (SFBC), which is capable of achieving other dimensional coding over space and frequency have been proposed in [11]. In addition, coding through spatial and frequency dimension offers implementation advantages [12]. However, in SFBC-OFDM, channel parameters need to be known at the receiver to recover the transmitted symbols. Therefore, channel estimation with acceptable level of accuracy and hardware complexity has become an important research topic for MIMO-OFDM systems.

Two approaches for channel estimation have been proposed in the literature. Blind channel estimation [13][14] which are based on the exploitation of the statistical information of the received symbols. It seems to be very attractive due to its bandwidth-saving advantage. However, the blind technique is limited to slow time varying channels and has higher complexity at the receiver. On the other hand, pilot aided channel estimation [15] using pilot sequences scattered in the transmitted signal and known at the receiver is simple to implement and can be applied to different types of channels although the use of pilots affect the data rate.
Channel estimation for OFDM systems in slow fading channels have been extensively studied [16][17]. These channel estimation techniques were developed under the assumption that the channel state does not change over one OFDM symbol, an assumption that may not suffice for mobile applications. Channel estimation for OFDM systems in fast fading channels is more challenging because the channel impulse response function varies within one OFDM symbol. In [18], both time domain and frequency domain correlations of the frequency response of rapid dispersive fading channels is exploited and proposed a minimum mean-square-error based channel estimator. In a similar work, Moon et. all introduced a channel estimation algorithm by adopting a Gaussian interpolation filter or a cubic spline interpolation filter [19]. However, both the algorithms in [18] and [19] require knowledge of channel statistics, which may not be available.
Channel estimation is critical in LTE Downlink MIMO-OFDM system. Many researchers assume that the length of the cyclic prefix should be greater than the channel length. But sometimes because of the channel behavior, cyclic prefix can be shorter than the channel length. At that condition channel estimation is difficult because of ISI and ICI introduced [9][20]. Different methods have been developed for converting the frequency-selective channel into multiple frequency-flat channels and eliminating the inter-symbol interference with a cyclic-prefix [21].

1.1 Organization of the Paper

Paper is organized as follows. Section 2 describes the LTE-A system model. LTE downlink physical structure is described in section 3. Various channel estimation techniques are explained in section 4. Simulation and discussion is carried out in section 5. Section 6 describes the conclusion of the work.

2 LTE Advanced System Model

In this chapter, we describe the LTE advanced down link communication system[20]. The structure of the transmitter is also presented depending on the 3GPP LTE advanced standard.

2.1 OFDM System Model

As shown in figure 1, binary information is first grouped, coded and mapped according to the modulation in a signal mapper. Different types of modulation schemes like quadrature phase shift keying (QPSK) or 16 QAM (Quadrature Amplitude Modulation) are used. After the guard band is inserted, an N-point inverse discrete-time Fourier transform $IDFT_N$ block transforms the data sequence into time domain. Following the IDFT block, a cyclic extension of time length T_G, chosen to be larger than the expected delay spread, is inserted to avoid intersymbol and intercarrier interferences. In this work different channel lengths relative to cyclic prefix are considered for comparison. The D/A converter contains low-pass filters with bandwidth $1/T_S$, where T_S is the sampling interval. The channel is modeled as an impulse response $g(t)$ followed by the complex additive white Gaussian noise (AWGN) $w(t)$, where α_m is a complex values and $0 \leq \tau_m T_S \leq T_G$.

$$g(t) = \sum_{m=1}^{L} \alpha_m \delta(t - \tau_m T_S) \qquad (1)$$

where α_m is the zero mean complex Gaussian random variable, g(t) is a time limited pulse train, Ts is the sampling period, τ_m is the delay of N^{th}

impulse, where the first delay is assumed to be zero and other impulse delay have been uniformly distributed over the length of the cyclic prefix. The requirement for orthogonality is the continuous time channel having length that is shorter than the cyclic prefix [22].
At the receiver, after passing though the analog-to-digital converter (ADC) and removing the CP, the DFT_N is used to transform the data back to frequency domain. Lastly, the binary information data is obtained back after the demodulation and channel decoding.

2.2 System Equations

By using N-point DFT, system can be modeled as

$$\bar{y} = DFT_N\left[IDFT_N(x) * \frac{\bar{g}}{\sqrt{N}} + \bar{w}\right]^T \qquad (2)$$

$$where, \bar{x} = [x_0, x_1, \dots, x_{N-1}]^T,$$
$$\bar{y} = [y_0, y_1, \dots, y_{N-1}]^T$$
$$\bar{w} = [w_0, w_1, \dots, w_{N-1}]^T$$

* denote the cyclic convolution and $\bar{g} = [g_0, g_1, \dots, g_{N-1}]^T$ is determined by sinc functions [16].

$$g_k = \frac{1}{\sqrt{N}}\sum_{m=0}^{N}\alpha_m e^{-j\frac{\pi}{N}}(k+(n-1)\tau_m)\frac{\sin(\pi\tau_m)}{\sin\left(\frac{\pi}{N}(\tau_m-k)\right)} \quad (3)$$

The amplitude α_m are the complex valued and $0 \le \tau_m T_s \le T_g$, k is an integer value in the range of 0 to N – 1. If τ_m is an integer then all the energy from i_m is mapped to taps g_k but when τ_m is not an integer then τ_m energy is located near the original pulse location[23].
The overall system response (received signal y_k) can be defined as

$$y_k = x_k h_k + n_k \qquad (4)$$

Here, k = 0, 1, 2, …N-1, h_k is the channel attenuation vector for 0 to N – 1 channel and g_k is the channel energy for N channels, where

$$h_k = [h_0, h_1, \cdots, h_{N-1}]^T = DFT(g_k)$$

For simplicity equation (4) can be written as follows:

$$\bar{y} = \bar{x}F\bar{g} + \bar{n} \qquad (5)$$

F is the DFT matrix.

2.3 MIMO OFDM System Model

Figure 2 depicts a high level block diagram of the MIMO OFDM system. We consider MIMO-OFDM systems with two transmit antennas and two receive antennas. The total number of subcarriers is *N*. Basically, the MIMO-OFDM transmitter has N_t parallel transmission paths which are very similar to the single antenna OFDM system, each branch performing serial-to-parallel conversion pilot insertion, N-point IDFT and cyclic extension before the final transmitted signals are up-converted to RF (Radio Frequency) and transmitted. It is worth noting that the channel encoder and the digital modulation, in some spatial multiplexing systems, can also be done per branch, where the modulated signals are then space-frequency coded using the Alamouti algorithm [5] before transmitting from multiple antennas not necessarily implemented jointly over all the N_t branches.

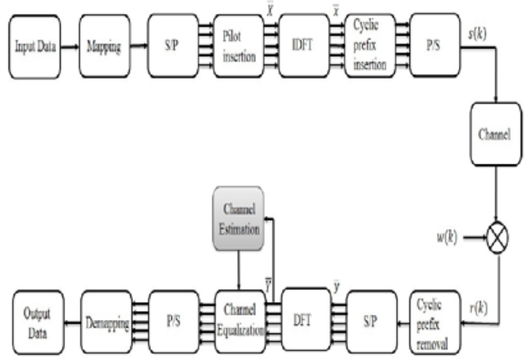

Figure 1: SISO OFDM System Block Diagram

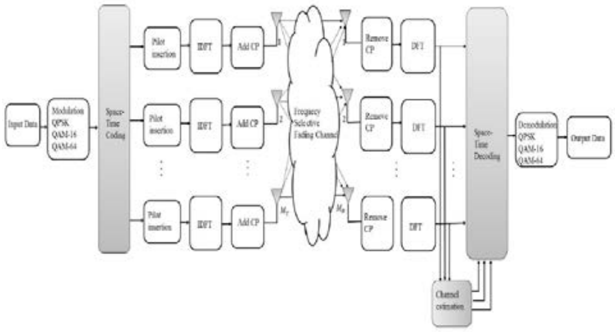

Figure 2: MIMO OFDM System Block Diagram

Subsequently at the receiver, the CP is removed and N-point DFT is performed per received branch. Next, the transmitted symbol per transmit antenna is combined and outputted for the subsequent operations like digital demodulation and decoding. Finally all the input binary data are recovered with certain bit error rate (BER).

As a MIMO signaling technique, N_t different signals are transmitted simultaneously over N_t x N_r transmission paths and each of those N_r received signals is a combination of all the N_t transmitted signals and the distorting noise. It brings in the diversity gain for enhanced system capacity as we desire. Meanwhile compared to the SISO system, it complicates the design regarding to channel estimation and symbol detection due to the hugely increased number of channel coefficients. The data stream from each antenna undergoes OFDM

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modulation. The Alamouti Space Time and Frequency Code scheme has full transmit diversity gain and low complexity decoder, with the encoding matrix represented as referred in [15].

$$X = \begin{bmatrix} X_1 & -X_2^* \\ X_2 & X_1^* \end{bmatrix} \quad (6)$$

The vectors X_1 and X_2 are modulated using the inverse DFT and after adding a cyclic prefix as a guard time interval, two modulated blocks $X_1^g \ and \ X_2^g$ are generated and are then transmitted by the first and second transmit antennas respectively. Assuming that the guard time interval is more than the expected largest delay spread of a multipath channel, the received signal will be the convolution of the channel and the transmitted signal. Assuming that the channel is static during an OFDM block, at the receiver side after removing the CP, the DFT output as the demodulated received signal can be expressed as

$$\begin{bmatrix} Y_1 \\ Y_2 \\ \vdots \\ Y_{N_R} \end{bmatrix} = \begin{bmatrix} H_{1,1} & H_{1,2} & \cdots & H_{1,N_t} \\ H_{2,1} & H_{2,2} & \cdots & H_{2,N_t} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N_r,1} & H_{N_r,2} & \cdots & H_{N_r,N_t} \end{bmatrix} \begin{bmatrix} X_1 \\ X_2 \\ \vdots \\ X_{N_t} \end{bmatrix} + \begin{bmatrix} W_1 \\ W_2 \\ \vdots \\ W_{N_r} \end{bmatrix} (7)$$

in the above equation $[W_1, W_2, \cdots, W_{N_r}]$ denotes AWGN and $H_{m,n}$ is the MIMO channel gain between the m^{th} receive and n^{th} transmit antenna pair. The n^{th} column of $\boldsymbol{H}$ is often referred to as the spatial signature of the n^{th} transmit antenna across the receive antenna array.

Knowing the channel information at the receiver, Maximum Likelihood detection can be used for decoding of received signals for two antenna transmission system.

3 LTE Downlink Physical Layer

The design of LTE physical payer is heavily influenced by the requirements for high peak transmission rate (100 Mbps Downlink (DL)/ 50 Mbps Uplink (UL)), spectral efficiency and multiple channel bandwidths (1.25 MHz-20 MHz).

3.1 LTE Frame Structure

Two types of radio frame structures are designed for LTE: Type-1 frame structures applicable to Frequency Division Duplex (FDD) and type-2 frame structure are related to Time Division Duplex (TDD). LTE frame structures are given in details in [3]. Figure 3 illustrates the structure of the LTE radio frame. The duration of one LTE radio frame is 10 ms. It is composed of 20 slots of 0.5 ms described in figure 3. Each time slot consists of either 7 or 6 OFDM symbols depending on the length of the CP (normal or extended). In LTE downlink physical layer, 12 consecutive subcarriers are grouped into one Physical Resource Block (PRB). A PRB has the duration of 1 time slot. Within one PRB, there are 84 resource elements for normal CP and 72 resource elements for extended CP. [3]

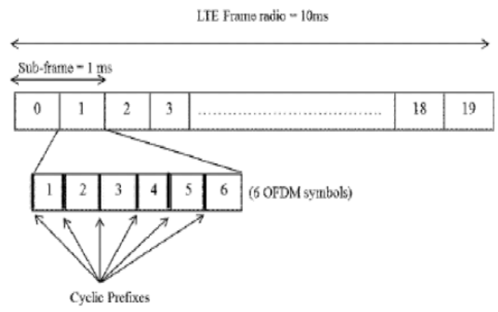

Figure 3 LTE Radio Frame Structure

3.2 Physical Resource Block (PRB)

In each slot, the transmitted signal consists of one or several physical resource blocks (PRB) where each individual resource element (RE) corresponds to one OFDM subcarrier during the respective OFDM symbol interval.

LTE supports scalable bandwidth from 1.25 MHz to 20 MHz and the physical resource block configuration parameters depend on the overall transmission bandwidth of the system. A physical resource block is the smallest element of resource allocation assigned by the base scheduler. For FDD mode, the frame structure is same for both downlink and uplink.

4 MIMO OFDM Channel Estimation

Channel estimation has important impact on the receiver performance. In LTE systems, channel estimation is performed based on the reference signals. The reference symbols are placed in the first OFDM symbol and the fifth OFDM symbol of every time slot when short CP is used while they are placed in the first and the fourth OFDM of every time slot when long CP is used. Figure 4 shows the reference signal pattern for two antennas.

The task now is to estimate the channel responses given the pilot signals X and the received signal Y.

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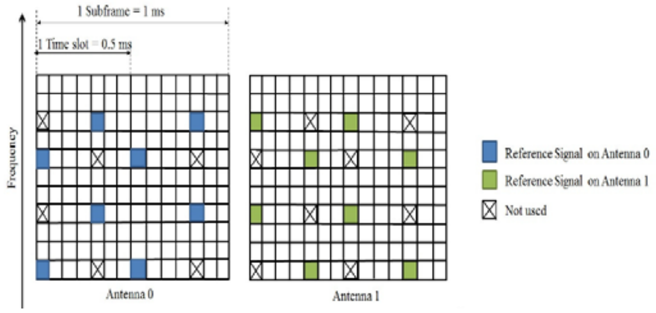

Figure 4 LTE Downlink Reference signal pattern for two antennas [20]

In this work, two linear channel estimators are studied: the Least Square (LS) and the Linear Minimum Mean Square Error (LMMSE).

4.1 Least Square (LS)

The LS estimate of such system is obtained by minimizing the square distance between the Received signal *Y* and the original signal *X*.

The least square estimates (LS) of the channel at the pilot subcarriers can be obtained as [16],

$$\hat{H}^{LS} = (x)^{-1}y \qquad (8)$$

where, x = transmitted symbols, y = received symbols.

The least square (LS) channel estimation is a simple estimation technique with very low complexity. It does not require any prior knowledge of the channel statistics. It is widely used because of its simplicity. However, it suffers from a high mean square error.

4.2 Linear Minimum Mean Square Error (LMMSE)

The LMMSE channel estimation employs the channel statistics to minimize the MSE estimate of the channel responses given in (5) is [16]:

$$H^{LMMSE} = R_{H\bar{H}}\left(R_{\bar{H}\bar{H}} + \sigma_\mu^2((XX^H)^{-1})\right)^{-1} \hat{H}^{LS} \quad (9)$$

$R_{H\bar{H}}$ is the cross-correlation matrix between all subcarriers and the subcarriers with reference Signals. $R_{\bar{H}\bar{H}}$ is the autocorrelation matrix of the subcarriers with reference signals. The problem with the LMMSE estimator is that it needs to apply the inversion $(XX^H)^{-1}$any time the data changes. In order to reduce the complexity of the LMMSE estimator, the term $(XX^H)^{-1}$ will be replaced by its expectation E $\{(XX^H)^{-1}\}$. If we assume the same signal constellation on all tones and equal probability on all constellation points, we obtain E $\{(XX^H)^{-1}\}$ – E $\{\left|{}^{1}/_{X_k}\right|^2\}$. By defining the average SNR as E $[|X_k|^2]$, the simplified LMMSE estimator becomes,

$$H^{LMMSE} = R_{H\bar{H}}\left(R_{\bar{H}\bar{H}} + \sigma_\mu^2((XX^H)^{-1})\right)^{-1} \hat{H}^{LS} \qquad (10)$$

Where β is scaling factor depending on the signal constellation (i.e. $\beta = 1$ for QPSK and $\beta = 16/9$ for 16-QAM) and is the identity matrix.

4.3 Proposed Iterative Channel Estimation

The channel estimation accuracy can be improved by adding virtual pilots using an iterative channel estimation and data detection. The hard decision symbols can be used as virtual pilots. Thus, there will be iteration between the decision and the channel estimation block at the receiver, which is a kind of decision feedback equalization technique, as seen in Fig 5[24]. Initially the channel is estimated using LS or MMSE method which gives initial estimation now this estimated data is detected and again fed back for iterative estimation that gives the improved performance than the conventional LS or MMSE method. This improved performance of LS is closer to the MMSE method. If we increase the number of iteration than mean square error performance is also improved. In iterative channel estimation initial channel estimation is performed by using pilot symbols and transmitted symbols are decoded. Then hard decision symbols are obtained by using these symbols. These symbols are sent back to the channel estimation block as virtual pilot symbols. After that, new channel estimation is performed.

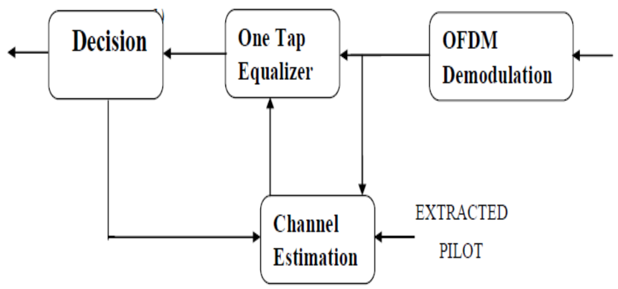

Figure 5: OFDM receiver for iterative channel estimation

The cost function for iteration is defined as

$$\hat{H}_n = \hat{H}_{n-1} - \mu \times e \times X \qquad (11)$$

where n = The iteration state, e = signal error, μ = step size, H = channel matrix, X = input data vector.

The iterative process is continued until the determined criteria will be reached, which it could be an especial MSE. As an initial estimate, LS and MMSE algorithms are used and successive iterations check for the error and according to cost function, modifies the next channel matrix. Important factor in channel estimation in this iterative method is the step size which influences on successive estimations after every iteration. Choice of the step size plays important role in convergence of the estimation algorithm.

5 Simulation Results and Discussions

Simulation is carried out for MIMO OFDM system using different channel estimation techniques. 2 x 2 antenna configurations are considered under the LTE downlink system under the effect of channel length. The modulation scheme of QPSK and 16 QAM are considered. The number of subcarrier in each OFDM symbol is N = 300, and the length of CP is L_{CP} = 36. 100 LTE radio frames are sent through a frequency-selective channel. The frequency selective fading channel responses are randomly generated with a Rayleigh probability distribution.

Channel model for OFDM with sampling interval N_s is given by:

$$h(n) = \delta(n) + \delta(n - 0.5N_s) + \delta(n - 3.5N_s) \quad (14)$$

Further to this, perfect carrier phase synchronization and zero offset in frequency of OFDM signals are assumed. Simulations are carried out using different channel lengths and CP of 36. The instantaneous mean square error (MSE) is defined as average error within an OFDM block.

Simulation has been carried out using LS and LMMSE channel estimation techniques as a primary estimate. Further iterations are carried out as per the cost function defined in (11).

It has been observed from figure 6 that LMMSE performs better than LS and iterative LS is improved compared to LMMSE and iterative LMMSE is best of all methods.

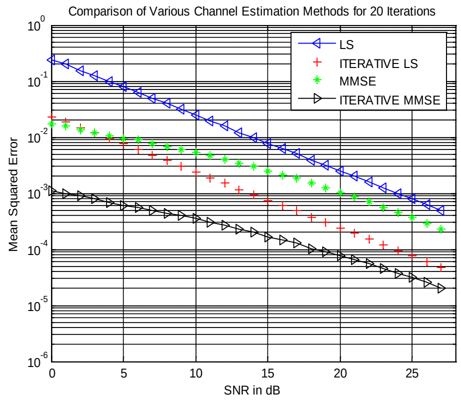

Figure 6: MSE performance of Iterative algorithm for 20 iterations for SISO OFDM

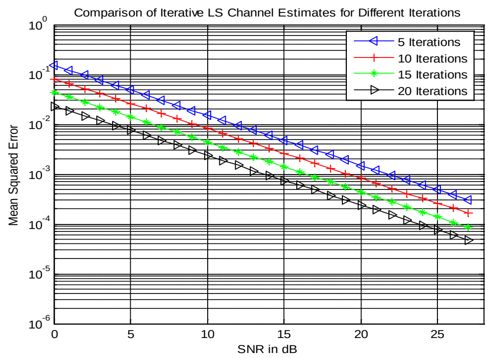

Figure 7: Comparison of Iterative LS for different iterations for SISO OFDM

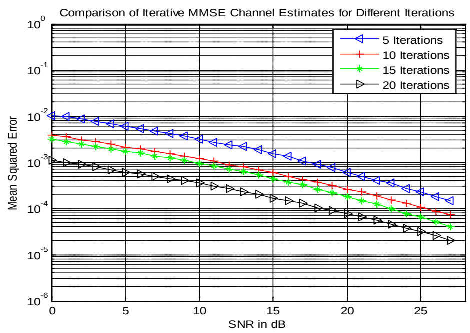

Figure 8: Comparison of Iterative LMMSE for different iterations for SISO OFDM

Figure 7 and 8 compares MSE of iterative LS and LMMSE techniques for different iterations. It has been found that as iterations increases, MSE decreases. Figures 9 and 10 show the BER performance of LTE downlink for 2x2 MIMO with different channel lengths. It has been observed that as the channel length increases, BER increases.

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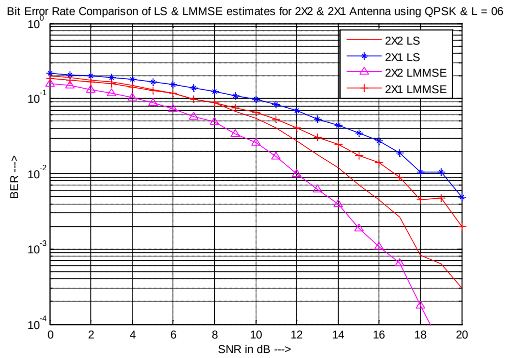

Figure 9: BER Comparison of LS/LMMSE MIMO OFDM for L = 06

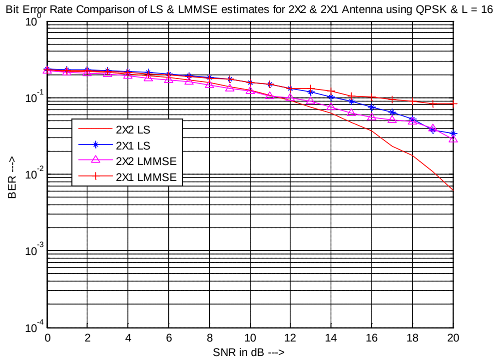

Figure 10: BER Comparison of LS and LMMSE for L = 16

Figure 11 shows iterative LS and LMMSE estimation results for different L. It can be seen that as L increases, BER decreases.

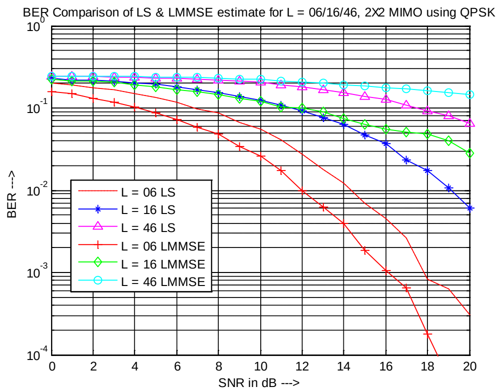

Figure 11: BER Comparison of LS and LMMSE for different channel length (MIMO-OFDM)

Figure 12 shows graph of iterative LS and LMMSE for different modulation schemes. It shows that as modulation increases, BER decreases. Figure 13 gives MSE results of different iterations for both methods and it can be observed that as iteration increases estimate improves. Figure 14 compares BER for both techniques using different iterations. As shown in figure 14, for higher iterations BER also improves.

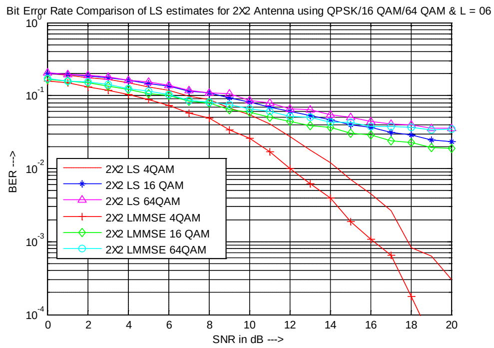

Figure 12: BER Comparison of LS and LMMSE MIMO OFDM for different modulations

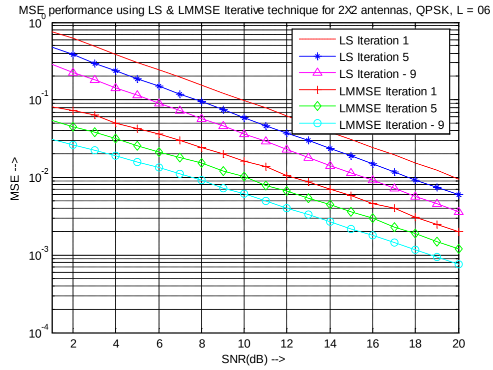

Figure 13: MSE Comparison of Iterative LS-MIMO OFDM for different iterations

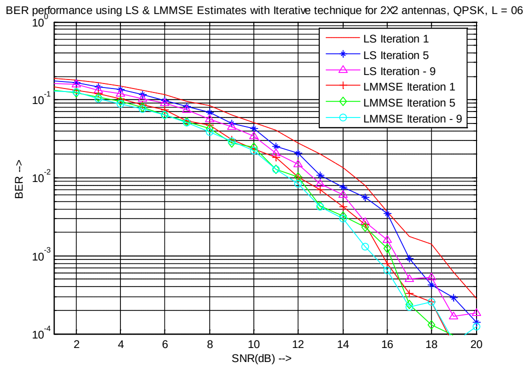

Figure 14: BER Comparison of Iterative LS & LMMSE-MIMO OFDM for different iterations

6 Conclusion

Simulation results show that in the case where the CP length is equal to or longer than the channel length, the LMMSE performs better than LS estimator but at the cost of the complexity because it depends on the channel and noise statistics. And by comparing QPSK and 16 QAM modulated technique, we can see that the BER performance is lost in 16 QAM than the QPSK technique as increasing SNR values.

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Iterative channel estimation improves the BER further to LMMSE estimate. As the iterations increases, it gives better performance. Even though the channel length increases, with sufficient number of iterations we can consistently achieve better performance than both the methods.

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