# Ls And QR Decompostion Channel Estimation Alagorithms for Mimo-OFDM Systems

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**Abstract**
Telecommunications in the current information age is increasingly relying on the wireless link. This is because wireless communication has made possible a variety of services ranging from voice to data and now to multimedia. Consequently, demand for new wireless capacity is growing rapidly at a very alarming rate. In a bid to cope with challenges of increasing demand for higher data rate, better quality of service, and higher network capacity, there is a migration from Single Input Single Output (SISO) antenna technology to a more promising Multiple Input Multiple Output (MIMO) antenna technology. On the other hand, Orthogonal Frequency Division Multiplexing (OFDM) technique has emerged as a very popular multi-carrier modulation technique to combat the problems associated with physical properties of the wireless channels such as multipath fading, dispersion, and interference. The combination of MIMO technology with OFDM techniques, known as MIMO-OFDM Systems, a major challenge in SISO-OFDM and MIMO-OFDM system is estimation of accurate channel state information (CSI) in order to make possible coherent detection of the transmitted signal at the receiver end of the system. In this paper, we propose iterative channel estimation based on QR Decomposition for MIMO OFDM systems. The aim of this paper is to investigate the effectiveness of QRD (QR decomposition) to reduce the computational complexity of channel estimation algorithms in MIMO-OFDM system, and design high performance channel estimation for this system by using iterative technique.

***Index Terms:***
*MIMO, OFDM, Channel Estimation*

## 1. Introduction:

MIMO employs multiple antennas at the transmitter and receiver sides to open up additional sub channels in spatial domain. Since parallel channels are established over the same time and frequency, high data rates without the need of extra bandwidth are achieved [7, 8]. Using this advantage have made the combination of MIMO-OFDM an attractive technique for future high data rate systems [10–12]. As in many other coherent digital wireless receivers, channel estimation is also an integral part of the receiver designs in coherent MIMO-OFDM systems [13]. In wireless systems, transmitted information reaches to receivers after passing through a radio channel. For conventional coherent receivers, the effect of the channel on the transmitted signal must be estimated to recover the transmitted information [14]. As long as the receiver accurately estimates how the channel modifies the transmitted signal, it can recover the transmitted information .Channel estimation can be avoided by using differential modulation techniques, however, such systems result in low data rate and in some cases, channel estimation at user side can be avoided if the base station performs the channel estimation and sends a pre-distorted signal [20]. However, for fast varying channels, the pre-distorted signal might not bear the current channel distortion, causing system degradation. Hence, systems with a channel estimation block are needed for the future high data rate systems. Channel estimation is a challenging problem in wireless systems. Unlike other guided media, the radio channel is highly dynamic. The transmitted signal travels to the receiver by undergoing many detrimental effects that corrupt the and often place limitations on the performance of the system. Transmitted signals are typically reflected and scattered, arriving at receivers along multiple paths. Also, due to the mobility of transmitters, receivers, or scattering objects, the channel response can change rapidly over time. Most important of all, the radio channel is highly random and the statistical characteristics of the channel are environment dependent. Multipath propagation, mobility, and local scattering cause the signal to be spread in frequency, time, and angle. These spreads, which are related to the selectivity of the channel, have significant implications on the received signal. Channel estimation performance is directly related to these statistics. Different techniques are proposed to exploit these statistics for better channel estimates. There has been some studies that cover these estimation techniques, however these are limited to the comparison of few of the channel estimation techniques [21–24]. This paper efficient channel estimation is firstly designed, and then it is simplified using alternative mathematical expressions.

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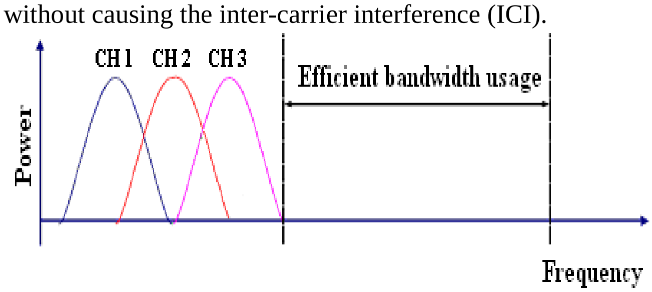
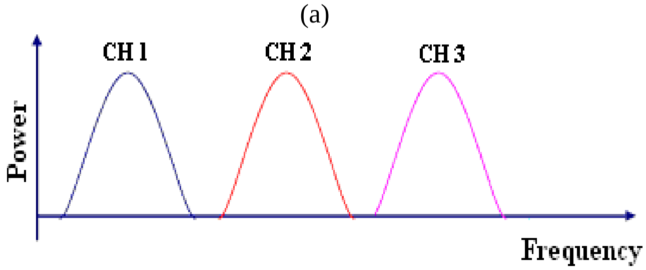
## 2. Realted Work

### 2.1. OFDM:

The OFDM concept is based on the splitting of data stream with a high-rate into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. Thus, there is an increase in symbol duration for the lower rate parallel subcarriers which in turn reduces the relative amount of dispersion that is caused by multipath delay spread, in time. However, in a bid to completely eliminate the intersymbol interference (ISI), a guard time is introduced in every OFDM symbol. As such, OFDM technology is seen as a scheme that transforms a frequency selective fading channel to a set of parallel flat fading sub-channels. Consequently, the receiver structure is drastically simplified, and the time domain waveforms of the sub-carriers become orthogonal to each other. In contrast to the normal Frequency Division Multiplexing (FDM) scheme where the subcarriers are non-overlapping, the signal frequency spectrum associated with different subcarriers overlap in frequency domain as shown in Figure 1.3. The introduction of guard band between the different carriers in the conventional FDM, in a bid to get rid of the interchannel interference, results in an inefficient use of the scarce and costly frequency spectrum resource. The overlapping of these subcarriers in the OFDM Systems makes possible efficient utilization of available bandwidth without causing the inter-carrier interference (ICI).




(a)



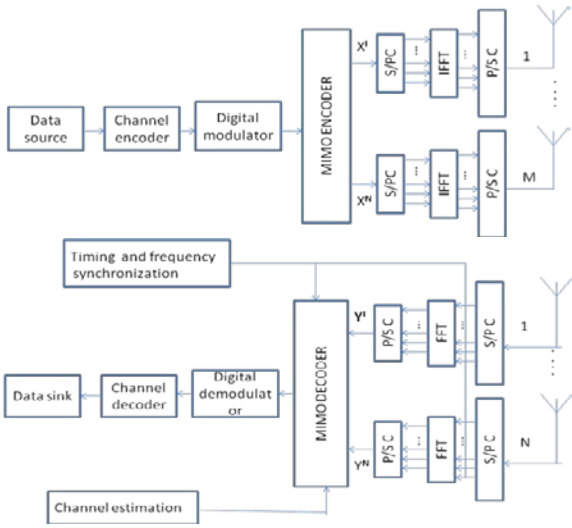

(b)

Figure.1 Comparison between (a) OFDM and (b) Conventional FDM

Direct implementation of OFDM Systems is computationally complex because of the large number of subcarriers involved which would require an equal number of sinusoidal oscillators for coherent demodulation. However, a breakthrough to the OFDM implementation came in 1971 when Weinstein and Ebert [21] proposed an effective way of implementing the scheme through the application of Discrete Fourier Transform (DFT), which drastically reduces the implementation complexity of the OFDM modems. This substantial reduction in implementation complexity was attributable to the simple realization that the DFT makes use of a set of harmonically related sinusoidal and cosinusoidal basis functions, whose frequency is an integer multiple of the lowest nonzero frequency of the set, which is referred to as the basis frequency. These harmonically related frequencies can therefore be used as the set of carriers required by the OFDM system. By employing a Fast Fourier Transform (FFT), an efficient implementation of the DFT, the computational complexity of OFDM could further be reduced. Recent advances in very-large-scale-integration (VLSI) technology have, however, enabled the availability of economical, high-speed and large-size integrated circuits for the implementation of FFT and (Inverse Fast Fourier Transform) IFFT. The use of these IFFT and FFT methods for the implementation of both the OFDM transmitter and receiver respectively reduces the number of operation to Klog2K from K2, if DFT techniques are used instead, where K is the number of subcarriers.

### 2.2. MIMO-OFDM SYSTEM MODEL:

The multiple transmitting and receiving antennas can be employed with OFDM to enhance the communication capacity and quality of mobile wireless Systems [25-28]. MIMO as described above is known to boost capacity. In the case of high data-rate transmission, the multipath nature of the communication environment causes the MIMO channels to become frequency-selective. However, as elucidated earlier, OFDM transmission scheme can convert such frequency-selective MIMO channels into an array of parallel frequency-flat MIMO channels by which the receiver complexity is drastically reduced. The combination of these two powerful techniques, MIMO and OFDM to form Multiple Input Multiple Output-Orthogonal Frequency Division Multiplexing (MIMO-OFDM) Systems, is very attractive, and is considered one of the most promising solutions to improve the signal rate of broadband wireless communication Systems. Figure.1 shows the basic model of MIMO-OFDM system with M and Nr number of antenna at the transmitter and receiver respectively. In this model, MIMO transmission is assumed to be OSTBC (Orthogonal Space-Time Block Coded). Therefore the block of user information after mapping in MPSK modulator is coded by the MIMO-STBC encoder with the matrix dimension of P×M. Where P is the number of time interval needed to transmit this matrix by M number of transmit antenna. It should be mentioned here that every elements of this coded matrix is an OFDM block with 64 symbols. Every columns of this matrix before transmission is fed to M number OFDM module. In this module, before adding cyclic prefix, IFFT transformation is performed for each element of the encoded matrix.




S/P C= Serial to Parallel Converter, P/S C= Parallel to Serial Converter,
FFT= Fast Fourier Transform, IFFT= Inverse Fast Fourier Transform

Fig. 2 MIMO-OFDM TRANCEVIER Model

If a column of encoded matrix which enter to the OFDM block is (X1, X2,…,Xm)T in frequency domain then the output of OFDM module will be (x1, x2,…,xm)T in time domain. Each element of encoded matrix Xk before OFDM module has a length of N= 64 symbols while after OFDM module change to xk in time domain with the length of 80 symbols.

The received signal after distortion by frequency selective channel and AWG noise at antenna j from antenna i can be represented by “Equation (1)”.

$$y^{ji}(n) = \sum_{l=0}^{L-1} h_l^{ji}(n) \cdot x^i(n-l) + w^j(n), i = 1,2,....M \quad (1)$$

where is $h_l^{ji}(n)$ is lth channel coefficient between received antenna j and transmitted antenna i at time n. WJ(n) is AWGN with zero mean and variance one. The “Equation (1)” in vector form can be rewritten by “Equation (2), (3), (4), (5)”.

$$y^{ji} = \left[ y_0^{ji} \quad y_1^{ji} \quad ,...., \quad y_{N+N_{CP}}^{ji} \right]^T \quad (2)$$

$$\begin{pmatrix} h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 & 0 & 0 \\ h^{ij}(1) & h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 & 0 \\ \vdots & h^{ij}(1) & h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 \\ h^{ij}(l-1) & \vdots & h^{ij}(1) & \ddots & 0 & \cdots & 0 & 0 \\ 0 & h^{ij}(l-1) & \vdots & \ddots & \ddots & 0 & \cdots & 0 \\ \vdots & 0 & h^{ij}(l-1) & \vdots & \ddots & \ddots & 0 & \vdots \\ 0 & \vdots & \vdots & \ddots & \vdots & \ddots & \ddots & 0 \\ 0 & 0 & 0 & 0 & h^{ij}(l-1) & \cdots & h^{ij}(1) & h^{ij}(0) \end{pmatrix} \quad (3)$$

$$x^{ji} = \left[ x_0^{ji} \quad x_1^{ji} \quad ,...., \quad x_{N+N_{CP}}^{ji} \right]^T \quad (4)$$

$$w^{ji} = \left[ w_0^{ji} \quad w_1^{ji} \quad ,...., \quad w_{N+N_{CP}}^{ji} \right]^T \quad (5)$$

The received signal at antenna j, is summation of transmitted signal by all transmit antenna i=1,2,…M. Hence it can be represented by as

$$y^i(n) = \sum_{i=0}^{M-1} \sum_{l=0}^{L-1} h_l^{ji}(n).x^i(n) + w^j(n) \quad (6)$$

Equation (6) in vector form can be rewritten by Equations (3), (7), (8), (9) as

$$y^j(n) = \left[ y_0^j, \quad y_{1,}^j \quad ,...., \quad y_{N+N_{CP}}^j \right]^T \quad (7)$$

$$x^j(n) = \left[ x_0^j, \quad x_{1,}^j \quad ,...., \quad x_{N+N_{CP}}^j \right]^T \quad (8)$$

$$v^j(n) = \left[ v_0^j, \quad v_{1,}^j \quad ,...., \quad v_{N+N_{CP}}^j \right]^T \quad (9)$$

Received signal vector at time interval t by antenna j=1,2,…,Nr can be represented as

$$y = \begin{pmatrix} \left[ y_0^1, \quad y_{1,}^1 \quad ....., \quad y_{N+N_{CP}}^1 \right]^T \\ \left[ y_0^2, \quad y_1^2, \quad ......, \quad y_{N+N_{CP}}^2 \right]^T \\ . \\ . \\ \left[ y_0^{N_r} \quad y_1^{N_r} \quad ......, \quad y_{N+N_{CP}}^{N_r} \right]^T \end{pmatrix} \quad (10)$$

$$y = \begin{pmatrix} \sum_{i=1}^{M} toeplitz(h^{i1}).\left[ x_0^i, \quad x_{1,}^i \quad ....., \quad x_{N+N_{CP}}^i \right] \\ \sum_{i=1}^{M} toeplitz(h^{i2})\left[ x_0^i, \quad x_1^i, \quad ......, \quad x_{N+N_{CP}}^i \right] \\ . \\ . \\ \sum toeplitz(h^{iNr}\left[ x_0^i \quad x_1^i \quad ......, \quad x_{N+N_{CP}}^i \right] \end{pmatrix}$$

Toeplitz function is a channel matrix function which can be defined as toeplitz(h) =

$$\begin{pmatrix} h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 & 0 & 0 \\ h^{ij}(1) & h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 & 0 \\ \vdots & h^{ij}(1) & h^{ij}(0) & 0 & \cdots & 0 & 0 & 0 \\ h^{ij}(l-1) & \vdots & h^{ij}(1) & \ddots & 0 & \cdots & 0 & 0 \\ 0 & h^{ij}(l-1) & \vdots & \ddots & \ddots & 0 & \cdots & 0 \\ \vdots & 0 & h^{ij}(l-1) & \vdots & \ddots & \ddots & 0 & \vdots \\ 0 & \vdots & \vdots & \ddots & \vdots & \ddots & \ddots & 0 \\ 0 & 0 & 0 & 0 & h^{ij}(l-1) & \cdots & h^{ij}(1) & h^{ij}(0) \end{pmatrix} (11)$$

After removing cyclic prefix and FFT transformation by OFDM demodulator, received signal in frequency domain can be represented as

$$Y_k^j = \sum_{i=1}^{M} H_K^{ji}.S_K^i + W_k^j \quad (12)$$

Wkj is AWGN in frequency domain and it can be calculated using

$$W_k^j = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} w^j.e^{-j2\pi kn/N} \quad (13)$$

Receive signal in vector form can be represented as

$$Y^j = \sum_{i=1}^{M} S^i H_K^{ji} + W^j \quad (14)$$

=S.HJ +WJ

In more detail each of variable in Equation (14) can be written as in Equations (15),(16),(17) and (18)

$$Y^j = \left[Y_1^j(n), \quad y_{1,}^j \quad ,...., \quad y_N^j(n)\right]^T \quad (15)$$

$$H_K^{ji} = \left(\left[H_1^{(1,J)}, \ H_2^{(1,J)}, \ \cdots \ ,H_N^{(1,J)}\right] \ ,\cdots, \ \left[H_1^{(M,J)}, \ H_2^{(M,J)}, \ \cdots \ ,H_N^{(M,J)}\right]\right)^T \quad (16)$$

$$S = \begin{pmatrix} S_1^1(n) & 0 & \cdots & 0 & \cdots & S_1^M(n) & 0 & \cdots & 0 \\ 0 & S_2^1(n) & \cdots & 0 & \cdots & S_1^M(n) & 0 & \vdots & 0 \\ \vdots & 0 & \ddots & \vdots & \vdots & \vdots & 0 & \ddots & \vdots \\ 0 & \cdots & 0 & S_N^1(1) & \cdots & 0 & \cdots & 0 & S_1^M(n) \end{pmatrix} \quad (17)$$

$$W^j = \left[W_1^j(n), \quad W_2^j(n), \quad \cdots \quad W_N^j(n)\right]^T \quad (18)$$

Received signal by antenna [1, 2 …Nr] in frequency domain after removing cyclic prefix and FFT transformation can be written as

$$Y = S \times H + W \quad (19)$$

In more detail each of variables in Equation (19) can be written as in Equations (17), (20), (21) and (22)

$$Y = \begin{pmatrix} Y_1^1(n) & y_N^2(n) & \cdots & y_N^{N_r}(n) \\ y_2^1(n) & y_N^2(n) & \cdots & y_N^{N_r}(n) \\ \vdots & \vdots & \vdots & \vdots \\ y_N^1(n) & y_N^2(n) & \cdots & y_N^{N_r}(n) \end{pmatrix} \quad (20)$$

$$H = \begin{pmatrix} H_1^{(1,1)} & H_1^{(1,2)} & \cdots & H_1^{(1,N_r)} \\ H_2^{(1,1)} & H_2^{(1,2)} & \cdots & H_2^{(1,N_r)} \\ \vdots & \vdots & \vdots & \vdots \\ H_N^{(1,1)} & H_N^{(1,2)} & \cdots & H_N^{(1,N_r)} \\ \vdots & \vdots & \vdots & \vdots \\ H_1^{(M,1)} & H_1^{(M,2)} & \cdots & H_1^{(M,N_r)} \\ H_2^{(M,1)} & H_2^{(M,2)} & \cdots & H_2^{(M,N_r)} \\ \vdots & \vdots & \vdots & \vdots \\ H_N^{(M,1)} & H_N^{(M,2)} & \cdots & H_N^{(M,N_r)} \end{pmatrix} \quad (21)$$

$$W = \begin{pmatrix} W_1^1(n) & W_N^2(n) & \cdots & W_N^{N_r}(n) \\ W_2^1(n) & W_N^2(n) & \cdots & W_N^{N_r}(n) \\ \vdots & \vdots & \vdots & \vdots \\ W_N^1(n) & W_N^2(n) & \cdots & W_N^{N_r}(n) \end{pmatrix} \quad (22)$$

## 2.3. LS ESTIMATION

From Equation (12), it can be seen that for estimation of channel component between receive antenna j and transmit antenna i=1,2,…M, the number of subcarriers which has to be estimated is M×N. where N is the number of subcarriers. In the other words for every receive antenna j=1,2,…Nr vector $H_k^{ji}$ in Equation(16) has to be estimated.

If one OFDM training block with N subcarriers transmitted from every of transmitted antenna, then from the model for every receive antenna there will be N equation with N×M unknown, hence these equations are under determined and can not be solved. For solving this problem there are two solutions, first solution is transmitting M OFDM blocks which in practical case is not applicable. Second solution is reducing the unknown elements by looking at an alternate representation of the received signal, called the transform-domain estimator that was first proposed by van de Beek in [4] for OFDM systems and well explained in [2] for MIMO-OFDM system. Base on this method CFR (Channel Frequency Response) can be expressed in terms of the CIR (Channel Impulse Response) through the Fourier transformation. Hence, the received signal model in “Equation (14)” can be expressed in terms of the CIR. The benefit of this representation is that usually the length of the CIR is much less than the number of subcarriers of the system. CIR representation can be achieved using following transformation

$$H^{(j,i)} = F \cdot h^{(j,i)} \quad (23)$$

Where $h^{(j,i)}$ is the (L×1) channel impulse vector and F is Fourier transform in vector form, and it can be represented as

$$F = \begin{pmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{\frac{-j2\pi(1)(1)}{N}} & \cdots & e^{\frac{-j2\pi(1)(L-1)}{N}} \\ \vdots & \vdots & \vdots & e^{\frac{-j2\pi(2)(L-1)}{N}} \\ 1 & e^{\frac{-j2\pi(N-1)(2)}{N}} & \cdots & \vdots \\ 1 & e^{\frac{-j2\pi(N-1)(1)}{N}} & \cdots & e^{\frac{-j2\pi(N-1)(L-1)}{N}} \end{pmatrix}_{N\times L} \quad (24)$$

To extend the matrix Fourier transform to operate on multiple channels following matrix in Equation (25) can be defined as

$$\Gamma = \begin{pmatrix} F & 0 & \cdots & 0 \\ 0 & F & \cdots & 0 \\ \vdots & \cdots & \ddots & \vdots \\ 0 & \cdots & 0 & F \end{pmatrix}_{M\times DIM(F)} \quad (25)$$

By using this definition, transformation of CFR to CIR in Equation (14) can be done as

$$Y^j = S^j H_K^j + W^j = X^j.\Gamma.h^j + W^j$$
$$A.h^J + W^j \quad (26)$$

By applying LS algorithm on Equation (26), channel component can be estimated using Equation (27). By this transformation one OFDM block is enough to estimate the channel. The only condition is N≥M×L

$$\tilde{h}^{j} = (A^{H}.A)^{-1}.A^{H}.Y \quad (27)$$

## 2.4. QR DECOMPOSITION

QR decomposition is just an alternative for calculating matrix inversion. There are different methods for QR decomposition. Here Householder algorithm is used. The steps of QRD algorithm to solve LS problem can be presented as follow[5]:1. Making the LS error function for Equation (27) as represented in Equation (28).

$$\varepsilon = Y - A\tilde{h} \text{ and if } \varepsilon = 0 \Rightarrow Y = A\tilde{h} \quad (28)$$

2. Decompose W into Hermitian matrix Q and upper triangular matrix R using Householder algorithm as Equation (29).

$$Y = A\tilde{h} = Q_{M\times M}\cdot\begin{bmatrix} R \\ 0 \end{bmatrix}_{M\times N}.\tilde{h} \quad (29)$$

3. Multiply Hermitian of Q to both side of Equation (29). The result can be represented as in Equation (30).

$$\begin{bmatrix} R \\ 0 \end{bmatrix}_{M\times N}\cdot\tilde{h} = Q_{M\times M}{}^{H}\cdot Y \quad (30)$$

4. Finally, solve the channel using back substitution

## 2.5. Complexity Comparison Between LS and QRD Algorithm

The advantage of using QR decomposition is to reduce the computational complexity of the LS channel estimation. In this research, the computational complexity in terms of number of mathematical operations has been measured. The derivations are based on an Mt-by-Mr MIMO-OFDM system with N subcarriers and a channel length of L. The known matrix A has dimensions (N ×L.Mt). For simplicity in notation L.Mt is denoted by M. For a consistent comparison, the complex operations are converted to real operation equivalents.

Table 2 shows the real equivalent operations for the various complex operations. In addition, each type of real operations has different levels of complexity when implemented in the hardware. For example multiplications, additions, and subtractions can be set to 1 FLOPs (Floating Point Operations), divisions to 6 FLOPs, and square roots to 10 FLOPs (table-1). It should be emphasized that counting of the number operations is only an estimate of the computational complexity of the algorithms. A more exact measure would be to implement the algorithm in hardware and count the number of instructions and processing time required. However, in computer simulations, FLOP counts can give a good indication of the relative complexity of different algorithm.

Table 1. Number of flops in every real operation

| Operation | # No of Flops |
|---|---|
| Multiplication, addition and subtraction | 1 |
| Division | 6 |
| Square root | 10 |

Table 2. Number of real operations in every complex operation

| **Complex Operation** | **No of Real Operations** | | |
|---|---|---|---|
| | Multiplication | Division | Subtraction & Addition |
| Multiplication | 4 | 2 | 0 |
| Division | 6 | 3 | 2 |
| Subtraction /Addition | 0 | 0 | 2 |
| Complex magnitude | 2 | 0 | 1 |

# 3. Iterative QRD Channel Estimation

In order to complete data-aided channel estimation, pilot signals can be spaced separated in the transmitted symbols. In the receiver, the channel impulse response can be estimated at the positions of pilot signals by several algorithms, such as least square method. The other channel information at the data signals can be obtained by interpolating the estimated channel impulse response[6].

In this paper,we propose a iterative LS-QRD channel estimation algorithm for MIMO OFDM system.At first step ,an LS-QRD channel estimate is obtained by using (31)

$$\hat{H}^{j}_{QRD} = \Gamma\tilde{h}^{j} \quad (31)$$

$\tilde{h}$ can be obtained from $\begin{bmatrix} R \\ 0 \end{bmatrix}_{M\times N}\cdot\tilde{h} = Q_{M\times M}{}^{H}\cdot Y$

Where

$$\Gamma = \begin{bmatrix} F & F & \cdots & F \\ 0 & F & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & F \end{bmatrix} \quad F = e^{-j2\pi mn/N}, \quad 0 \le m \le N-1, 0 \le n \le L-1$$

And then the channel estimate is set as intial $\hat{H}^{j}_{QRD}{}^{(k)}, k = 0$ .Secondly, the receiver uses the estimated channel to help the detection/decision of data signals. The detection data can be obtained by zero forcing method

$$\hat{X}^{i}_{ZF} = \hat{H}^{(k)}_{QRD}{}^{\Phi}Y^{j} \quad (32)$$

Where $\hat{H}^{(k)}_{QRD}{}^{\Phi}$ is pseudo inverse of $\hat{H}^{j}_{QRD}$

$$\hat{H}_{QRD}^{(k)} = \begin{bmatrix} \hat{H}_{QRD,diag}^{1,1(k)} & \hat{H}_{QRD,diag}^{2,1(k)} & \cdots & \hat{H}_{QRD,diag}^{N_t,1(k)} \\ \hat{H}_{QRD,diag}^{1,2(k)} & \hat{H}_{QRD,diag}^{2,2(k)} & \cdots & \hat{H}_{QRD,diag}^{Nr,2(k)} \\ \vdots & \vdots & \vdots & \vdots \\ \hat{H}_{QRD,diag}^{1,N_r(k)} & \hat{H}_{QRD,diag}^{2,N_r(k)} & \cdots & \hat{H}_{QRD,diag}^{N_t,N_r(k)} \end{bmatrix}$$

And then the channel estimation treats the detected signals as known data to perform a next stage channel estimation iteratively and the index k adds 1. Go to the first step and repeat the process till the mean-square-error of channel estimate is converged or the expected iterations reach. By utilizing the iterative channel estimation and signal detection process we can reduce the estimation error caused by channel interpolation between pilots. The accuracy of the channel estimation can be improved by increasing the number of iteration process.

## 4. Simulation Results

The system specification for this simulation can be summarized in Table 3. For this simulation the channel has L=16 paths where the amplitude of each path varies independently according to the Rayleigh distribution with an exponential power delay profile [10], and can be represented as in Equation (33). The results can be classified into two parts; Performance comparison and complexity comparison results. These are presented in the next sections.

Table 3. Simulation parameters for MIMO-OFDM

| System | MIMO(STBC)-OFDM |
|---|---|
| #Rx Antenna | 2 |
| #Tx Antenna | 2 |
| Channel | Frequency selective, Rayleigh fading |
| Noise | AWGN |
| #Sub carrier | 64 |
| #Cyclic Prefixes | 16 |
| Cahnnel length | 16 |
| Trms (RMS delay spread) | 25 ns |
| Ts-Sampling Frequency | 1/80 MHz |

$$h_i = N(0,1/2\sigma^2) + jN(0,1/2\sigma_l^{\,2}) \qquad l = 1,2,......L-1 \qquad (33)$$

Where $\sigma_l^2 = (1 - e^{\frac{-T_s}{T_{RMS}}}) \times e^{\frac{-LT_s}{T_{RMS}}}$ and for normalization $\sum_{l=0}^{L-1} \sigma_l^{\,2} = 1$ and L approximated by

L= $\dfrac{10T_{rms}}{T_s}$

### 4.1. Iterative QRD Algorithm

The bit error rate (BER) and MSE Performance of iterative QRD channel estimation method are shown in figure 2 and 3.Iteration number r=0 means the conventional QRD channel estimation method . After about 2 iterations ,the BER and MSE performance of iterative channel estimation are much closed to that of ideal one .

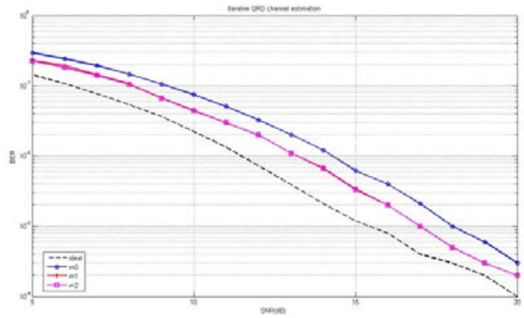


Figure.2 BER Performance of iterative channel estimation with different iteration numbers

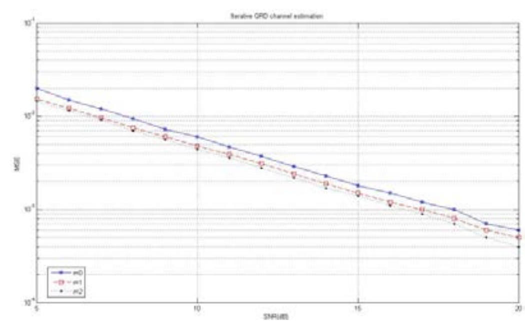


Figure.3 MSE Performance of iterative channel estimation with different iteration numbers.

The use of iterative QRD method improve the performance in terms of lower channel estimation error. From the results it can be concluded that the iterative QRD channel estimation algorithm have high performance efiiciency in terms of BER and MSE.However, in the next section the benefit of QRD, which is the significant reduction of the complexity of the system, is portrayed.

### 4.2 Complexity Comparison Between LS And QRD Algorithm

Using the system parameters for the MIMO-OFDM system specified in Table 3, the number of operations for a 2 transmit antenna system with a channel length of 6 and 16 was calculated for the two algorithms. In this section, the complexity comparison in terms of FLOPs count is performed for two algorithms. Figures 4 and 5 show the complexity comparison of both algorithms.
In Figure 4, the number of channel length is hold at 6

while the number of transmit antenna vary from 1 to 6. The results show the computational complexity of LS algorithm is higher than QRD.

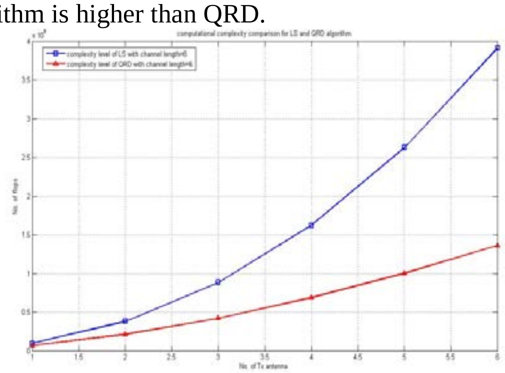


Figure. 4 Complexity comparisons between LS and QRD with channel length=6

The results in Figure 5 is more highlighted which the number of channel length increase to 16. Increasing the channel length increases the number of unknown parameters, thereby will increase the complexity of the channel estimation. It shows that the LS increases exponentially as the channel length increases and has much higher complexity than the QRD for long channel lengths

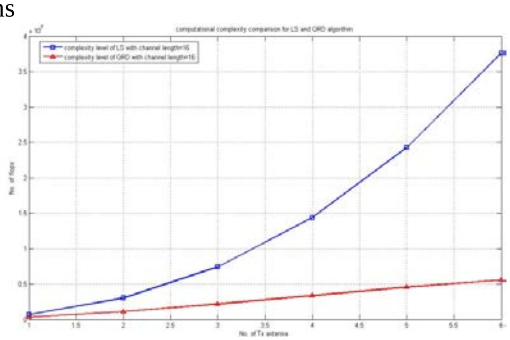


Figure.5 Complexity comparisons between LS and QRD with channel length=16

Figure 6 shows the simulation result using 2 transmit antenna while the number of channel component vary from 1 to 16. The previous conclusion for computational complexity can be made here. In Figure 7, the number of transmit antenna is increased to 8 while the channel is changed from 1 to 16. As expected, when the number of antennas increases, both estimation techniques increase in complexity because the size of the unknown matrix **A** increases. The general trend of the QRD method is that it increases almost linearly with the number of transmits antennas of the system. The LS method increases exponentially at a considerably higher rate than the QRD methods. Therefore, the QRD is especially preferable for higher number of transmit antennas since it does not explode in complexity as the LS solution. Finally the Numerical example for computational complexity comparison between two channel estimation algorithms is provided in Table 4.

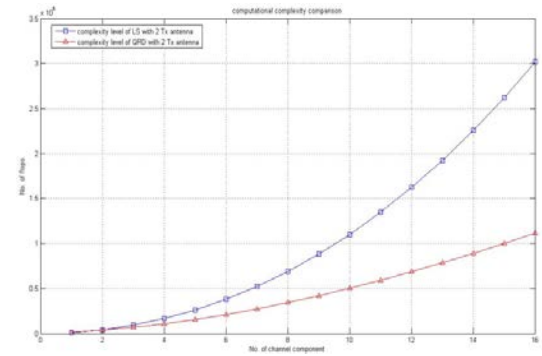


Figure. 6 Complexity comparisons between LS and QRD with TX=2

Table 4 .Number of complex operations and Flops in two algorithms with L=16 and Tx =4 (A: # of complex ultiply, B: # of add/sub, C: # of complex division, D: # of square root, E: # of complex magnitude, F: # of flops).

| Algorithm | A | B | C | D | E | F |
|---|---|---|---|---|---|---|
| LS | 790528 | 7822772 | 6176 | 0 | 0 | 14373568 |
| QRD | 187220 | 1840471 | 192 | 2144 | 128 | 3376126 |

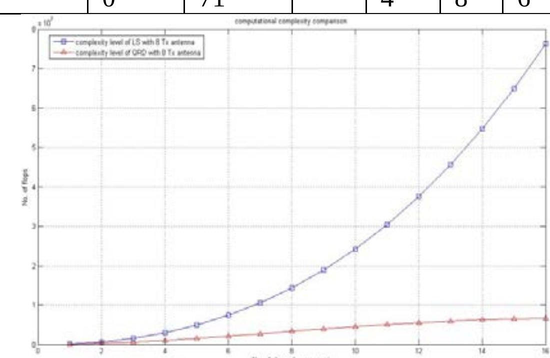


Figure 7. Complexity comparisons between LS and QRD with TX=8

The results prove that QRD method is lower in complexity than LS method. The results in Table 4 show that the total number of operation for the LS method is much higher than the QRD method. For this simulation scenario using QRD achieves a complexity reduction by approximately 77%. This verifies that the QRD has significantly lower complexity than that of direct LS estimation via the pseudoinverse, hence a better option for channel estimation.

## CONCLUSION

The simulation results prove that Iterative QRD channel estimation algorithm has good performance efficiency it can provide better mean square error and bit error rate performance than conventional methods. However the computational complexity of the QRD channel estimation is much lower than LS algorithm. In addition, computational complexity for QRD channel estimation is approximately linearly proportional with number of transmit antenna and channel length, whereas for LS algorithm is exponentially proportional with the number of transmit antenna and channel length. As finding indicate; using QRD channel estimation, computational complexity of the system for above particular scenario which mentioned in table-4 can dramatically decrease by 77 %.

Finally it can be concluded that Iterative QR decomposition can be an ultimate solution for high performance efficiency and reduction computational complexity .

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