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# Linear Approximation Signal Detection Scheme in MIMO-OFDM Systems

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**Abstract:** In this paper, a linearly approximate signal detection scheme is proposed in multiple input multiple output-orthogonal frequency division multiplexing (MIMO-OFDM) systems. The huge MIMO-OFDM system, which uses many transmit antennas and high order modulation, requires a detection scheme at the receiver with very low complexity for practical implementation. In the proposed detection scheme, one  $N \times N$  MIMO-OFDM system is divided into  $N/2 \ 2 \times 2$  MIMO-OFDM systems for linear increase of complexity. After the partial zero-forcing (ZF), decision feedback equalizer (DFE) and QR decomposition- $M$  algorithm (QRD- $M$ ) are applied to each  $2 \times 2$  MIMO-OFDM system. Despite nonlinear detection schemes, the overall complexity of the proposed algorithm increases almost linearly because the DFE and the QRD- $M$  are applied to  $2 \times 2$  MIMO-OFDM systems. Also, the value of  $M$  in the QRD- $M$  is fixed according to position of the center point in constellation for efficient signal detection. In simulation results, the proposed detection scheme has higher error performance and lower complexity than the conventional ZF. Also, the proposed detection scheme has very lower complexity than the conventional DFE, with slight loss of error performance.

**Keywords:** MIMO-OFDM; partial ZF; DFE; QRD- $M$

## 1. Introduction

Multiple input multiple output-orthogonal frequency division multiplexing (MIMO-OFDM) system, which is a combination of the MIMO and OFDM, is a core technology for high speed mobile communication systems. The MIMO systems provide a high channel capacity and high data rate, without additional bandwidth and high transmit power when compared to single input-single output (SISO) systems [1,2]. However, received signal is a mixed form of the several distorted transmit signals that go through the fading channel. At the receiver, maximum likelihood (ML) detection has optimal error performance [3,4]. The ML calculates the squared Euclidean distance (SED) between the received signals and all of the reference signals in the constellation. However, the ML cannot be implemented in the huge MIMO-OFDM system due to very high complexity. Unlike the ML, the zero-forcing (ZF), which is popular for linear detection can be implemented easily in the huge MIMO-OFDM system [5–8]. Also, ZF based decision fusion schemes have been studied for low complexity in huge MIMO systems [9,10]. However, the error performance for the linear detection scheme is poor as the number of transmit antennas increases and its error performance is not acceptable in real-time systems. The decision feedback equalizer (DFE) and QR decomposition- $M$  algorithm (QRD- $M$ ), which are nonlinear detection schemes, have higher error performance than the linear detection schemes [11–18]. Among several nonlinear detection schemes, the DFE and the QRD- $M$  use QR receiver for low complexity. However, the complexity for the nonlinear detection scheme increases exponentially as the number of transmit antennas increases, and these algorithms cannot

be implemented in the huge MIMO-OFDM systems. Specifically, the complexity for the QRD- $M$  increases exponentially when the modulation order is high, due to tree structures. In [19–22], many detection algorithms were developed to reduce the complexity for tree search algorithms. In [19–21], the proposed algorithms are based on sphere decoding (SD) and  $K$ -best decoding. These algorithms have a very lower complexity than the conventional detection scheme and the ML, with similar error performance. Specifically, [19,20] represented complexities as the average number of metric operations at all layers with respect to signal-to-noise ratio (SNR). Also, the average number of metric operations is decreased very sharply according to increased SNR. However, the number of metric operations converges to certain point and the decrease of the complexity is limited although the value of SNR goes to infinity. Also, [22] proposed the semidefinite relaxation (SDR) based scheme. However, the complexity order is still high to use in the huge MIMO-OFDM system.

In this paper, linearly approximate signal detection scheme is proposed for easy implementation in the huge MIMO-OFDM system. The proposed scheme is composed of three stages. The first stage is partial ZF, which is a core scheme to reduce the whole complexity for the proposed algorithm. The purpose of the partial ZF is that it divides one  $N \times N$  MIMO-OFDM system into several  $2 \times 2$  MIMO-OFDM systems. After the partial ZF, the number of  $2 \times 2$  MIMO-OFDM systems is  $N/2$ . The second and the third stages are DFE and QRD- $M$  to increase the error performance because noise power may be increased in the partial ZF. Also, the DFE and the QRD- $M$  are applied to  $2 \times 2$  MIMO-OFDM systems, and it does not require high complexity. Due to the partial ZF, the whole complexity for the proposed algorithm increases linearly according to the increased number of transmit antennas.

## 2. System Model

In this system model, the MIMO-OFDM system has  $N$  transmit antennas and receive antennas. Also, it is assumed that  $N = 2^P$  ( $P \geq 2$ ) is considered where  $P$  is natural number, which is larger than one. At the MIMO-OFDM transmitter, an input data is demultiplexed into  $N$  data substreams. Then,  $N$  data substreams go through digital modulation, inverse fast Fourier transform (IFFT) for multi-carrier transmission. Finally, guard interval (GI) is added to reduce the effect of inter-symbol interference (ISI) and inter carrier interference (ICI). At the receiver,  $N \times 1$  MIMO received vector  $\mathbf{Y}$  is as follows,

$$\mathbf{Y} = \mathbf{H}\mathbf{X} + \mathbf{Z}, \quad (1)$$

where  $\mathbf{H}$  is  $N \times N$  rich scattering complex Rayleigh flat fading channel matrix,  $\mathbf{X}$  is  $N \times 1$  transmit vector, and finally,  $\mathbf{Z}$  is  $N \times 1$  zero mean additive white Gaussian noise vector. The channel matrix  $\mathbf{H}$  is represented as follows,

$$\mathbf{H} = \begin{bmatrix} H_{11} & H_{12} & \cdots & H_{1N} \\ H_{21} & H_{22} & \cdots & H_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N1} & H_{N2} & \cdots & H_{NN} \end{bmatrix}, \quad (2)$$

where an element  $H_{ij}$ ,  $i, j = 1, 2, \dots, N$  denotes channel coefficient from the  $j$ -th transmit antenna to the  $i$ -th receive antenna. Also, each  $H_{ij}$ ,  $i, j = 1, 2, \dots, N$  is independent and identically distributed (i.i.d) random variables.

## 3. Conventional Detection Schemes

In this section, conventional detection schemes are represented, i.e., ZF, DFE, and QRD- $M$ , which are basic schemes for explanation of the proposed detection.

### 3.1. ZF

The ZF removes inter-antenna interference (IAI) perfectly by multiplying pseudo-inverse channel matrix by received vector. The pseudo-inverse channel matrix  $\mathbf{G}$ , which is a least square (LS) solution to estimate the transmit signal, is as follows,

$$\mathbf{G} = \left( \mathbf{H}^H \mathbf{H} \right)^{-1} \mathbf{H}^H. \tag{3}$$

Then, the estimated MIMO transmit vector  $\hat{\mathbf{X}}_{ZF}$  is as follows,

$$\hat{\mathbf{X}} = \mathbf{G} \mathbf{Y} = \mathbf{X} + \bar{\mathbf{Z}}, \tag{4}$$

where  $\bar{\mathbf{Z}} = \mathbf{G} \mathbf{Z}$  is a modified noise vector.

### 3.2. DFE

For higher error performance than the linear detection, the DFE uses ordered decision feedback and successive interference cancellation at each layer. The DFE has three stages, i.e., ordering, QR decomposition, and finally, decision feedback and IAI cancellation. At first, the DFE decides the cancelling order with respect to the norm value of each row of the  $\mathbf{G}$  in Equation (3) to minimize the error propagation. Likewise, each column of the  $\mathbf{H}$  is also ordered with respect to above order and the ordered channel matrix  $\mathbf{H}_S$  is defined as follows,

$$\mathbf{H}_S = \mathbf{H} \mathbf{U} \tag{5}$$

where  $\mathbf{U}$  is column change matrix, which is composed of 1 and 0.

Then,  $\mathbf{H}_S$  is decomposed by using QR decomposition. In the QR decomposition,  $\mathbf{Q}$  denotes unitary quadrature matrix, which is satisfied with  $\mathbf{Q}^H \mathbf{Q} = \mathbf{I}$  and  $\mathbf{R}$  denotes the upper triangular matrix. The QR decomposition of the  $\mathbf{H}_S$  is as follows,

$$\begin{aligned} \mathbf{H}_S &= \mathbf{Q} \mathbf{R} \\ &= \begin{bmatrix} Q_{11} & Q_{12} & \cdots & Q_{1N} \\ Q_{21} & Q_{22} & \cdots & Q_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ Q_{N1} & Q_{N2} & \cdots & Q_{NN} \end{bmatrix} \begin{bmatrix} R_{11} & R_{12} & \cdots & R_{1N} \\ 0 & R_{22} & \cdots & R_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & R_{NN} \end{bmatrix}. \end{aligned} \tag{6}$$

Using Equation (6), the MIMO received vector in Equation (1) can be rewritten as follows,

$$\mathbf{Y} = \mathbf{H} \mathbf{X} + \mathbf{N} = \mathbf{H}_S \mathbf{X}_S + \mathbf{N} = \mathbf{Q} \mathbf{R} \mathbf{X}_S + \mathbf{N}, \tag{7}$$

where  $\mathbf{X}_S = \mathbf{U}^{-1} \mathbf{X}$  is ordered transmit vector.

Multiplying  $\mathbf{Q}^H$  on both sides of Equation (7), the resulted  $N \times 1$  vector  $\mathbf{V}$  is as follows,

$$\mathbf{V} = \mathbf{Q}^H \mathbf{Y} = \mathbf{R} \mathbf{X}_S + \tilde{\mathbf{N}}, \tag{8}$$

where  $\tilde{\mathbf{N}} = \mathbf{Q}^H \mathbf{N}$  is modified noise vector which has the same statistical property as existing noise vector  $\mathbf{N}$  because  $\mathbf{Q}$  is unitary orthogonal matrix.

The vector form of  $\mathbf{V}$  without noise is as follows,

$$\begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{bmatrix} = \begin{bmatrix} R_{11} & R_{12} & \cdots & R_{1N} \\ 0 & R_{22} & \cdots & R_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & R_{NN} \end{bmatrix} \begin{bmatrix} X_{S,1} \\ X_{S,2} \\ \vdots \\ X_{S,N} \end{bmatrix}. \tag{9}$$

Finally, decision feedback and interference cancellation are operated from the bottom layer to the top layer in Equation (9). The estimated MIMO transmit symbol at the  $n$ -th layer  $\hat{X}_n$  is as follows,

$$\hat{X}_{S,n} = Q\left(\left(V_n - \sum_{i=n+1}^N R_{ni}\hat{X}_i\right)/R_{nn}\right), \tag{10}$$

where  $Q(\cdot)$  denotes quantization operator.

To estimate the MIMO transmit vector,  $\hat{\mathbf{X}}_S$  is multiplied by  $\mathbf{U}$  as follows,

$$\hat{\mathbf{X}} = \mathbf{U}\hat{\mathbf{X}}_S. \tag{11}$$

The DFE has higher error performance than the ZF due to detection ordering. However, the complexity of the DFE is higher than the ZF.

### 3.3. QRD-M

In Equation (9),  $|S|$  times SEDs between  $Z_N$  and reference symbol  $C_k$ ,  $k = 1, 2, \dots, |S|$  are calculated at the  $N$ -th layer, where  $|S|$  is constellation size. The SED between  $Z_N$  and the  $k$ -th reference symbol  $C_k$  is as follows,

$$E_{N,k} = |V_N - R_{NN}C_k|^2. \tag{12}$$

With respect to  $k$  in Equation (12), the SED vector  $\mathbf{E}_N$  is as follows,

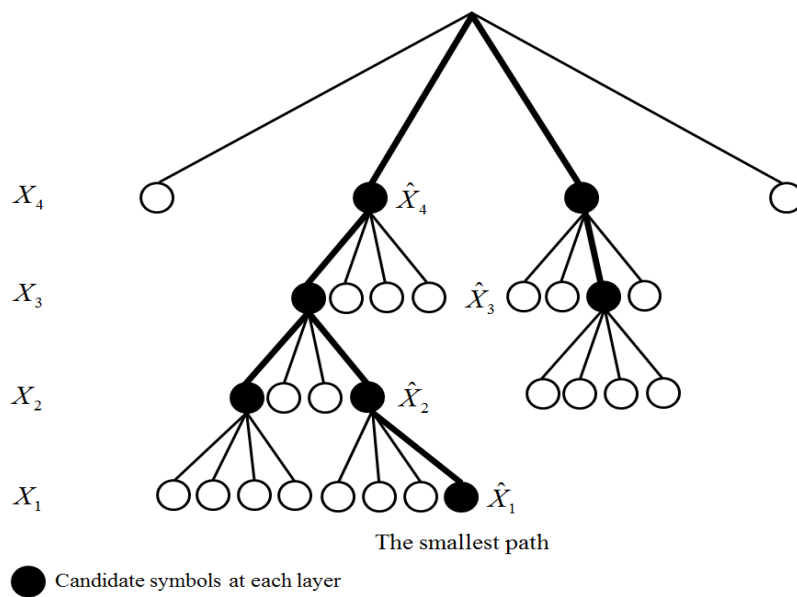
$$\mathbf{E}_N = \begin{bmatrix} E_{N,1} & E_{N,2} & \cdots & E_{N,|S|} \end{bmatrix}. \tag{13}$$

In Equation (13), the small  $M$  values are selected and its corresponding symbols are selected as candidate symbols at the  $N$ -th layer. The selected  $M$  candidates are extended to the  $(N - 1)$ -th layer. At the  $m$  ( $1 \leq m \leq N - 1$ )-th layer,  $M|S|$  times SEDs are calculated. However, accumulated SED (ASED) is calculated for accurate estimation. The ASED between  $V_m$  and  $C_k$ , which considers the ASED for the  $d_{m+1}$ -th candidate symbol, is as follows,

$$E_{m,k}^{d_{m+1}} = \left| V_m - \left( R_{mmm}C_k + \sum_{i=m+1}^{N_t} R_{mi}\hat{X}_i^{d_i} \right) \right|^2 + E_{m+1,d_{m+1}}^{d_{m+2}}, \tag{14}$$

where  $\hat{X}_i^{d_i}$  is temporarily estimated symbol at the  $i$ -th layer corresponding to the  $d_i$ -th candidate symbol. At the first layer, transmit symbols are estimated by selecting the candidate that has the smallest ASED with respect to  $k$ .

Figure 1 shows the tree structure of the conventional QRD-M ( $M = 2$ ) in  $4 \times 4$  MIMO-OFDM system using quadrature phase shift keying (QPSK) modulation. In Figure 1, bold circles are candidate symbols at each layer. The complexity of the conventional QRD-M increases exponentially in huge MIMO-OFDM systems because the conventional QRD-M calculates  $M|S|$  times ASEDs at each layer.



**Figure 1.** The tree structure of the conventional QR decomposition- $M$  algorithm (QRD- $M$ ) ( $M = 2$ ) in  $4 \times 4$  multiple output-orthogonal frequency division multiplexing (MIMO-OFDM) system using quadrature phase shift keying (QPSK) modulation.

#### 4. Proposed Detection Scheme

To reduce the complexity of nonlinear detection scheme, the proposed detection scheme is composed of three stages, i.e., partial ZF (slightly different to the conventional ZF in Section 3.1), DFE, and QRD- $M$ . For easy understanding and detailed explanation, the proposed detection scheme is explained in  $4 \times 4$  MIMO-OFDM system without noise vector as follows,

$$\begin{bmatrix} Y_1 \\ Y_2 \\ Y_3 \\ Y_4 \end{bmatrix} = \begin{bmatrix} H_{11} & H_{12} & H_{13} & H_{14} \\ H_{21} & H_{22} & H_{23} & H_{24} \\ H_{31} & H_{32} & H_{33} & H_{34} \\ H_{41} & H_{42} & H_{43} & H_{44} \end{bmatrix} \begin{bmatrix} X_1 \\ X_2 \\ X_3 \\ X_4 \end{bmatrix}. \quad (15)$$

##### 4.1. Partial ZF Stage

The partial ZF stage nulls not all signals unlike the conventional ZF. The purpose of the partial ZF is to simply divide  $4 \times 4$  channel matrix into two  $2 \times 2$  channel matrices by using cancelling process. That is, one matrix contains only  $X_1$  and  $X_2$ , another matrix contains only  $X_3$  and  $X_4$ . The cancelling process is as follows,

$$\begin{bmatrix} \tilde{Y}_3 \\ \tilde{Y}_4 \end{bmatrix} = \begin{bmatrix} Y_1 \\ Y_2 \end{bmatrix} - \mathbf{W}^T \begin{bmatrix} Y_3 \\ Y_4 \end{bmatrix}, \quad (16)$$

where  $\mathbf{W} = \begin{bmatrix} w_{11} & w_{12} \\ w_{21} & w_{22} \end{bmatrix}$  is a weight matrix as follows,

$$\begin{bmatrix} H_{31} & H_{32} \\ H_{41} & H_{42} \end{bmatrix}^T \begin{bmatrix} w_{11} & w_{12} \\ w_{21} & w_{22} \end{bmatrix} = \begin{bmatrix} H_{11} & H_{12} \\ H_{21} & H_{22} \end{bmatrix}^T. \quad (17)$$

The modified received vector  $\begin{bmatrix} \tilde{Y}_3 & \tilde{Y}_4 \end{bmatrix}^T$  in Equation (16) is as follows,

$$\begin{bmatrix} \tilde{Y}_3 \\ \tilde{Y}_4 \end{bmatrix} = \begin{bmatrix} \tilde{H}_{31} & \tilde{H}_{32} \\ \tilde{H}_{41} & \tilde{H}_{42} \end{bmatrix} \begin{bmatrix} X_3 \\ X_4 \end{bmatrix}, \quad (18)$$

where  $\tilde{H}_{ij}$  is modified channel coefficient.

From above results, the received vector has only  $X_3$  and  $X_4$ . So, the dimension of the modified channel matrix in Equation (18) is half of the dimension of existing channel matrix in Equation (15). The modified channel matrix is as follows,

$$\begin{bmatrix} \tilde{H}_{31} & \tilde{H}_{32} \\ \tilde{H}_{41} & \tilde{H}_{42} \end{bmatrix} = \begin{bmatrix} H_{13} & H_{14} \\ H_{23} & H_{24} \end{bmatrix} - \mathbf{W}^T \begin{bmatrix} H_{33} & H_{34} \\ H_{43} & H_{44} \end{bmatrix}. \quad (19)$$

Likewise, another modified received vector  $\begin{bmatrix} \tilde{Y}_1 & \tilde{Y}_2 \end{bmatrix}^T$  which has only  $X_1$  and  $X_2$  is obtained in a similar method. Another modified received vector is as follows,

$$\begin{bmatrix} \tilde{Y}_1 \\ \tilde{Y}_2 \end{bmatrix} = \begin{bmatrix} \tilde{H}_{11} & \tilde{H}_{12} \\ \tilde{H}_{21} & \tilde{H}_{22} \end{bmatrix} \begin{bmatrix} X_1 \\ X_2 \end{bmatrix} + \begin{bmatrix} \tilde{Z}_1 \\ \tilde{Z}_2 \end{bmatrix} \quad (20)$$

Therefore, the original  $4 \times 4$  MIMO-OFDM system is divided into two  $2 \times 2$  MIMO-OFDM systems.

#### 4.2. DFE Stage

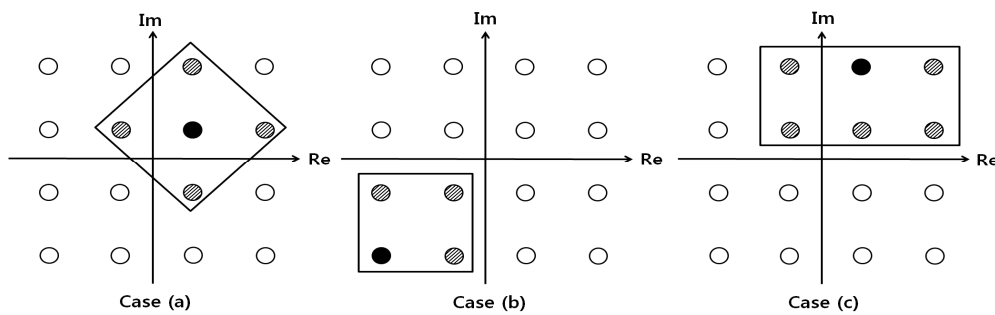
The conventional DFE in Section 3 is applied to the two systems in Equations (18) and (20), respectively. So, the complexity of the DFE stage is lower than the conventional DFE, which is simply applied in original  $4 \times 4$  MIMO-OFDM system. After the DFE stage, all of the transmit symbols from  $\hat{X}_1$  to  $\hat{X}_4$  are initially estimated.

#### 4.3. QRD-M Stage

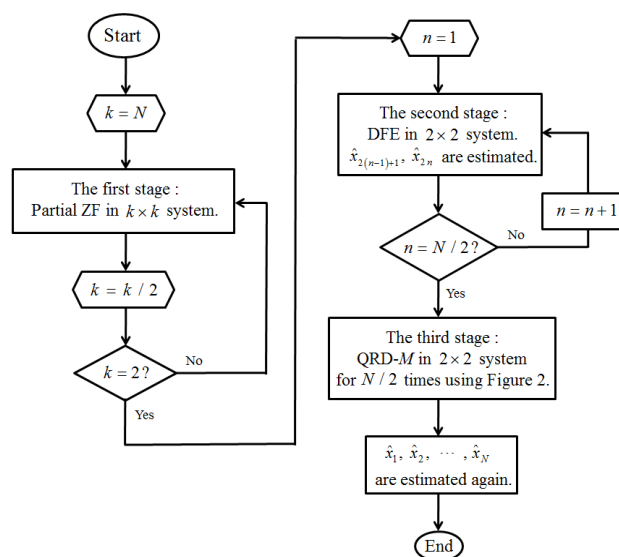
The QRD-M is applied to the two MIMO-OFDM systems in Equations (18) and (20). For low complexity QRD-M, the value of  $M$  is fixed with respect to the used modulation order, like Figure 2. Figure 2 shows the method for candidate selections in the 16-quadrature amplitude modulation (QAM) constellation with respect to three cases. The candidates are both center point (black colored circle) and neighbor points (dash-lined circles). The center point is a symbol, which is estimated by the DFE stage, as shown in Section 4.2. Also, the neighbor points are symbols that are within one point from the center point. The case (a) is that the center point is in the middle of the constellation and the case (b) is that the center point is in the edge of the constellation. Finally, the case (c) is that the center point is the nearest point from the edge point. The value of  $M$  is different according to the estimated center point, and it is fixed as 5, 4, 6 for cases (a–c), respectively. The neighbor points may have high probability for the original transmit symbol if the DFE stage fails to estimate the accurate transmit symbols. From the QRD-M, all of the transmit symbols are estimated again and all processes of the proposed detection scheme are completed. With a slight increase of the complexity, the estimated symbols are more reliable than the estimated symbols in the DFE stage. Table 1 represents the number of complex multiplications for the proposed detection scheme in  $N \times N$  MIMO-OFDM system. Also, it is assumed that one multiplication between two complex numbers requires four real multiplications. Finally, Figure 3 shows the flow chart for the proposed detection scheme. It is composed of three stages which are explained in Sections 4.1–4.3.

**Table 1.** The number of complex multiplications ZF: zero-forcing; DFE: decision feedback equalizer; QRD-M: QR decomposition-M algorithm).

Detection Scheme	The Number of Complex Multiplications
Conventional ZF	$8N^3 + 4N^2$
Conventional DFE	$12N^3 + 4N^2 + 4 \sum_{k=2}^N (k + 1) + 4$
Proposed scheme (Partial ZF stage)	$\sum_{n=2}^{N/2} (12n^2 + 4n)N$ ( $n = 2, 4, 8, \dots, N/2$ )
Proposed scheme (DFE stage)	$64N$
Proposed scheme (QRD-M ZF stage)	$(4M^2 + 8M)N$
Group Iterative Linear ZF in [7]	$14N^3 + 8N^2$
Hybrid ZF and QR Receiver in [8]	$15N^3 + 6N^2 + 4 \sum_{k=2}^{N/2} (k + 1) + 4$
Hybrid dynamic QRD-M and ZF in [16]	$12N^3 + 4N^2(1 + \lceil N/3 \rceil) + 4N(\lceil N/3 \rceil^2 + \lceil N/3 \rceil)8L + 4LM \sum_{l=2}^{N_l - \lceil N/3 \rceil + 1} (l + 1)$ where $\lceil N/3 \rceil$ denotes the integer number which is larger than $N/3$ or equal to $N/3$ .



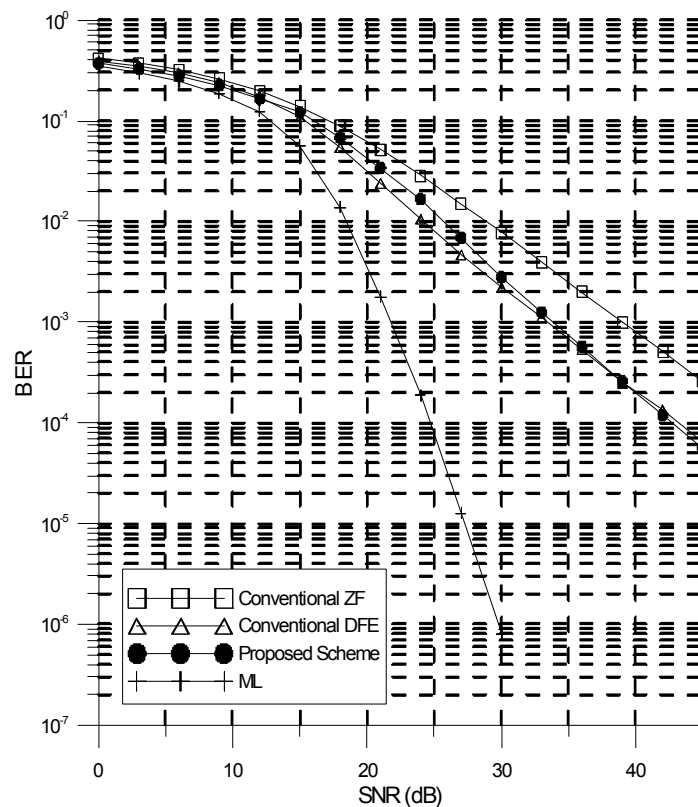
**Figure 2.** The method for candidate selections in the 16-QAM constellation (a): center point is in the middle of the constellation; (b): center point is in the edge of the constellation; (c): center point is the nearest point from the edge point (QAM: quadrature amplitude modulation).



**Figure 3.** The flow chart of the proposed detection scheme (ZF: zero-forcing; DFE: decision feedback equalizer).

### 5. Simulation Results

This section shows the simulation results for bit error rate (BER) performance and complexity. In the simulation results, the used modulation scheme is 16-QAM. Also, the number of used total subcarriers is 128 and length of the cyclic prefix (CP) is 32. Figure 4 shows the BER performances for the conventional ZF, DFE, and proposed detection scheme in  $4 \times 4$  MIMO-OFDM system. In this simulation, the transmit symbols go through 7-path Rayleigh fading channel. The proposed detection scheme uses MMSE based detection for robustness to noise effect and higher error performance. For the MMSE, the estimation of the noise power is very simple and it does not cause high complexity to the overall systems. Also, the BER performance for the ML is shown for comparison. The BER performance for the proposed detection scheme is higher than the conventional ZF because the DFE and QRD-M stages are helpful for the accurate estimation of transmits symbols with slight addition of the complex multiplications. Also, the BER performance for the proposed detection scheme is slightly lower than the conventional DFE at low SNR because the noise is amplified in the partial ZF and it causes severe error propagations. However, the gap of the BER performances between the conventional DFE and the proposed detection scheme becomes small, according to increased SNR.

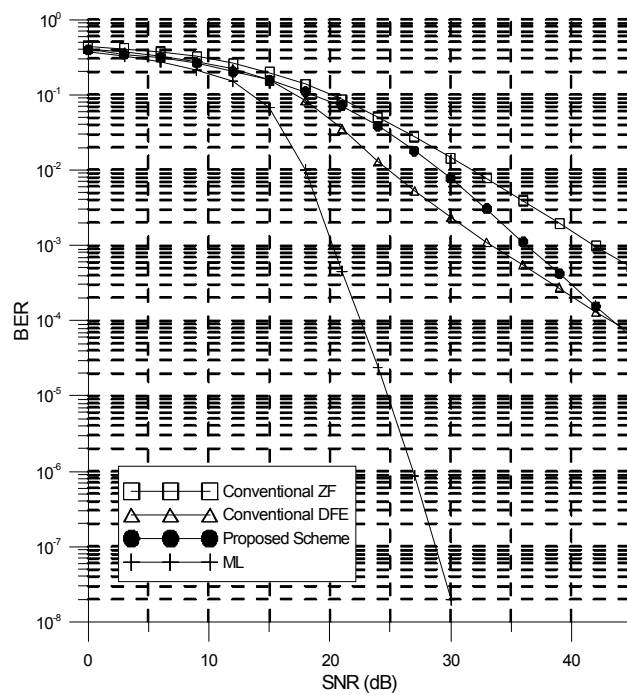


**Figure 4.** The BER performances for the conventional and the proposed detection scheme in  $4 \times 4$  MIMO-OFDM system (BER: bit error rate; ML: maximum likelihood).

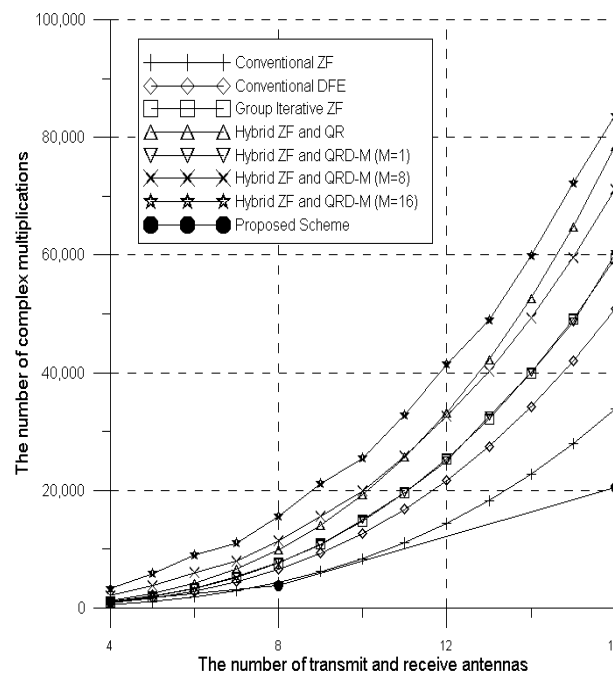
Figure 5 shows the BER performances for the conventional ZF, DFE and proposed detection scheme in  $8 \times 8$  MIMO-OFDM system. For various environments, the transmit symbols go through 10-path Rayleigh fading channel. Also, the BER performance for the ML is shown for comparison. Like in  $4 \times 4$  MIMO-OFDM system, the BER performance for the proposed detection scheme is higher than the conventional ZF, and is lower than the conventional DFE at low SNR. As a result, the error performance for the proposed detection scheme is slightly decreased, according to the increased number of transmit antennas due to severe noise enhancement in the partial ZF stage. However, the



complexity for the proposed scheme is increased almost linearly, according to the increased number of transmit antennas. This result is shown in Figure 6.



**Figure 5.** The bit error rate (BER) performances for the conventional and the proposed detection scheme in  $8 \times 8$  MIMO-OFDM system.



**Figure 6.** The required number of complex multiplications for the conventional and the proposed detection scheme (QRD-M: QR decomposition-M algorithm).

For further comparisons, Tables 2 and 3 compare the error performance and the complexity for the proposed detection scheme with the proposed simplified ML decoding (SMLD) in [3]. The number of transmit and receive antennas is four and eight. Also, the used modulation is 16-QAM. The parameter  $V$

denotes the number of candidate symbols at the top layer in the DFE algorithm. The error performance and complexity increase nonlinearly, according to increased  $V$ . Table 2 represents the required SNR to obtain the BER performance for  $10^{-3}$ . The required SNR for the SMLD is decreased according to the increased number of transmit antennas. However, the required SNR for the proposed detection scheme is increased according to the increased number of transmit antennas because the noise enhancement is very severe in the partial ZF stage. In a similar way, Table 3 represents the required number of complex multiplications. Like Table 2, the number of complex multiplications for the SMLD is variable according to  $V$ . The required number of complex multiplications for the proposed detection scheme increases linearly, according to the increased number of transmit antennas relative to the SMLD due to the partial ZF stage.

**Table 2.** The required signal-to-noise ratio (SNR) to obtain the BER performance for  $10^{-3}$  (BER: bit error rate; QAM: quadrature amplitude modulation; ML: maximum likelihood).

Detection Scheme	$N_t = N_r = 4$ Modulation: 16-QAM			$N_t = N_r = 8$ , 16-QAM Modulation: 16-QAM		
	$V = 1$	$V = 8$	$V = 16$	$V = 1$	$V = 8$	$V = 16$
Low complexity ML	29 dB	25.5 dB	25 dB	27 dB	25 dB	24.5 dB
Proposed scheme	33 dB			36 dB		

**Table 3.** The required number of complex multiplications (SMLD: simplified ML decoding).

Detection Scheme	$N_t = N_r = 4$ Modulation: 16-QAM			$N_t = N_r = 8$ Modulation: 16-QAM		
	$V = 1$	$V = 8$	$V = 16$	$V = 1$	$V = 8$	$V = 16$
SMLD	892	1284	1732	6708	8836	11268
Proposed scheme	1089			3842		

## 6. Conclusions

This paper proposes linearly approximate signal detection scheme in the MIMO-OFDM system. At the first stage, the proposed detection scheme uses the partial ZF to increase the complexity linearly. The partial ZF divides the huge MIMO-OFDM system into  $N/2 \times 2 \times 2$  MIMO-OFDM systems. After the partial ZF, the conventional DFE and QRD- $M$  are used. The conventional DFE and QRD- $M$  have high complexity in the huge MIMO-OFDM system. However, these nonlinear detection schemes are applied to  $2 \times 2$  MIMO-OFDM systems, and it makes the whole complexity linear. The simulation results show that the proposed detection scheme has a higher error performance than the conventional ZF and slightly lower error performance than the conventional DFE. Also, the required number of complex multiplications for the proposed detection scheme is lower than the conventional ZF and the conventional DFE, according to the increased number of transmit antennas due to the partial ZF divides the huge MIMO-OFDM system into several small MIMO-OFDM systems. Therefore, the whole complexity for the proposed algorithm increases linearly, although the number of transmit antennas increases. The proposed algorithm can be widely used in the huge MIMO-OFDM systems that require very low complexity for practical implementations.

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and all simulation results. Also, H.-K.S. provided the experimental materials for better computational simulations and revised critical errors of the manuscript.

**Conflicts of Interest:** The authors declare no conflict of interest.

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