Complexity reduction using QRD-M or SD in MIMO Interleaved SC-FDMA receiver with iterative detection

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Abstract*—The maximum likelihood detection (MLD) is known as the optimum separation method of spatially multiplexed signals in MIMO (Multiple Input Multiple Output) communication schemes. However, when the number of transmit antenna and the level of modulation are increased, the complexity of MLD becomes quite large and the reduction of complexity is the important issue. For reducing the complexity of MLD, QRD-M (QR Decomposition with M-Algorithm) was proposed, but it is a quasi-ML method and could not obtain the ML solution although the complexity is largely reduced. On the other hand, Sphere Decoding (SD) can obtain the ML solution like MLD with reduced complexity. In this paper, we propose the receiver structure for MIMO Interleaved SC-FDMA (MIMO Interleaved Single Carrier-Frequency Division Multiple Access) where the FDE (Frequency Domain Equalization) is firstly done for obtaining the tentative decision results and secondly using the tentative results the ISI (Inter-Symbol Interference) is cancelled by ISI canceller and then the MLD is made for separating the spatially multiplexed signals. Furthermore the output from MLD is fed back to ISI canceller and this feedback is iteratively for improving the BER. In order to reduce the complexity of the proposed receiver, we replace the MLD by QRD-M or SD. By examining the BER characteristics and the complexity reduction through computer simulations, we have verified the effectiveness of proposed receiver in which MLD is replaced by SD.

I. INTRODUCTION

Recently MIMO transmission techniques with multiple transmit and receive antennas are widely used to achieve the spatially multiplexed transmission and to increase the transmission rate in wireless communications. For MIMO spatially multiplexed transmission, MLD is known as the optimum signal separation method at the receiver side. However, MLD needs very high computational complexity and the reduction of complexity is a problem [1]. The SC-FDMA is used as the uplink wireless scheme in LTE (Long Term Evolution). The feature of its low PAPR (Peak to Average Power Ratio) characteristics decreases the burden of nonlinear amplifiers in UE (User Equipment) and the SC-DFMA is more suitable to uplink transmission than OFDM (Orthogonal Frequency Division Multiplexing) [2]. Moreover by employing the interleaved SC-FDMA where the subcarriers for each UE are deployed like a comb tooth, the PAPR of SC-FDMA is further reduced and the frequency diversity effect becomes large. We already proposed the MIMO SC-FDE and MIMO Interleaved SC-FDMA receivers with iterative detection where the receive signal is firstly detected by FDE (Frequency Domain Equalization), i.e., MMSE nulling, to obtain the tentative decision results and secondly the ISI cancellation from the signal to be detected using the tentative decision results is done followed by the MLD for the signal separation of spatially multiplexed signals [3], [4]. However, as the complexity of MLD increases as the power of modulation levels to the number of transmit antenna, the complexity reduction of MLD becomes an important issue. In this paper, we propose the novel receiver in which the MLD is replaced by QRD-M or SD to reduce the complexity of MLD in our previously proposed receiver structure, where QRD-M is the quasi-ML scheme but the SD can obtain the ML solution like MLD. Through computer simulations, we have examined the BER characteristics and the complexity reduction effect of proposed iterative receiver with ISI canceller and MLD replaced by QRD-M or SD in MIMO interleaved SC-FDMA. Consequently we verified the SD improves the complexity very much without degrading the BER.

II. MIMO Interleaved SC-FDMA receiver

A. Proposed transmitter and receiver structure

In Fig.1, the block diagram of transmitter and receiver for the uplink to be considered is shown. At the transmitter of each UE, the QAM (Quadrature Amplitude Modulation) signal is FFT (Fast Fourier Transform) transformed with N-points and converted to the frequency domain. The FFT points are then mapped to the interleaved frequency points like a comb tooth. After that, the frequency points are IFFT (Inverse FFT) converted with M points to obtain the time domain signal. The CP (Cyclic Prefix) is inserted and the signal is transmitted to the channel. At the receiver (Base station), after removing the CP, the FDE, i.e., MMSE nulling, is firstly done. The receive signal is FFT converted with M points and the frequency domain signal is obtained. The converted frequency points are de-mapped to each user subcarriers and the subcarriers are multiplied by the MMSE weight $\boldsymbol{G}_u(n)$ at frequency point n, i.e.

$$\boldsymbol{G}_u(n) = \boldsymbol{H}_u(n)^H \left\{ \boldsymbol{H}_u(n)\boldsymbol{H}_u(n)^H + n_T\sigma^2 \boldsymbol{I}_{n_T} \right\}^{-1} \tag{1}$$

where $\boldsymbol{H}_u(n)$ denotes the MIMO channel matrix at the frequency point n assigned to the user u, σ^2 the variance of noise, $\boldsymbol{I}_{n_T}$ the identity matrix with the size n_T which is the number of transmit antenna. After that, the frequency points are IFFT converted with N points and the time domain signal is obtained to be detected as the hard detected result $\hat{\boldsymbol{x}}_u$.

$$\hat{\boldsymbol{x}}_u = [\hat{\boldsymbol{x}}_u(1), \hat{\boldsymbol{x}}_u(2), \cdots \hat{\boldsymbol{x}}_u(N)] \tag{2}$$

We call $\hat{x}_u$ as the tentative decision through FDE. Using $\hat{\boldsymbol{x}}_u$, the receive signal replica due to ISI caused by the transmit signals other than the signal at time k to be detected, is generated and is subtracted from the receive signal. Using tentative decision of (2) and by letting the transmit signal of user u at time k to be detected being 0, the ISI replica is obtained as (3).

$$\begin{aligned}\hat{\boldsymbol{x}}_{uk} &= \left[\hat{\boldsymbol{x}}_u(1), \cdots, \hat{\boldsymbol{x}}_u(k-1), \boldsymbol{0}, \hat{\boldsymbol{x}}_u(k+1), \cdots, \hat{\boldsymbol{x}}_u(N) \right] \\ &= \begin{bmatrix} \hat{x}_{u1}(1), \cdots, \hat{x}_{u1}(k-1), 0, \hat{x}_{u1}(k+1), \cdots, \hat{x}_{u1}(N) \\ \hat{x}_{u2}(1), \cdots, \hat{x}_{u2}(k-1), 0, \hat{x}_{u2}(k+1), \cdots, \hat{x}_{u2}(N) \\ \vdots \\ \hat{x}_{un_T}(1), \cdots, \hat{x}_{un_T}(k-1), 0, \hat{x}_{un_T}(k+1), \cdots, \hat{x}_{un_T}(N) \end{bmatrix}\end{aligned} \tag{3}$$

Next, $\hat{\boldsymbol{x}}_{uk}$ in (3) is converted to frequency domain signal of $\hat{\boldsymbol{X}}_{uk}$ using FFT with N points. This procedure is done for all the users. The frequency domain ISI replica is made by multiplying $\hat{\boldsymbol{X}}_{uk}(n)$ by the channel matrix $\boldsymbol{H}_u(n)$ and the replica $\boldsymbol{H}_u(n)\hat{\boldsymbol{X}}_{uk}(n)$ is subtracted from the receive signal in frequency domain. Accordingly, the ISI components due to the transmit signals other than time k to be detected are removed. If the ISI cancellation is complete, then the condition where only the transmit signal at time k is transmitted to the receiver is achieved. The ISI cancelled receive signal $\boldsymbol{Z}_{uk}(n)$ in frequency domain is expressed as

$$\begin{aligned}&\boldsymbol{Z}_{uk}(n) = \boldsymbol{Y}_u(n) - \boldsymbol{H}_u(n)\hat{\boldsymbol{X}}_{uk}(n) \\ &= \begin{pmatrix} Y_{u1}(n) \\ Y_{u2}(n) \\ \vdots \\ Y_{un_R}(n) \end{pmatrix} - \begin{pmatrix} H_{u,11}(n) & H_{u,12}(n) & \cdots & H_{u,1n_T}(n) \\ H_{u,21}(n) & H_{u,12}(n) & \ddots & H_{u,2n_T}(n) \\ \vdots & \vdots & \ddots & \vdots \\ H_{u,n_R1}(n) & H_{u,n_R2}(n) & \cdots & H_{u,n_Rn_T}(n) \end{pmatrix} \begin{pmatrix} \hat{X}_{uk,1}(n) \\ \hat{X}_{uk,2}(n) \\ \vdots \\ \hat{X}_{uk,n_T}(n) \end{pmatrix}\end{aligned} \tag{4}$$

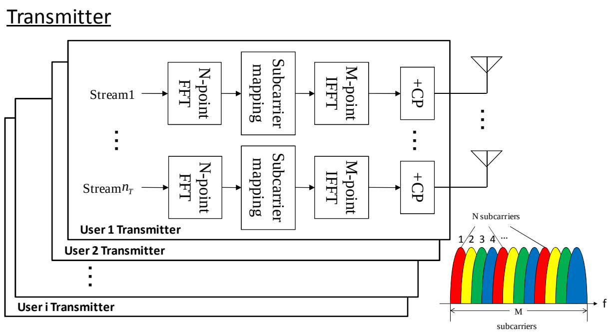
Next, for $\boldsymbol{Z}_{uk}(n)$ in (4), the signal separation of spatially multiplexed transmission is done using MLD. The total number of candidates of receive replica $\boldsymbol{x}'_u$ in MLD is K^{n_T} where K is the level of modulation. The candidate signal for MLD in time domain $\boldsymbol{x}'_{uk}$ is obtained by letting the transmit signal all 0 except for time k to be detected as follows.

$$\begin{aligned}\boldsymbol{x}'_{uk} &= [\ 0, \cdots, 0, \boldsymbol{x}'_u,\ 0, \cdots, 0\ ] \\ &= \begin{bmatrix} 0, \cdots, 0, x'_{u1}, 0, \cdots, 0 \\ 0, \cdots, 0, x'_{u2}, 0, \cdots, 0 \\ \vdots \\ 0, \cdots, 0, x'_{un_T}, 0, \cdots, 0 \end{bmatrix}\end{aligned} \tag{5}$$

where the matrix size is $n_T \times N$. $\boldsymbol{x}'_{uk}$ in (5) is then FFT converted with N points to obtain $\boldsymbol{X}'_{uk}(n)$. $\boldsymbol{X}'_{uk}(n)$ is multiplied by the channel matrix of $\boldsymbol{H}_u(n)$ which is assigned for user u at frequency point n and the candidate receive replica for MLD is obtained as $\boldsymbol{H}_u(n)\boldsymbol{X}'_{uk}(n)$. Then the squared distance between the ISI cancelled receive signal and the candidate MLD replica in frequency domain is calculated as

$$\sum_{n=1}^{N} \left\| \boldsymbol{Z}_{uk}(n) - \boldsymbol{H}_u(n)\boldsymbol{X}'_{uk}(n) \right\|^2 \tag{6}$$

where $\|*\|$ denotes the Euclidian norm. Eq.(6) is minimized over the total K^{n_T} MLD candidates and the MLD output of $\boldsymbol{x}'_{uk}$ which minimizes (6) is obtained. For each user, the same MLD operation is also done. The tentative decision result $\hat{\boldsymbol{x}}_u = \hat{\boldsymbol{x}}_u(k)$ in (2) is then replaced by $\boldsymbol{x}'_{uk}$ and the procedure proceeds from time k to time k+1. This ISI canceller with MLD procedure is repeated from time 1 to N. Accordingly the residual ISI components in tentative FDE decision results are more precisely removed and the spatially multiplexed signals are more accurately separated. After the processing for one FFT block is done, the obtained decision results for one block are again regarded as the tentative decision results. Then the MLD outputs are fed back to the ISI canceller at each FFT block and this feedback is iteratively done to lower the final BER.



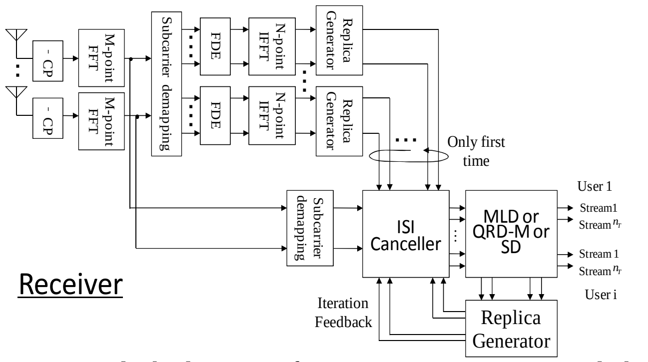


Fig.1 Block diagram of SC-FDMA transmitter and the proposed receiver structure with iterative feedback

B. *Complexity reduction of MLD by QRD-M or SD*

The number of candidate replicas in MLD increases exponentially as K^{n_T}. As the complexity reduction method of MLD, we illustrate the method utilizing the tree search with QR decomposition [1]. The receive signal vector is written as

$$\boldsymbol{y} = \boldsymbol{H}\boldsymbol{s} + \boldsymbol{n} \tag{7}$$

where $\boldsymbol{y}$ is the receive signal vector with $n_R \times 1$, $\boldsymbol{H}$ the channel matrix with $n_R \times n_T$, $\boldsymbol{s}$ the transmit signal vector with $n_T \times 1$ and $\boldsymbol{n}$ the receive noise vector with $n_R \times 1$. Using the QR decomposition, the channel matrix is decomposed into $\boldsymbol{H} = \boldsymbol{Q}\boldsymbol{R}$, where $\boldsymbol{Q}$ is the unitary matrix and $\boldsymbol{R}$ is the upper triangular matrix. By multiplying the Hermitian transpose $\boldsymbol{Q}^H$ by $\boldsymbol{y}$ from the left hand side, we obtain

$$\begin{aligned} \boldsymbol{y} &= \boldsymbol{Q}\boldsymbol{R}\boldsymbol{s} + \boldsymbol{n} \\ \boldsymbol{Q}^H\boldsymbol{y} &= \boldsymbol{R}\boldsymbol{s} + \boldsymbol{Q}^H n \\ \boldsymbol{y}' &= \boldsymbol{R}\boldsymbol{s} + \boldsymbol{n}' \end{aligned} \tag{8}$$

The squared Euclidian norm for MLD is then expressed as

$$\begin{aligned} \|\boldsymbol{y} - \boldsymbol{H}\hat{\boldsymbol{s}}\|^2 &= \|\boldsymbol{y} - \boldsymbol{Q}\boldsymbol{R}\hat{\boldsymbol{s}}\|^2 \\ &= \|\boldsymbol{y}' - \boldsymbol{R}\hat{\boldsymbol{s}}\|^2 \end{aligned} \tag{9}$$

As $\boldsymbol{R}$ is the upper triangular matrix, the detection of transmit signal is considered as the tree search problem from $\hat{s}_{n_T}$. The tree structure is shown in Fig.2 when $K = 2$ (BPSK) and $n_T = 4$, where the diverging number at each node and the depth of tree become $K = 2$ and $n_T = 4$ respectively. Eq.(9) is also expressed in components as

$$\left\| \begin{bmatrix} y_1' \\ y_2' \\ y_3' \\ y_4' \end{bmatrix} - \begin{bmatrix} r_{11} & r_{12} & r_{13} & r_{14} \\ 0 & r_{22} & r_{23} & r_{24} \\ 0 & 0 & r_{33} & r_{34} \\ 0 & 0 & 0 & r_{44} \end{bmatrix} \begin{bmatrix} \hat{s}_1 \\ \hat{s}_2 \\ \hat{s}_3 \\ \hat{s}_4 \end{bmatrix} \right\|^2 \tag{10}$$

As the tree search method for Fig.2 toward the width direction, M algorithm (QRD-M) is widely known. At each step, the squared distance norm for every branch is calculated, and arbitral m survival paths with the least cumulative squared distance metric are retained. The complexity of QRD-M algorithm is constant when m is determined and the QRD-M reduces the complexity of MLD very much, but could not obtain the ML solution, i.e., quasi-LM. On the other hand, the SD algorithm searches the tree of Fig.2 toward the depth direction. The SD first determines the initial sphere radius $C = \|\boldsymbol{y}' - \boldsymbol{R}\hat{\boldsymbol{s}}\|^2$ for some transmit candidate of $\hat{\boldsymbol{s}}$. Next SD searches the transmit signal vector which falls in the radius C toward the depth direction. When the cumulative distance metric exceeds the initial radius, then the subsequent search along the path is no more needed, thus the amount of calculation is saved. Therefore, when the initial sphere radius is small, the complexity reduction becomes effective. In other words, the higher the Eb/N0 is and the smaller the initial sphere radius is, more effectively complexity reduction is done. If the cumulative distance metric does not exceeds the initial radius till the bottom of tree, then the initial radius is replaced by the cumulative metric and the new radius is set. In the same manner the tree search is done for every path in the tree, thus the SD can obtain the ML solution.

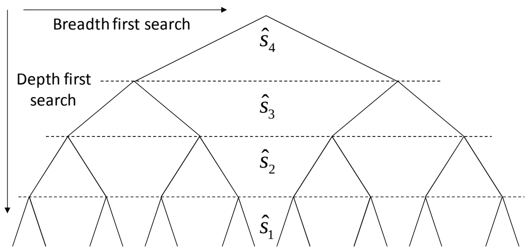


Fig.2 Tree structure of MLD when using QR decomposition

C. *Receiver structure when using QRD-M algorithm*

By using QRD-M instead of MLD in the receiver structure in Fig.1, we reduce the complexity of MLD. The same signal processing procedure mentioned in **2.** is done to cancel the residual ISI and to satisfy the condition as if only the transmit signal at time *k* is transmitted. After the ISI cancellation, the QRD-M is applied instead of MLD. The number of transmit signal candidates $\hat{\boldsymbol{s}}_u$ equals the number of modulation levels of K. Like in (5), the time domain transmit signal vector with N points in which the candidate transmit signal is located at time *k* and the transmit signals at other time instants are all set to 0 is generated. Then the time domain signal vector is converted to the frequency domain signal vector $\hat{\boldsymbol{S}}_u(n)$ using FFT with N points.

Then the channel matrix $\boldsymbol{H}_u(n)$ assigned to user *u* at subcarrier number *n* is QR decomposed.

$$\boldsymbol{H}_u(n) = \boldsymbol{Q}_u(n)\boldsymbol{R}_u(n) \tag{11}$$

The Hermitian transpose $\boldsymbol{Q}_u{}^H(n)$ is multiplied by the output $\boldsymbol{Z}_{uk}(n)$ of the ISI canceller from the left hand side.

$$\boldsymbol{Z}'_{uk}(n) = \boldsymbol{Q}_u^H(n)\boldsymbol{Z}_{uk}(n) \quad (n = 1 \sim N) \tag{12}$$

The squared metric for the minimization using $\boldsymbol{Z}'_{uk}(n)$ in (12) is given by

$$\sum_{n=1}^{N}\left\|\begin{pmatrix} Z'_{uk,1}(n) \\ Z'_{uk,2}(n) \\ \vdots \\ Z'_{uk,n_R}(n) \end{pmatrix} - \begin{pmatrix} R_{u,11}(n) & R_{u,12}(n) & \cdots & R_{u,1n_T}(n) \\ 0 & R_{u,22}(n) & \ddots & R_{u,2n_T}(n) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & R_{u,n_Rn_T}(n) \end{pmatrix}\begin{pmatrix} \hat{S}_{uk,1}(n) \\ \hat{S}_{uk,2}(n) \\ \vdots \\ \hat{S}_{uk,n_T}(n) \end{pmatrix}\right\|^2 \tag{13}$$

Using M algorithm, (13) is step by step calculated from the bottom to the top. The m survival paths with the least cumulative metrics are retained at each step initially from the bottom and the path which minimizes (13) is finally selected from the m survival paths which reach the top. The path obtained by ORD-M determines the output $\hat{\boldsymbol{s}}_u$. The signal processing afterward is the same as MLD.

D. *Receiver structure when using SD algorithm*

By using SD instead of MLD in the receiver structure in Fig.1, we reduce the complexity of MLD. The same signal processing procedure mentioned in **2.** is done to cancel the residual ISI and to satisfy the condition as if only the transmit signal at time *k* is transmitted. In the proposed SD, the initial radius is set using QRD-M. Eq.(13) is used for the search of initial radius using QRD-M The cumulative metric with small radius is firstly searched in the tree using QRD-M and we set this cumulative metric as the initial radius. Next the transmit signal candidate which satisfies the initial radius is searched toward the depth direction in the tree. When the cumulative distance metric exceeds the initial radius, then the subsequent search along the path is no more needed, thus the amount of calculation is saved. If the cumulative distance metric does not exceeds the initial radius till the bottom of tree, then the initial radius is replaced by the cumulative metric and the new radius is set. In the same manner the tree search is done for every path in the tree, thus the SD can obtain the ML solution. Eq.(13) is searched toward the upward direction with exhaustive search to obtain the ML solution of $\hat{\boldsymbol{s}}_u$. If the QRD-M finds the small initial radius, then the efficiency for searching the tree is improved. The iterative feedback afterwards follows the same as the one of MLD.

III. Computer simulation results

We made the computer simulations for Fig.1. The MLD, QRD-M or SD is used after the ISI canceller. The simulation conditions are given in Table.1. The BER characteristics are given in Fig.3 and Fig.5. The total numbers of complex additions and multiplications for calculating the squared norm metric are given in Fig.4 and Fig.6. In Fig.3, we show the lower bound for BER where the ISI canceller works ideally with the perfect knowledge of the transmit signals.

Table I simulation conditions

Number of UE	4
Number of transmit antennas in each UE	4
Number of receive antennas at BS	4
Modulation formats	QPSK
Number of total subcarriers	M=256
Number of subcarriers assigned to each user	N=64
Symbol length of QPSK	T
Guard interval length	T/16
Channel model	Equal power 16 paths quasi-static Rayleigh fading channel
Interval of delay time	T/256
Subcarrier assignment	IFDMA
Channel estimation	Perfect at BS
FDE	Nulling (MMSE)
Initial radius setting for SD	QRD-M (m=1)
Number of iterative feedbacks in the receiver	0, 1, 3 # means the number of MLD, QRD-M or SD

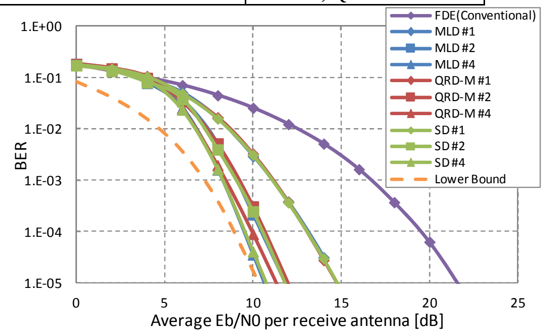


Fig.3 Comparison of BER characteristics

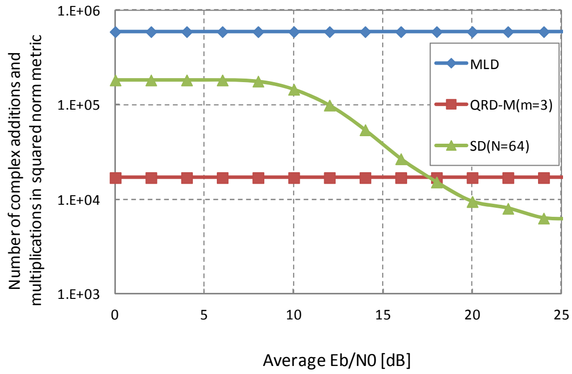


Fig.4 Comparison of total number of complex additions and multiplications in squared norm metric

In Fig.3, compared with the conventional FDE receiver, the proposed receiver using MLD with no iterative feedback improves the BER by about 7 dB at $\text{BER}=10^{-5}$. By increasing the number of iterative feedbacks, the proposed receiver further improves the BER and obtains the BER improvement of more than 10 dB which is close to the lower bound of BER. This is because the MLD outputs with high reliability are used as the improved decision results for making the accurate ISI replicas, and accordingly more exact ISI cancellation becomes possible followed by improved MLD performance too. We know the BER performance of QRD-M is inferior to MLD, but the BER of SD coincides with the MLD, thus the SD can obtain the ML solution. Regarding the number of complex additions and multiplications in squared norm metric in Fig.4, we show the case where there is no iterative feedback, i.e., only for one MLD, QRD-M or SD. Compared with MLD, SD improves the complexity by 1/3 and 1/100 in low Eb/N0 and high Eb/N0 regions respectively. SD also improves the complexity more