WIDELY LINEAR DETECTOR FOR QAM MIMO SYSTEM

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ABSTRACT

Detection algorithm based on widely linear filtering is proposed for the QAM modulated MIMO system. Unlike the existing works which focus on non circular signals (e.g. BPSK, PAM), this paper extends the widely linear detection to MIMO system transmitting circular signals (e.g. QAM, QPSK) by eliminating the I (or Q) component of the transmitted signals. The proposed algorithm has a ML-approaching performance with a computational complexity independent of the modulation order, whose upper bounder is $O(2^{N_T})$ (for QPSK) or $O(3^{N_T})$ (for M-QAM, $M > 4$), lower than the complexity ($O(M^{\sqrt{N_T}})$) of FSD (Fixed-complexity Sphere Decoding) for a practical interval of N_T, the number of transmitting antennas. Simulation results show that the proposed algorithm can achieve quasi-ML performance with complexity comparable with FSD(1, $\cdots$, M) for QPSK signal, and much lower complexity than FSD(1, $\cdots$, M) when the system is 16-QAM (64-QAM) modulated.

1. INTRODUCTION

Widely linear filtering has drawn an increasing interest in estimation [1], beamforming [2], DoA finding [3], and communications [4]. For example, [1] has provided a general scheme for widely linear estimation. Beamformers for the extraction of an unknown signal from non circular interferences are investigated in [2]. [5] gives a new insight into widely linear receivers for the BPSK, MSK, GMSK signals. Blind widely linear structures for multiuser detection of code-division multiple-access signals are proposed in [4]. However, the existing works dealing with widely linear filtering are mainly restricted in the cases where either the signal of interest is non-circular (e.g. BPSK, PAM, OQPSK, OQAM, GMSK MSK) [3]-[4], [5] [6] or the interference is non circular [2] [7]. An important and practical case, where circular signal (e.g. QPSK, QAM, PSK) being desired and interference being circular (e.g. QPSK or QAM modulated MIMO system) or combination of circular and non circular ones, has not been considered by these works. In such cases, widely linear filtering for PAM signal can not be applied directly. In this paper, we introduce a widely linear MIMO (Multiple Input Multiple Output [8]) detector for circular signals in the case of frequency-flat fading channels. The detection of circular signal corrupted by circular noise is discussed. It is shown that the proposed algorithm can achieve quasi-ML performance with an attractive computational complexity compared with FSD (Fixed-complexity Sphere Decoding [9], [10]) which efficiently "fixes" the order of complexity of SD (Sphere Decoding [11]). This paper is distinct from previous works in the following aspects: firstly: widely linear detection of circular signal in the context of circularity is considered, which is different from the previous work; secondly, widely linear filtering is applied by eliminating the I (or Q) component of the transmitted data. The final decision is made by joint detection of the transmitted vector including both I and Q components.

Notation: Upper case letters in boldface are used for matrices. Lower case letters in boldface denote the column vectors. $(\cdot)^H$ denotes Hermitian (conjugate transpose), $(\cdot)^T$ is the operation of transpose. The operation of complex conjugate is denoted by $(\cdot)^*$. $E[\cdot]$ denotes the expectation. $\mathbf{A}(:,k)$ represents the kth column of matrix $\mathbf{A}$. $\mathbf{I}_N$ is the $N \times N$ identity matrix. $\mathbf{0}$ represents zero matrix or vector. $\Re(\cdot)$ and $\Im(\cdot)$ represent the real part and imaginary part of $(\cdot)$ respectively.

2. MIMO COMMUNICATION SYSTEM MODEL

The MIMO system model considered here is a V-BLAST [12] system with N_R antennas at the receiver and N_T antennas at the transmitter, which can be described in equation (1):

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w} \tag{1}$$

where $\mathbf{y} = [y_1, y_2, \cdots, y_{N_R}]^T$ is the received vector. $\mathbf{x} = [x_1, x_2, \cdots, x_{N_T}]^T$ is the transmitted vector, $\mathbf{w} = [w_1, w_2, w_3, \cdots, w_{N_R}]^T$ is the additive white Gaussian noise vector, $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_T}]$ represents the frequency-flat channel. The original signal $\mathbf{x}$ is transmitted, and then distorted by the fading channel and noise, and detected by the receiver. The most famous receiver is the ZF (Zero Forcing) [12] detector which has a low computational complexity but with a limited performance. It is improved by MMSE (Minimum Mean Square Error) [13] [14] receiver which minimizes the mean square error distance between the estimated signal and its original counterpart. Compared to the ML (Maximum likelihood) algorithm which is theoretically optimal but with an unaffordable computational complexity, there is still a huge gap between MMSE and ML.

3. PROPOSED ALGORITHM: WIDELY LINEAR DETECTOR FOR COMPLEX SIGNALS

3.1 Second order statistics of the signal and proposed algorithm

The second order statistical characteristics of signal $\mathbf{y}$ are contained in its correlation matrix $\mathbf{R}_y$ and conjugate correlation matrix $\mathbf{R}_{yc}$, which are respectively defined by: $\mathbf{R}_y = E[\mathbf{y}\mathbf{y}^H]$, $\mathbf{R}_{yc} = E[\mathbf{y}\mathbf{y}^T]$.

In the following sections, the following assumptions are used:

1. $E[\mathbf{x}\mathbf{x}^H] = P\mathbf{I}_{N_T}$, where P is the average transmitting power of each transmitting antenna;

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2. $E[\mathbf{w}\mathbf{w}^H] = \sigma^2 \mathbf{I}_{N_R}$, where σ^2 is the power of noise at each receiving antenna;
3. $E[\mathbf{w}\mathbf{x}^H] = \mathbf{0}$, which means that signal and noise are independent.

In case of circular signal, conjugate correlation matrix $\mathbf{R}_{yc} = \mathbf{0}$, WL-MMSE discussed for SAIC (Single Antenna Interference Cancelation) in [5] is reduced to the conventional MMSE detector. Unfortunately, the circular signals such as QPSK and M-QAM (M is the total number of the constellation points which depends on the modulation scheme, $M > 4$.) signals are widely used in communication systems because of their higher spectral efficiency. For such circular signals, we propose a technique to make them non-circular by eliminating the I (or Q) component of the transmitted signal. One notes that the true signal $\mathbf{x}$ can be estimated by $\widehat{\mathbf{x}}_{M-SIC}$ using MMSE-OSIC [15] as follows:

$$\overline{\mathbf{x}} = Q[\widehat{\mathbf{x}}_{M-SIC}] \tag{2}$$

where $Q[\cdot]$ denotes the operation of quantization appropriate to the constellation in use. We assume

$$\overline{\mathbf{x}} = \mathbf{x} + \mathbf{\Delta} \tag{3}$$

where $\mathbf{\Delta}$ is the error vector of estimation. We define

$$\begin{array}{lll} \mathbf{x} = \mathbf{x}_R + j\mathbf{x}_I & \mathbf{x}_R = \Re(\mathbf{x}) & \mathbf{x}_I = \Im(\mathbf{x}) \\ \overline{\mathbf{x}} = \overline{\mathbf{x}}_R + j\overline{\mathbf{x}}_I & \overline{\mathbf{x}}_R = \Re(\overline{\mathbf{x}}) & \overline{\mathbf{x}}_I = \Im(\overline{\mathbf{x}}) \\ \mathbf{\Delta} = \mathbf{\Delta}_R + j\mathbf{\Delta}_I & \mathbf{\Delta}_R = \Re(\mathbf{\Delta}) & \mathbf{\Delta}_I = \Im(\mathbf{\Delta}) \end{array} \tag{4}$$

In system (1), $\mathbf{x}_I$ can be eliminated from $\mathbf{y}$ as follows:

$$\mathbf{z} = \mathbf{y} - j\mathbf{H}\mathbf{x}_I \tag{5}$$

According to equation (3) and (4), $\mathbf{x}_I = \overline{\mathbf{x}}_I - \mathbf{\Delta}_I$, a non circular system is given by:

$$\mathbf{z} = \mathbf{y} - j\mathbf{H}\overline{\mathbf{x}}_I + j\mathbf{H}\mathbf{\Delta}_I = \mathbf{H}\mathbf{x}_R + \mathbf{w} \tag{6}$$

The real part $\mathbf{x}_R$ can be considered as a PAM signal, and similarly, the following matrix is constructed based on the equation (6):

$$\widetilde{\mathbf{z}} = \widetilde{\mathbf{H}}\mathbf{x}_R + \widetilde{\mathbf{w}} \tag{7}$$

where $\widetilde{\mathbf{z}} = [\mathbf{z}^T \;\; \mathbf{z}^H]^T$. Therefore $\mathbf{x}_R$ is estimated as:

$$\widehat{\mathbf{x}}_{R-E} = \mathbf{C}_{EMMSE}\widetilde{\mathbf{z}} \tag{8}$$

where

$$\mathbf{C}_{EMMSE} = \frac{1}{2}P\widetilde{\mathbf{H}}^H\mathbf{R}^{-1} \tag{9}$$

where

$$\mathbf{R} = E[\widetilde{\mathbf{z}}\widetilde{\mathbf{z}}^H] = \begin{bmatrix} \mathbf{R}_z & \mathbf{R}_{zc} \\ \mathbf{R}_{zc}^* & \mathbf{R}_z^* \end{bmatrix} \tag{10}$$

$$\begin{array}{c} \mathbf{R}_z = E[\mathbf{z}\mathbf{z}^H] = \frac{1}{2}P\mathbf{H}\mathbf{H}^H + \sigma^2\mathbf{I}_{N_R} \\ \mathbf{R}_{zc} = E[\mathbf{z}\mathbf{z}^T] = \frac{1}{2}P\mathbf{H}\mathbf{H}^T \end{array} \tag{11}$$

The signal vector $\mathbf{x}$ can be estimated as

$$\check{\mathbf{x}} = \widehat{\mathbf{x}}_{R-E} + j\check{\mathbf{x}}_I \tag{12}$$

where $\check{\mathbf{x}}_I$ is estimate of $\mathbf{x}_I$ by $\mathbf{x}_I = \overline{\mathbf{x}}_I - \mathbf{\Delta}_I$. The candidate vector $\mathbf{s}$ is obtained by quantization $\mathbf{s} = Q[\check{\mathbf{x}}]$ if $\mathbf{\Delta}_I$ is known.

3.2 Evaluation of $\mathbf{\Delta}_I$

To get an exact expression of equation (6), the main task is the evaluation of $\mathbf{z}$, which requires to estimate the error vector $\mathbf{\Delta}_I$. For the evaluation of $\mathbf{\Delta}_I$, we should consider the following $N_T + 1$ cases theoretically:
Case 1: The totally correct detection of the imaginary part $\mathbf{x}_I$, then $\mathbf{\Delta}_I = \mathbf{0}$.
Case 2: Only one element, $\mathbf{x}_I(k_1)$, of $\mathbf{x}_I$ is wrongly estimated, we assume $\overline{\mathbf{x}}_I(k_1) - \mathbf{x}_I(k_1) = m_1 d$, d is the distance between two nearest neighbor points in the plane of the constellation.
Case3, $\cdots$, case $N_T + 1$ are corresponding to the situations where 2, $\cdots$, N_T elements are wrongly estimated, respectively. According to different cases, $\mathbf{\Delta}_I$ can be expressed mathematically as follows:

$$\begin{cases} 1: \; \mathbf{\Delta}_I = \mathbf{0} \text{ with probability } p_0 \\ 2: \; \mathbf{\Delta}_I = \mathbf{P}^{k_1-1}[m_1 d, 0, \cdots, 0]^T \text{ with probability} \\ \qquad p_1,\; k_1 \in \{1,\; 2,\; \cdots,\; N_T\},\; k_1 \text{ denotes} \\ \qquad \text{the position index of the error, } m_1 \text{ is non} \\ \qquad \text{zero integer.} \\ 3: \; \mathbf{\Delta}_I = \mathbf{P}^{k_1-1}[m_1 d, 0, \cdots 0]^T + \mathbf{P}^{k_2-1}[m_2 d, 0, \cdots 0]^T \\ \qquad \text{with probability } p_2,\; m_1, m_2 \text{ are non zero} \\ \qquad \text{integers},\; k_1, k_2 \in \{1,\; 2,\; \cdots,\; N_T\}, \\ \qquad k_1 \neq k_2, \text{ denote the position indexes of the} \\ \qquad \text{two errors} \\ \qquad \cdots\cdots\cdots\cdots \\ N_T + 1: \mathbf{\Delta}_I = (\sum_{i=1}^{N_T} \mathbf{P}^{k_i-1}[m_i d, 0, \cdots, 0]^T), \\ \qquad \text{with probability } p_{N_T} \\ \qquad k_1, \ldots, k_{N_T} \in \{1,\; 2,\; \cdots,\; N_T\} \\ \qquad k_1 \neq k_2 \neq \cdots \neq k_{N_T},\; m_i \text{ is non zero integer,} \\ \qquad \text{denote the position indexes of the } N_T \text{ errors} \end{cases} \tag{13}$$

where $p_0 \gg p_1 > p_2 > \cdots > p_{N_T}$ and $\sum_{i=0}^{N_T} p_i = 1$ hold in a wide interval of SNR value. We define $\mathbf{P}^0 = \mathbf{I}$, $\mathbf{P}$ is a $N_T \times N_T$ matrix given by

$$\mathbf{P} = \begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ 1 & 0 & \cdots & 0 & 0 \\ \vdots & 1 & \ddots & \vdots & \vdots \\ 0 & 0 & \ddots & 0 & 0 \\ 0 & 0 & \cdots & 1 & 0 \end{bmatrix}$$

Candidate vectors corresponding to cases stated above can be calculated by (12). Finally, likelihood test is performed among the obtained candidate vectors to determine the one which minimizes $\|\mathbf{y} - \mathbf{H}\mathbf{s}\|^2$. In practical implementation, we just need to consider the first L ($L \leq N_T + 1$) cases which occur with much greater probabilities. For a specific modulation scheme, the number of the candidate vectors can be sharply reduced by the simplification of $\mathbf{\Delta}_I$.

3.3 M-QAM signals

Theoretically, m_i has multiple choices to determine $\mathbf{\Delta}_I$ if $\mathbf{x}$ is M-QAM modulated ($M > 4$) signal, however, the wrong decision occurs with the greatest probability between two nearest neighbor points in the plane of constellation. As a result,

in equation (13), we choose $m_i = \pm 1$ to simplify the consideration, m_i has only one value 1 or -1 for the marginal point of the constellation plane. Equation (13) can be simplified as follows:

$$\begin{cases} 1: \ \mathbf{\Delta}_I = \mathbf{0} \\ 2: \ \mathbf{\Delta}_I = [l_1 d, \ l_2 d, \ \cdots, \ l_{N_T} d]^T \\ \qquad l_i \in \{-1, \ 0, \ 1\}, \sum_{i=1}^{N_T} |l_i| = 1 \\ 3: \ \mathbf{\Delta}_I = [l_1 d, \ l_2 d, \ \cdots, \ l_{N_T} d]^T \\ \qquad l_i \in \{-1, \ 0, \ 1\}, \sum_{i=1}^{N_T} |l_i| = 2 \\ \qquad \cdots\cdots\cdots\cdots \\ N_T+1: \ \mathbf{\Delta}_I = [l_1 d, \ l_2 d, \ \cdots, \ l_{N_T} d]^T \\ \qquad l_i \in \{-1, \ 0, \ 1\}, \sum_{i=1}^{N_T} |l_i| = N_T \end{cases} \tag{14}$$

There are at most $2^{n-1} C_{N_T}^{n-1}$ candidate vectors for case n, where C_k^l denotes the number of l-combinations of an k-element set. We have at most $\sum_{n=1}^{L} (2^{n-1} C_{N_T}^{n-1})$ candidate vectors if the first L cases in equation (14) are considered.

3.4 QPSK signal (4-QAM)

Every element of $\mathbf{x}_I$ has only two possible values ($\frac{d}{2}$ and $\frac{-d}{2}$) if $\mathbf{x}$ is QPSK modulated signal since every point in QPSK constellation is marginal. We can turn to its contrary polarity if the estimate $\overline{\mathbf{x}}_I(k)$ is not correct. Therefore, $\mathbf{\Delta}_I$ is expressed as follows:

$$\begin{cases} 1: \ \mathbf{\Delta}_I = \mathbf{0} \\ 2: \ \mathbf{\Delta}_I = 2[l_1 \overline{\mathbf{x}}_I(1), \ l_2 \overline{\mathbf{x}}_I(2), \ \cdots, \ l_{N_T} \overline{\mathbf{x}}_I(N_T)]^T \\ \qquad l_i \in \{0, \ 1\}, \sum_{i=1}^{N_T} l_i = 1 \\ 3: \ \mathbf{\Delta}_I = 2[l_1 \overline{\mathbf{x}}_I(1), \ l_2 \overline{\mathbf{x}}_I(2), \ \cdots, \ l_{N_T} \overline{\mathbf{x}}_I(N_T)]^T \\ \qquad l_i \in \{0, \ 1\}, \sum_{i=1}^{N_T} l_i = 2 \\ \qquad \cdots\cdots\cdots\cdots \\ N_T+1: \ \mathbf{\Delta}_I = 2[l_1 \overline{\mathbf{x}}_I(1), \ l_2 \overline{\mathbf{x}}_I(2), \ \cdots, \ l_{N_T} \overline{\mathbf{x}}_I(N_T)]^T \\ \qquad l_i \in \{0, \ 1\}, \sum_{i=1}^{N_T} l_i = N_T \end{cases} \tag{15}$$

One should note that there are $C_{N_T}^{n-1}$ candidate vectors for case n. Since $p_0 \gg p_1 > p_2 > \cdots > p_{N_T}$, significant improvement can be achieved by considering only the first two $(L = 2)$ or three $(L = 3)$ cases, which correspond to $C_{N_T}^0 + C_{N_T}^1$ candidate vectors or $C_{N_T}^0 + C_{N_T}^1 + C_{N_T}^2$ candidate vectors. We have $\sum_{n=1}^{L} (C_{N_T}^{n-1})$ candidate vectors if the first L cases in equation (15) are considered.

3.5 Computation of candidate vectors

The candidate vectors are given by formula (12), however, it is time consuming to repeat the matrix multiplication in formula (8). It is necessary to optimize the structure of formula (8). Combining formula (8) and (6), we have:

$$\widehat{\mathbf{x}}_{R-E} = \mathbf{m} + \mathbf{M}\mathbf{\Delta}_I \tag{16}$$

where $\mathbf{M} = \mathbf{C}_{EMMSE}[j\mathbf{H}^T \ \ -j\mathbf{H}^H]^T$, $\mathbf{m} = \mathbf{C}_{EMMSE}[(\mathbf{y} - j\mathbf{H}\overline{\mathbf{x}}_I)^T \ \ (\mathbf{y}^* + j\mathbf{H}^*\overline{\mathbf{x}}_I)^T]^T$. For the M-QAM $(M > 4)$ signals, $\mathbf{\Delta}_I$ has the general form:

$$\mathbf{\Delta}_I = [l_1 d, \ l_2 d, \ \cdots, \ l_{N_T} d]^T \tag{17}$$

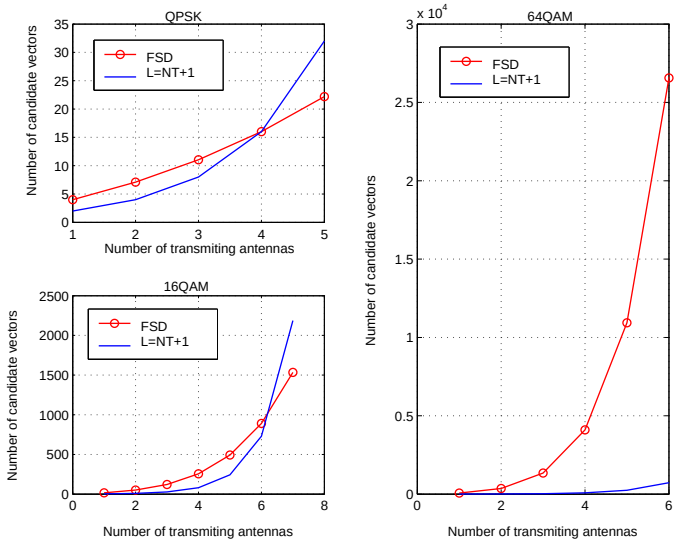

Figure 1: Comparison of complexity order between FSD $(M^{\sqrt{N_T}})$ and proposed algorithm (2^{N_T})

where $l_i \in \{1, \ -1, \ 0\}$. Equation (16) can be written as follows:

$$\widehat{\mathbf{x}}_{R-E} = \mathbf{m} + d \sum_{i=1}^{N_T} [\mathbf{M}(:, i) l_i] \tag{18}$$

In case of QPSK signal, $\mathbf{\Delta}_I$ can be expressed as follows:

$$\mathbf{\Delta}_I = [2 l_1 \overline{\mathbf{x}}_I(1), \ 2 l_2 \overline{\mathbf{x}}_I(2), \ \cdots, \ 2 l_{N_T} \overline{\mathbf{x}}_I(N_T)]^T \tag{19}$$

$$\widehat{\mathbf{x}}_{R-E} = \mathbf{m} + \sum_{i=1}^{N_T} 2[\mathbf{M}(:, i) l_i \overline{\mathbf{x}}_I(i)] \tag{20}$$

where $l_i \in \{1, \ 0\}$. The candidate vectors can be calculated according to equation (12). Thus time-consuming matrix multiplication in formula (8) is replaced by simple addition. Efficiency of computation is improved.

4. COMPLEXITY ANALYSIS

The complexity of the proposed algorithm mainly lies in the computation of matrix inverse at each layer and the likelihood test among the candidate vectors. To compare its complexity with FSD, we can only consider the number of candidate vectors. Because FSD should also calculate the matrix inverse at each level to determine its detection order. As discussed in section 3.3 and 3.4, $L = N_T + 1$ if all cases are considered. We have $\sum_{n=1}^{N_T+1} (2^{n-1} C_{N_T}^{n-1}) = 3^{N_T}$ candidate vectors for the detection of M-QAM signal (M>4), and $\sum_{n=1}^{N_T+1} (C_{N_T}^{n-1}) = 2^{N_T}$ for the detection of QPSK signal. The upper bound of the complexity order is $O(3^{N_T})$ and $O(2^{N_T})$, for the detector of M-QAM signal and QPSK signal, respectively. It is independent of the scheme of modulation, and is totally determined by transmitting antenna number N_T. One notes that lower bound on the average complexity of sphere decoding has been shown to be exponential [16]. Theoretically, FSD algorithm should maintain at the order of

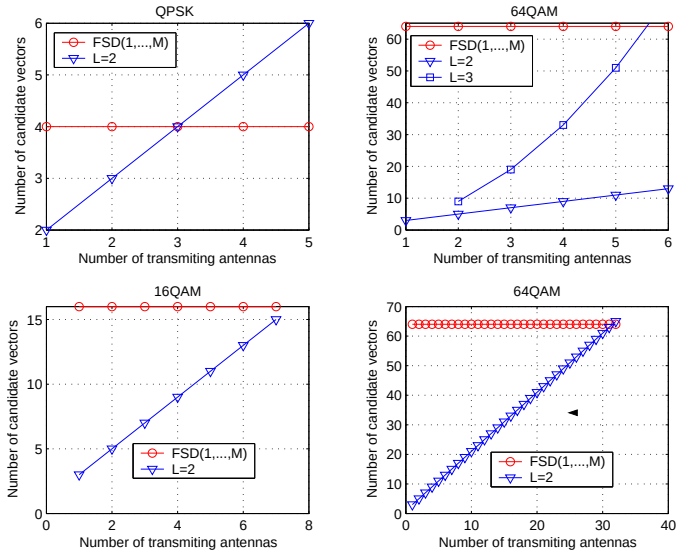

Figure 2: Comparison of complexity between FSD $(1,\cdots,M)$ and proposed algorithm $(L=2)$

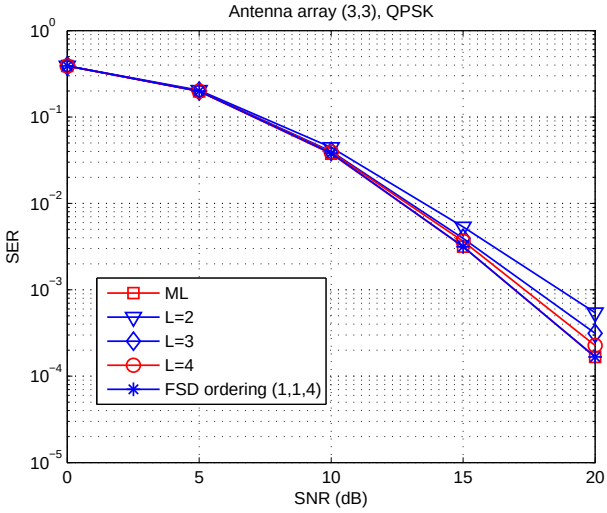

Figure 3: Performance comparison (QPSK, $N_T = N_R = 3$)

$O(M^{\sqrt{N_T}})$ to achieve near ML detection [10]. Fig. 1 compares the complexity between FSD and the proposed algorithm with $L = N_T + 1$. Fig. 1 shows that the complexity of the proposed algorithm is lower than FSD when the number of transmitting antennas N_T is located in quite an applicable interval, more accurately, $N_T \in [1,4]$ for QPSK signal and $N_T \in [1,6]$ for 16-QAM signal. For 64-QAM signal, the proposed algorithm always has much lower complexity in the whole applicable interval of transmitting antenna number.

Practically, FSD is implemented as FSD(1, $\cdots$, M) for simplicity, which means FSD(1, $\cdots$, M) has M candidates vectors. The proposed algorithm $(L=2)$ has $1+N_T$ candidate vectors for QPSK, and $1+2N_T$ for M-QAM. It is easy to verify that the proposed algorithm $(L=2)$ still has less candidate vectors than FSD(1, $\cdots$, M) when $N_T \leq 7$ (or $N_T \leq 31$, respectively) for 16-QAM signal (or 64-QAM, respectively), as shown in Fig. 2.

5. SIMULATION RESULTS

The SER (Symbol Error Rate) performance of the proposed algorithm has been examined by means of the Monte-Carlo simulation. The simulated MIMO systems are V-Blast systems. In addition, SNR is defined as $SNR = PN_T/\sigma^2$ and, in the figures of this paper, SNR is represented by dB. Fig. 3 is obtained based on a QPSK 3×3 system. It shows that there are little distinctions among the performance curves including ML, FSD(1, $\cdots$, M) and the proposed algorithm ($L=2$, $L=3$ and $L=4$). Fig. 4 simulates a 3×4 16-QAM system, we can observe that the SER curves are almost overlapped by one another. Although the better SER performance can be achieved with larger L theoretically, different values of L make only a little difference in performance, especially between $L = N_T + 1$ and $L = N_T$. However, one should note that $L = N_T$ has 2^{N_T} less candidates vectors than $L = N_T + 1$ for 16-QAM or 64-QAM. Similar situation is showed by Fig. 5 which is simulated based on a 64-QAM 4×5 system. For $L \in [1, N_T+1]$, $L=2$ (or 3) is a preferred choice, because the proposed algorithm with $L=2$ (or $L=3$ for 64QAM) can achieve much less complexity than FSD(1, $\cdots$, M) with a little performance degradation.

6. CONCLUSION

Efficient detection algorithm of flexible computational complexity for QAM MIMO system is proposed. The proposed algorithm extends widely linear filtering to the circular signals (e.g. QAM, QPSK). It is shown that the proposed algorithm ($L=2$ or $L=3$) has a ML-approaching performance, and that its complexity is comparable with FSD$(1,\cdots,M)$ for QPSK signal, and lower than the corresponding FSD$(1,\cdots,M)$ algorithm for 16-QAM or 64-QAM signal in quite an applicable interval of transmitting antenna number. The application of the proposed algorithm is not confined to QPSK or QAM signals, it is also effective in the situation where other circular signals (e.g. PSK) are employed.

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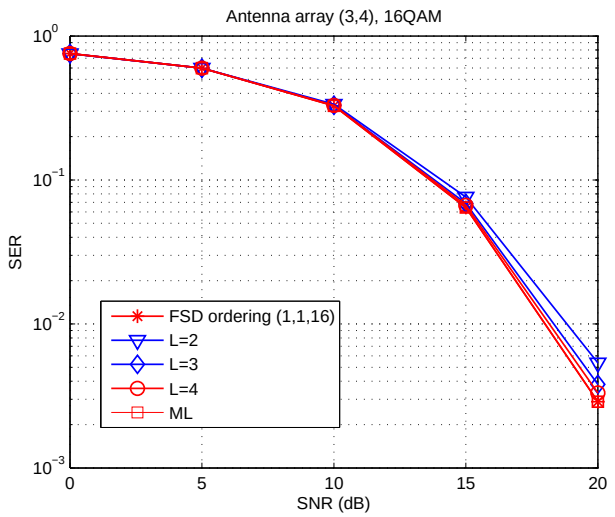

Figure 4: Performance comparison (16-QAM, $N_T = 3, N_R = 4$)

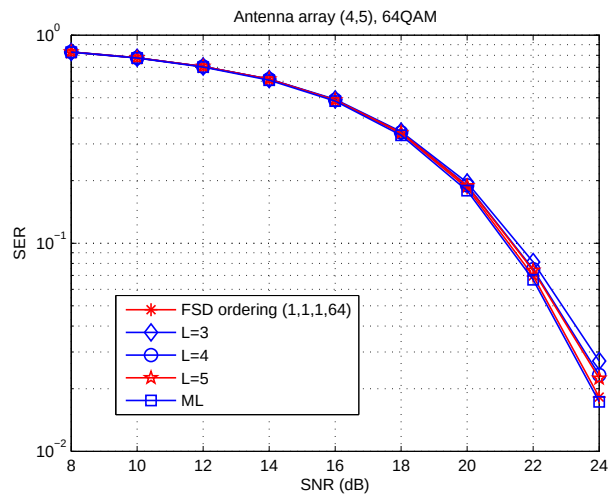

Figure 5: Performance comparison (64-QAM, $N_T = 4, N_R = 5$)

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