Enhanced Groupwise Detection with a New Receive Combiner for Spatial Multiplexing MIMO Systems

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***Abstract*—In this letter, we propose a new groupwise receive combiner design for multiple-input multiple-output spatial multiplexing systems. The conventional group detection (GD) suffers from a considerable performance loss since the noise components are not taken into account. The output signal-to-interference-plus-noise ratio (SINR) is defined in each subgroup in order to consider both the desired signal and noise statistics. Adopting the real-valued representation, we provide an optimal receive combiner which maximizes the SINR with a general group size. The simulation results show that the proposed scheme achieves a large performance gain over the conventional GD in coded systems. Also, when combining with near-optimal detection algorithms such as sphere decoder, the proposed GD scheme offers a comparable performance with significant reduced complexity.**

***Index Terms*—Multiple-input multiple-output (MIMO), group detection, spatial multiplexing, maximum-likelihood (ML) detection.**

I. INTRODUCTION

IN recent years, the demands for high data rate wireless communication have increased rapidly for wireless multimedia services. Multiple-input multiple-output (MIMO) systems are popular approaches to meet such demands [1][2]. Especially, spatial multiplexing (SM) methods enable extremely high spectral efficiency by transmitting independent streams of data simultaneously through multiple transmit antennas [2][3]. Several receiver algorithms have been proposed to improve the detection performance for SM systems. One of the simplest algorithms is a linear processing based on zero-forcing (ZF) or minimum mean square error (MMSE) criteria. While maximum likelihood detection (MLD) serves as the optimal solution, the complexity of the MLD increases exponentially with the number of transmit antennas. The vertical Bell Labs layered space-time (V-BLAST) architecture [4] is one of well-known detection methods for MIMO systems with manageable complexity. However, the performance of the V-BLAST scheme based on ordered successive interference cancelation (OSIC) detection is limited by error propagation [3].

To further enhance the performance of the V-BLAST, group detection (GD) schemes [5][6] can be employed with a good compromise between performance and complexity. The GD scheme partitions transmitted signals into several subgroups and employs the MLD for each subgroup. Compared to the full MLD, the GD approach enables us to reduce the complexity with a small performance loss. In [7], a different approach to the GD is presented for closed-loop MIMO systems, where the orthogonalized spatial multiplexing [8] is applied at the transmitter. Also, a recent work in [9] has analyzed successive group decoders and their optimal partitioning for uplink multiuser MIMO channels.

One major drawback of the conventional GD [5] is that it suffers from a considerable performance loss because the noise is ignored. In order to overcome a noise enhancement issue, we propose an enhanced groupwise detection with a new receiver combiner design method based on the signal-to-interference-plus-noise ratio (SINR) maximization. In [10], an approach of maximizing the signal-to-leakage-and-noise ratio (SLNR) was proposed for designing transmit beamforming matrices in multi-user MIMO systems. In contrast, in this letter, we define the received SINR in each subgroup to design an effective GD scheme for open-loop point-to-point MIMO systems. Particularly, we employ the real-valued system representation in designing the groupwise linear combiner. By considering the real and imaginary parts separately, we can take into account the interference on the component level of symbols, and thus we can achieve a performance gain compared to the complex-valued model. The simulation results verify that the proposed scheme achieves a large performance gain over the conventional GD in coded systems. Also, by adjusting the group size properly, our scheme becomes single-symbol decodable while outperforming the linear MMSE receiver. Furthermore, we show that the proposed scheme offers a good performance gain with much reduced complexity when combined with near-optimal detection algorithms such as sphere decoder [11]. In addition, a simple group selection method is presented which improves the performance with a small increase in computational complexity.

Throughout this letter, the following notations are used. Normal letters represent scalar quantities, bold face letters indicate vectors, and boldface uppercase letters designate matrices. With a bar accounting for complex variables, for any complex notation $\overline{c}$, we denote the real and imaginary part of $\overline{c}$ by $\Re[\overline{c}]$ and $\Im[\overline{c}]$, respectively. The superscript $(\cdot)^T$ denotes the

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transpose of a vector or a matrix. The determinant and trace of a matrix are denoted by $\det(\cdot)$ and $\mathrm{Tr}(\cdot)$, respectively, and $\mathbf{I}_d$ indicates an identity matrix of size d.

II. System Description

In this section, we consider an SM scheme for open-loop MIMO systems with N_t transmit and N_r receive antennas. We define a complex transmitted symbol vector as $\overline{\mathbf{x}} = [\overline{x}_1 \cdots \overline{x}_{N_t}]^T$. Each element of $\overline{\mathbf{x}}$ is chosen from a QAM signal set $\mathcal{M}_c$ of size M_c and $\overline{x}_i$ is given as $\overline{x}_i = x_{i,I} + jx_{i,Q}$ where $j = \sqrt{-1}$. The receiver is assumed to have perfect knowledge of the channel. This work can be easily applied to frequency selective channels where orthogonal frequency division multiplexing (OFDM) modulation is used to convert the broadband MIMO channel into multiple narrowband MIMO channels [12].

The N_r dimensional complex received signal vector $\overline{\mathbf{y}} = [\overline{y}_1 \cdots \overline{y}_{N_r}]^T$ is given as

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

where the MIMO channel matrix $\overline{\mathbf{H}} \in \mathbb{C}^{N_r \times N_t}$ is represented by

$$\overline{\mathbf{H}} = \begin{bmatrix} \overline{h}_{1,1} & \cdots & \overline{h}_{1,N_t} \\ \vdots & \ddots & \vdots \\ \overline{h}_{N_r,1} & \cdots & \overline{h}_{N_r,N_t} \end{bmatrix},$$

and $\overline{\mathbf{n}} \in \mathbb{C}^{N_r \times 1}$ denotes the additive white Gaussian noise vector with zero mean and the covariance matrix $\sigma_n^2 \mathbf{I}_{N_r}$. Here we assume that $\mathbb{E}[\overline{\mathbf{x}}\overline{\mathbf{x}}^H] = \sigma_s^2 \mathbf{I}_{N_t}$. The channel coefficients $\overline{h}_{j,i}$ of $\overline{\mathbf{H}}$ represent the channel response between the i-th transmit and the j-th receive antenna, and have an independent and identically distributed (i.i.d.) complex Gaussian distribution.

III. Proposed Groupwise Detection

In this section, we propose a new groupwise linear combiner and the corresponding group selection method. The concept of the GD was originally suggested for multi-user detection in code-division multiple access [6] and applied to MIMO systems to reduce the complexity of the MLD [5]. We first introduce the equivalent real-valued channel model which enables us to consider the interference in the symbol components. The real-valued representation of the system (1) can be written as

$$\mathbf{y} = \begin{bmatrix} \Re[\overline{\mathbf{y}}] \\ \Im[\overline{\mathbf{y}}] \end{bmatrix} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{2}$$

where $\mathbf{x} \in \mathbb{R}^{2N_t \times 1}$ and $\mathbf{n} \in \mathbb{R}^{2N_r \times 1}$ are respectively given by $\mathbf{x} = \left[\Re[\overline{\mathbf{x}}^T]\ \Im[\overline{\mathbf{x}}^T]\right]^T = [x_{1,I} \cdots x_{N_t,I}\ x_{1,Q} \cdots x_{N_t,Q}]^T$, $\mathbf{n} = \left[\Re[\overline{\mathbf{n}}^T]\ \Im[\overline{\mathbf{n}}^T]\right]^T$, and $\mathbf{H} \in \mathbb{R}^{2N_r \times 2N_t}$ denotes the real-valued representation of the channel matrix $\overline{\mathbf{H}}$ as

$$\mathbf{H} = \begin{bmatrix} \Re[\overline{\mathbf{H}}] & -\Im[\overline{\mathbf{H}}] \\ \Im[\overline{\mathbf{H}}] & \Re[\overline{\mathbf{H}}] \end{bmatrix}.$$

For an M_c-ary QAM modulation system, the real-valued symbols $x_{i,I}$ and $x_{i,Q}$ are chosen from a $\sqrt{M_c}$-PAM signal set $\mathcal{M}_r$. Although we consider QAM in this letter, we note that the system can experience diversity order improvement if $\overline{\mathbf{x}}$ is PAM modulated [13].

A. Groupwise receive combiner design

Now we design the receive filter $\mathbf{G}$ which separates the transmitted signals into K independent subgroups. Although the proposed GD can generate subgroups with different sizes, we focus on the case where all subgroups have the same size $L\,(\geq 2)$ and subgroups are not overlapped with each other, i.e., $2N_t = KL$. Then, $\mathbf{H}$ is represented by

$$\mathbf{H} = [\mathbf{H}_1 \ \ \mathbf{H}_2 \ \ \cdots \ \ \mathbf{H}_K]$$

where $\mathbf{H}_k$ denotes the $2N_r \times L$ channel matrix for the k-th subgroup $(1 \leq k \leq K)$.

First, we define the receive combining matrix $\mathbf{G}$ of size $2N_r \times 2N_t$ as

$$\mathbf{G} = [\mathbf{G}_1 \ \ \mathbf{G}_2 \ \ \cdots \ \ \mathbf{G}_K]$$

where $\mathbf{G}_k$ is the $2N_r \times L$ combining matrix for the k-th subgroup. By applying the linear filter $\mathbf{G}$ to the real-valued system model in (2), the combiner output vector $\mathbf{z}$ becomes

$$\mathbf{z} = \mathbf{G}^T\mathbf{y} = \mathbf{H}_G\mathbf{x} + \mathbf{G}^T\mathbf{n}$$

where the effective channel $\mathbf{H}_G$ is

$$\mathbf{H}_G = \mathbf{G}^T\mathbf{H} = \begin{bmatrix} \mathbf{G}_1^T\mathbf{H}_1 & \mathbf{G}_1^T\mathbf{H}_2 & \cdots & \mathbf{G}_1^T\mathbf{H}_K \\ \mathbf{G}_2^T\mathbf{H}_1 & \mathbf{G}_2^T\mathbf{H}_2 & \cdots & \mathbf{G}_2^T\mathbf{H}_K \\ \vdots & & \ddots & \vdots \\ \mathbf{G}_K^T\mathbf{H}_1 & \mathbf{G}_K^T\mathbf{H}_2 & \cdots & \mathbf{G}_K^T\mathbf{H}_K \end{bmatrix}.$$

Here, $\mathbf{G}_k^T\mathbf{H}_k$ is the k-th effective channel submatrix of size $L \times L$ and $\mathbf{G}_k^T\mathbf{H}_j$ accounts for residual interference from the j-th subgroup $(j \neq k)$ to the k-th subgroup.

Our goal is to design $\mathbf{G}_k$ which minimizes the residual interference while taking the noise component into account. We can obtain the submatrices independently for each desired subgroup. Denoting $\mathbf{x}_k \in \mathbb{R}^{L \times 1}$ as the real-valued transmit signal vector for the k-th subgroup, the groupwise linear combiner output $\mathbf{z}_k \in \mathbb{R}^{L \times 1}$ at the k-th subgroup is given as

$$\begin{aligned} \mathbf{z}_k &= \mathbf{G}_k^T\mathbf{H}_k\mathbf{x}_k + \mathbf{G}_k^T \sum_{j=1,\ j\neq k}^{K} \mathbf{H}_j\mathbf{x}_j + \mathbf{G}_k^T\mathbf{n} \\ &= \mathbf{G}_k^T\mathbf{H}_k\mathbf{x}_k + \mathbf{G}_k^T\widetilde{\mathbf{H}}_k\widetilde{\mathbf{x}}_k + \mathbf{G}_k^T\mathbf{n} \end{aligned} \tag{3}$$

where $\widetilde{\mathbf{H}}_k = [\mathbf{H}_1 \cdots \mathbf{H}_{k-1}\ \mathbf{H}_{k+1} \cdots \mathbf{H}_K]$ and $\widetilde{\mathbf{x}}_k = \left[\mathbf{x}_1^T \cdots \mathbf{x}_{k-1}^T\ \mathbf{x}_{k+1}^T \cdots \mathbf{x}_K^T\right]^T$.

In equation (3), we define the combiner output SINR at the k-th subgroup as

$$\begin{aligned} \mathrm{SINR}_k &\triangleq \frac{\mathbb{E}\left(\|\mathbf{G}_k^T\mathbf{H}_k\mathbf{x}_k\|^2\right)}{\mathbb{E}\left(\|\mathbf{G}_k^T\widetilde{\mathbf{H}}_k\widetilde{\mathbf{x}}_k\|^2\right) + \mathbb{E}\left(\|\mathbf{G}_k^T\mathbf{n}\|^2\right)} \\ &= \frac{\sigma_s^2/2\ \mathrm{Tr}\left(\mathbf{H}_k^T\mathbf{G}_k\mathbf{G}_k^T\mathbf{H}_k\right)}{\sigma_s^2/2\ \mathrm{Tr}\left(\widetilde{\mathbf{H}}_k^T\mathbf{G}_k\mathbf{G}_k^T\widetilde{\mathbf{H}}_k\right) + \sigma_n^2/2\ \mathrm{Tr}\left(\mathbf{G}_k\mathbf{G}_k^T\right)} \\ &= \frac{\sigma_s^2\mathrm{Tr}\left(\mathbf{G}_k^T\mathbf{H}_k\mathbf{H}_k^T\mathbf{G}_k\right)}{\sigma_s^2\mathrm{Tr}\left(\mathbf{G}_k^T\widetilde{\mathbf{H}}_k\widetilde{\mathbf{H}}_k^T\mathbf{G}_k\right) + \sigma_n^2\mathrm{Tr}\left(\mathbf{G}_k^T\mathbf{G}_k\right)} \\ &= \frac{\mathrm{Tr}\left(\mathbf{G}_k^T\mathbf{H}_k\mathbf{H}_k^T\mathbf{G}_k\right)}{\mathrm{Tr}\left[\mathbf{G}_k^T\left(\widetilde{\mathbf{H}}_k\widetilde{\mathbf{H}}_k^T + \sigma_n^2/\sigma_s^2\mathbf{I}_{2N_r}\right)\mathbf{G}_k\right]}. \end{aligned} \tag{4}$$

Now, we need to select $\mathbf{G}_k$ for $k = 1, \cdots, K$ which maximizes the k-th subgroup SINR of (4) as

$$\arg\max_{\{\mathbf{G}_k\}} \frac{\mathrm{Tr}\left(\mathbf{G}_k^T \mathbf{A}_k \mathbf{G}_k\right)}{\mathrm{Tr}\left(\mathbf{G}_k^T \mathbf{B}_k \mathbf{G}_k\right)}$$

where $\mathbf{A}_k = \mathbf{H}_k \mathbf{H}_k^T$ and $\mathbf{B}_k = \widetilde{\mathbf{H}}_k \widetilde{\mathbf{H}}_k^T + \sigma_n^2/\sigma_s^2 \mathbf{I}_{2N_r}$. Note that the $2N_r \times 2N_r$ matrices $\mathbf{A}_k$ and $\mathbf{B}_k$ are real symmetric. Also $\mathbf{A}_k$ and $\mathbf{B}_k$ are positive semi-definite and positive definite, respectively.

From the generalized eigenvalue problem $\mathbf{A}_k \mathbf{x} = \lambda \mathbf{B}_k \mathbf{x}$ [14], the matrices $\mathbf{A}_k$ and $\mathbf{B}_k$ can be simultaneously diagonalized as

$$\begin{array}{rcl} \mathbf{X}_k^T \mathbf{B}_k \mathbf{X}_k & = & \mathbf{I}_{2N_r} \\ \mathbf{X}_k^T \mathbf{A}_k \mathbf{X}_k & = & \mathrm{diag}\left(\lambda_1, \cdots, \lambda_{2N_r}\right) \end{array} \quad (5)$$

where $\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_{2N_r}$ are non-negative generalized eigenvalues and the columns of $\mathbf{X}_k$ are the corresponding generalized eigenvectors. Then, the optimal $\mathbf{G}_k$ for the above maximization problem corresponds to $\mathbf{X}_k^{1:L}$, the first L columns of $\mathbf{X}_k$ in (5) [10]. Thus, $\mathbf{G}_k$ can be obtained using the Cholesky factorization and solving a standard eigenvalue problem.

Denoting the Cholesky factorization of $\mathbf{B}_k$ as $\mathbf{B}_k = \mathbf{R}_k^T \mathbf{R}_k$, the solution $\widehat{\mathbf{G}}_k$ is given by

$$\widehat{\mathbf{G}}_k = \mathbf{X}_k^{1:L} = \mathbf{R}_k^{-1} \mathbf{Q}_k^{1:L} \quad (6)$$

where $\mathbf{Q}_k^{1:L}$ consists of the eigenvectors corresponding to the L largest eigenvalues of $\mathbf{C}_k = \mathbf{R}_k^{-T} \mathbf{A}_k \mathbf{R}_k^{-1}$ [14]. From equations (4) and (5), the corresponding SINR for the k-th subgroup is then determined by

$$\mathrm{SINR}_k = \frac{1}{L} \sum_{j=1}^{L} \lambda_{k,j}$$

where $\lambda_{k,j}$ denotes the j-th largest eigenvalue of $\mathbf{C}_k$. Also, (5) implies $\widehat{\mathbf{G}}_k^T \mathbf{B}_k \widehat{\mathbf{G}}_k = \mathbf{I}_L$, which means that $\widehat{\mathbf{G}}_k$ whitens the noise plus interference signal from other groups.

Now, the MLD is performed in each of K independent subgroups to jointly detect L real symbols. The number of search candidates for the proposed scheme reduces to $(\sqrt{M_c})^L = M_c^{L/2}$ in comparison to $M_c^{N_t}$ in the original MLD. Therefore, we can effectively decrease the detection complexity by adjusting the group size L appropriately. Note that when L is set to 2, the proposed scheme becomes single complex-symbol ML decodable.

B. Capacity of the proposed scheme

In this subsection, we study the maximum information rate which can be achieved from the proposed GD. In general GD systems with group size L, the capacity of the k-th decoding group when combined with $\mathbf{G}_k$ can be expressed by

$$C_{\mathrm{GD},k} = \frac{1}{2} \log\det\left(\mathbf{I}_L + \left(\mathbf{G}_k^T \mathbf{B}_k \mathbf{G}_k\right)^{-1} \mathbf{G}_k^T \mathbf{A}_k \mathbf{G}_k\right). \quad (7)$$

Then, by substituting (6) into (7), we have

$$\begin{aligned} C_{\mathrm{GD},k} &= \frac{1}{2} \log\det\left(\mathbf{I}_L + \mathbf{H}_k^T \mathbf{R}_k^{-1} \mathbf{Q}_k^{1:L} (\mathbf{Q}_k^{1:L})^T \mathbf{R}_k^{-T} \mathbf{H}_k\right) \\ &= \frac{1}{2} \log\det\left(\mathbf{I}_L + \mathbf{H}_k^T \mathbf{B}_k^{-1} \mathbf{H}_k\right) \end{aligned} \quad (8)$$

where $\mathbf{Q}_k^{1:L} (\mathbf{Q}_k^{1:L})^T$ can be removed from the equation by using the definition of $\mathbf{Q}_k^{1:L}$. Note that (8) is the maximum mutual information for the k-th group when the interference from other groups is regarded as Gaussian noise. Therefore, we confirm that the proposed receiver (6) which maximizes the group SINR also achieves the maximum capacity in each decoding group.

The total capacity that the proposed GD can achieve is then given by

$$C_{\mathrm{GD}} = \sum_{k=1}^{K} C_{\mathrm{GD},k} = \frac{1}{2} \sum_{k=1}^{K} \log\det\left(\mathbf{I}_L + \mathbf{H}_k^T \mathbf{B}_k^{-1} \mathbf{H}_k\right). \quad (9)$$

Here, we notice that the receive filter $\mathbf{G}$ which achieves (9) is not unique. Namely, there exists an alternate form solution for $\mathbf{G}_k$ which connects the equations (7) and (8), although it does not satisfy the conditions (5) for our generalized eigenvalue problem. One such example is given by the MMSE based group decoder $\mathbf{G}_k = \mathbf{B}_k^{-1} \mathbf{H}_k (\mathbf{H}_k^T \mathbf{B}_k^{-1} \mathbf{H}_k)^{-1/2}$. Note that this equivalent filter may make computations simpler.

C. Group selection method

In order to further enhance the performance of the proposed scheme, a group selection (GS) can be considered. In [5], an adaptive GS method was proposed based on the correlation matrix of $\mathbf{R} = \mathbf{H}^H \mathbf{H}$. However, this algorithm suffers from high complexity cost as it allows an overlapped choice, and the channel correlation alone does not account for the detection performance much. In this letter, we present a GS criterion which maximizes the sum achievable rate C_{GD} given in (9). By rearranging $2N_t$ columns of $\mathbf{H}$ before combining, we can choose the best combination of K subgroups among N total choices which guarantees

$$\begin{aligned} \hat{n} &= \arg\max_{1 \leq n \leq N} C_{\mathrm{GD}}(n) \\ &= \arg\max_{1 \leq n \leq N} \sum_{k=1}^{K} \sum_{j=1}^{L} \log\left(1 + \lambda_{k,j}(n)\right). \end{aligned} \quad (10)$$

Here, an integer number N indicates the number of all possible group combinations computed as $N = \binom{2N_t}{L}\binom{2N_t - L}{L} \cdots \binom{L}{L} / K!$. Note that N is small for practical configurations. From the criterion (10), the receiver determines the best pairing set which maximizes the performance.

IV. Simulation Results

In this section, we present simulation results for the proposed groupwise detection in flat fading channels and compare them with conventional schemes. We assume that the channel is fixed during one frame and varies independently from frame to frame. Also, we employ a 64-state convolutional encoder with polynomials (133, 171) in octal notation with code rates $R_c = 1/2$ and $3/4$, where the rate $3/4$ code is obtained by puncturing. In order to avoid complex LLR calculations of the MLD, we adopt list sphere decoder (LSD) [11] in each subgroup as a near optimal algorithm.

Figures 1 and 2 describe the simulation results for $N_r = N_t = 4$. The group size L of the conventional and the proposed GD is set to 4, and the GS is not considered here. From Figure 1, we observe that the proposed GD provides

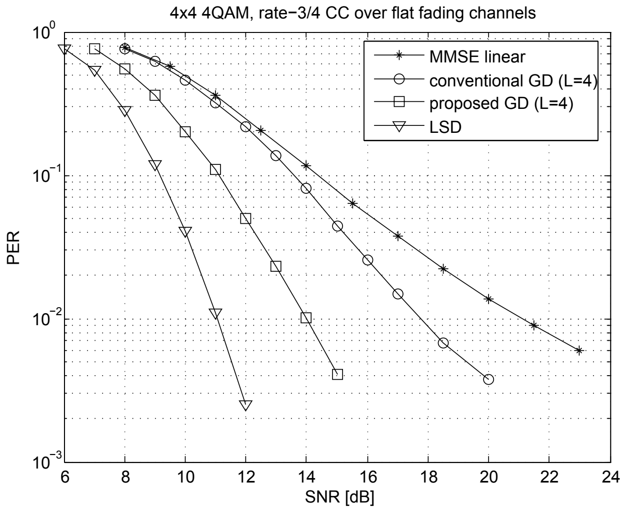

Fig. 1. PER performance comparison of the proposed method with other schemes for $N_r = N_t = 4$ over 4QAM.

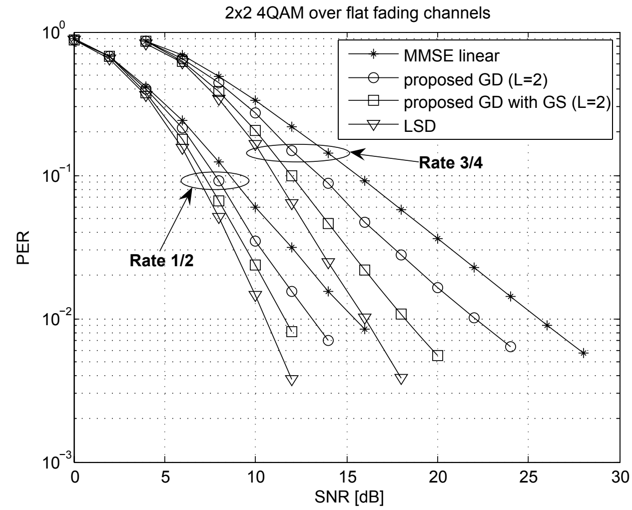

Fig. 3. PER performance comparison for $N_r = N_t = 2$ over 4QAM in the case of $L = 2$.

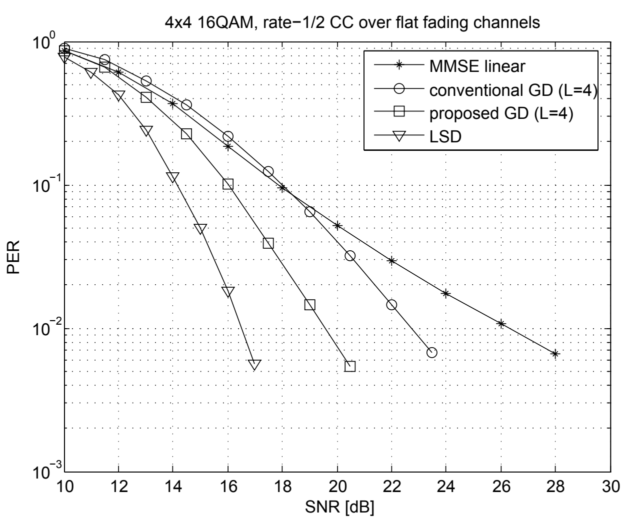

Fig. 2. PER performance comparison of the proposed method with other schemes for $N_r = N_t = 4$ over 16QAM.

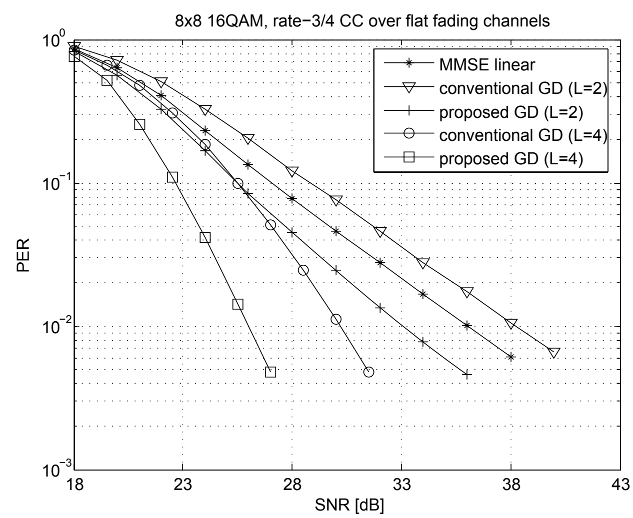

Fig. 4. PER performance comparison for $N_r = N_t = 8$ over 16QAM.

a 4dB gain at a packet error rate (PER) of 10^{-2} over the conventional GD [5] for 4QAM with $R_c = 3/4$. The conventional single-group LSD performs best, however, whose complexity is significantly high with $N_t = 4$ as will be noted in Table I. Figure 2 illustrates the PER comparison toward 16QAM modulation and $R_c = 1/2$, where the same tendency is observed between the performance of the proposed GD and existing techniques.

Next, we consider a MIMO channel with $N_r = N_t = 2$ for 4QAM, and the GS is applied. In this case, the proposed GD becomes single complex-symbol decodable with $L = 2$ and thus the detection complexity is proportional only to $\mathcal{O}(M_c)$. Figure 3 shows that the proposed method with grouping provides gains of 2dB and 5.5dB over the MMSE linear receiver at a rate $1/2$ and rate $3/4$, respectively. Here the total number of grouping N is only 3, and thus the searching complexity for the best grouping in (10) is minimal.

In Figure 4, we plot the results for larger antenna arrays of $N_r = N_t = 8$, which is currently discussed as a primary setting in the 3GPP LTE−Advanced next-generation standard [15]. We apply 16QAM and $R_c = 3/4$. The curves show that with the group size 2 and 4, the proposed GD achieves 3dB and 10dB gains, respectively, over the baseline MMSE receiver with increased complexity. Also, our scheme outperforms the conventional GD by 5dB for both $L = 2$ and 4.

Finally, we address the computational complexity of the proposed scheme. The proposed GD with GS requires NK times filter updates and K times subgroup LLR computations. For updating $\mathbf{G}_k$, $\frac{1}{3}N_r(2N_r + 1)(6N_t + 4N_r + 12L - 5) + 4N_r(N_r + 1)LN_{\text{itr}}$ floating-point multiplications are required in total including Cholesky factorization, back-substitution and power method[1] [17]. Here, N_{itr} indicates the number of iterations for the power method and determined from 2 to 6 in a practical range of antenna dimensions. Table I shows that

[1]The power method computes only the dominant singular components [16], which is suitable for finding $\mathbf{Q}_k^{1:L}$.

TABLE I
NUMBER OF MULTIPLICATIONS IN CODED SYSTEMS AT A PACKET ERROR RATE OF 10^{-2} (K=2 AND N=1 FOR PROPOSED GD)

	Conventional LSD	Proposed GD with LSD
$N_r = N_t = 2$, 4QAM	1,994	500
$N_r = N_t = 4$, 4QAM	64,988	7,790
$N_r = N_t = 4$, 16QAM	80,668	8,332

the proposed GD combined with the groupwise LSD saves about 75% and 90% of the number of real multiplications for $N_r = N_t = 2$ and 4, respectively, compared to the full size LSD at the expense of a small performance loss.

V. CONCLUSIONS

The GD technique exhibits a good tradeoff between complexity and performance in the design of signal detectors for SM systems. The main objective of this work is to design an enhanced groupwise receiver for the GD. By considering the noise statistics and adopting the real-valued representation, the proposed GD scheme maximizes the received SINR defined in each subgroup with a general group size. By maintaining a moderate group size, we can obtain reasonable performance and practical complexity. Also, we present a simple GS method based on the SINR. The simulation results confirm that the proposed method achieves a considerable performance gain compared to the conventional GD and other existing SM schemes in coded systems.

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