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A Grouped Pre-Coding Aided Spatial Modulation for MIMO Systems

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ABSTRACT Pre-coding aided spatial modulation (PSM) is a recently introduced concept that achieves low complexity and low cost at the receiver. However, it is only suitable for the symmetric or under-determined system due to the utilization of the linear pre-coding. In this paper, a novel PSM-based transmission strategy termed as grouped pre-coding aided spatial modulation (group PSM) is proposed for the over-determined MIMO system where the number of receive antennas is larger than the number of transmit antennas. In our group PSM scheme, the receive antennas are equally divided into several groups. A single receive antenna group is selected during each time slot and the index of the receive antenna group is used for conveying extra information bits. We design a low complexity detection method for the proposed scheme and derive the bit error rate expression. To further improve the performance of group PSM, we propose a power allocation method, where the power allocated for each receive antenna group is optimized based on maximizing the minimum Euclidean distance between received group PSM signal constellations. We also generalize our group PSM scheme in order to improve the multiplexing gain and flexibility, where a particular subset of receive antenna groups is activated so as to convey more information bits. Our simulation results validate the accuracy of the theoretical analysis and demonstrate that both the proposed group PSM and generalized group PSM schemes achieve good performance and significantly increase the scalability of PSM in practice.

INDEX TERMS Multiple-input-multiple-output (MIMO), pre-coding aided spatial modulation, power allocation, bit error rate (BER).

I. INTRODUCTION

Spatial modulation (SM) has been considered as one promising paradigm for emerging multiple-antenna techniques [1]–[6], where only a single transmit antenna is activated in each time slot. Except for the conventional modulated symbol, in the SM system, the transmit antenna index is also employed to convey additional information. Recent studies show that the SM system outperforms the conventional multiple-input multiple-output (MIMO) systems in terms of energy efficiency and computational burden [7].

In contrast to existing SM-based schemes, the pre-coding aided spatial modulation (PSM) was proposed in [8], [9] as a new technique for 5G massive MIMO downlink transmission [10]. Unlike the SM technique, PSM invokes the conventional MIMO linear pre-coding technique in conjunction with SM to exploit the receive spatial domain. Specifically, PSM activates all transmit antennas to focus the transmit power only towards a single receive antenna by utilizing the linear pre-coding method such as zero-forcing (ZF). Thus, the receiver can easily demodulate transmitted signals without channel state information (CSI) at the receiver [8], which decreases the complexity and cost of the receiver. The mathematical framework for computing the upper bound of the average bit error rate (BER) of the PSM system based on the minimum mean square error (MMSE) pre-coding was provided in [11].

In [9], a generalized pre-coding aided spatial modulation (GPSM) was presented to increase the flexibility of PSM by activating a specific subset of receive antennas during each time slot. In [12], the closed-form BER of GPSM and conventional PSM systems were derived, and the performance of GPSM in comparison to the conventional MIMO linear pre-coding scheme was discussed. In order to accommodate the time-varying channel, a pair of power allocation algorithms were proposed in [10] for the PSM system by minimizing the upper bound of the instantaneous bit error probability. In [13], the authors proposed a novel unified power allocation algorithm for PSM systems based on the maximum minimum Euclidean distance. Moreover, PSM also

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has been studied in many communication scenarios. The conventional full spatial multiplexing was combined with PSM in [14] to improve bandwidth efficiency. In [15], the PSM was combined with simultaneous wireless information and power transfer (SWIPT) to achieve better flexibility for the optimization of the rate-energy region. Besides, two beamforming methods were proposed for the PSM-SWIPT system in [16], where only the index of the receive antenna is used for conveying information. The antenna selection strategy for secure PSM systems was studied in [17]. The PSM scheme was utilised for V2X communication systems in [18] by applying the lattice reduction principle. Both in PSM and GPSM schemes, all transmit antennas are activated for emitting the pre-coded information while all receive antennas are used for the data mapping. In [19], a joint transmitter-receiver spatial modulation (JSM) was introduced to obtain both the multiplexing gain and the diversity gain at the transmitter and the receiver, respectively. In order to further improve the reliability of the JSM scheme, the power allocation algorithm and the antenna selection algorithm were provided in [20]. However, the conventional PSM scheme assumes that the number of transmit antennas is larger than that of receive antennas in order to facilitate the linear pre-coding process. On the other hand, the receive antenna subset selection algorithms were proposed for PSM systems in over-determined scenarios to improve the system performance [21], [22]. Specifically, a fast receive antenna subset selection criterion was proposed for PSM based on a fast greedy incremental algorithm [21]. To further improve the BER performance, [22] presented two efficient receive antennas selection methods by using the condition number and eigenvalue of the channel matrix and the Wishart distribution. However, the receive antenna subset selection based PSM leads to substantial idle system resources when the number of receive antennas is significantly lager than that of transmit antennas.

Against the background, this paper considers the over-determined MIMO system, e.g., the uplink massive MIMO system. We aims to employ the index of all receive antennas to convey information bits so as to achieve higher data rate transmissions. The main contributions of this paper are given as follows.

1) We propose a novel PSM-based scheme termed as grouped pre-coding aided spatial modulation (group PSM) for MIMO systems. In the proposed scheme, the receive antennas are equally divided into several groups for performing a small dimension of PSM. Therefore, the information bits are carried by three units: the index of the selected receive antenna group, the index of activated antenna of the selected group and the conventional modulated symbol.
2) Based on the proposed group PSM system, we present a low complexity detection method and derive its BER expression.
3) In order to further improve the BER performance, a new power allocation method is proposed for the group PSM scheme based on minimizing the maximum Euclidean distance between the received symbols.
4) We generalize the proposed group PSM in order to improve the multiplexing gain and flexibility of group PSM. In the generalized group PSM scheme, several receive antenna groups are selected in each time slot to carry more data symbols.

The remaining of this paper is organized as follows. In section II, the system model of the proposed group PSM is introduced and a low complexity detection method is proposed. Besides, the system performance is analyzed. Section III presents a power allocation method for group PSM scheme. The group PSM is generalized in Section IV and the simulation experiments are carried out in Section V. Finally, conclusions are drawn in Section VI.

Notations: In this paper, the operators $\mathrm{Tr}(\cdot)$, $(\cdot)^H$ and $(\cdot)^{-1}$ represent the trace, conjugate transpose and inverse of a matrix, respectively. Besides, we use $\|\cdot\|$ and $\lfloor\cdot\rfloor$ to represent the Euclidean norm operation and the floor operation, respectively. The operator $\mathbb{E}_{\mathbf{H}}[\cdot]$ denotes the statistical expectation with respect to the fading channel $\mathbf{H}$, $\Re$ represents taking the real part, and $p(\cdot)$ denotes the probability of an event. In addition, $\mathbf{I}_N$ denotes an identity matrix of N-dimension and $\binom{\cdot}{\cdot}$ denotes the binomial coefficient.

II. SYSTEM MODEL

We consider a MIMO system consisting of N_t transmit antennas and N_r receive antennas. The signal model at the receiver side can be described as

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

where $\mathbf{x}$ is the resultant transmit signal after pre-coding, $\mathbf{H}$ is the frequency flat Rayleigh fading channel matrix whose entries are independently and identically distributed (i.i.d.) random variables with zero mean and unit variance, $\mathbf{n}$ is the additive white Gaussian noise (AWGN) vector, and its entries are i.i.d. Gaussian noise with zero mean and variance σ_n^2. We assume that CSI is perfectly known at the transceiver for executing the linear pre-coding operation at the transmitter and the decoding operation at the receiver.

A. PRE-CODING AIDED SPATIAL MODULATION

The conventional PSM system selects one receive antenna according to part of information bits. And then the transmitter activates all transmit antennas to emit the pre-coded signal, in which the linear pre-coding method is adopted to enable only the selected receive antenna to obtain the transmitted symbol.

More specifically, in PSM, the first part of information bits are mapped to the modulated symbol, and the extra information bits are used for activating one out of N_r receive antennas (Here, we use "active" and "inactive" to represent whether the receive antenna receives the modulated symbol). Based on above mapping rules, the super-symbol $\mathbf{z} \in \mathbb{C}^{N_r \times 1}$

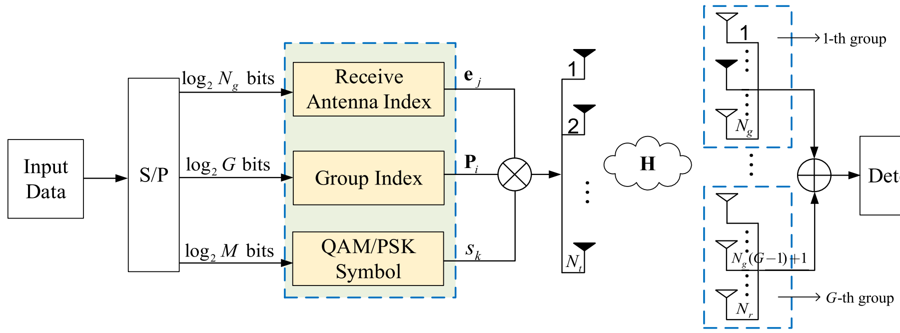

FIGURE 1. Schematic model of the proposed group PSM system.

of the PSM scheme can be written as

$$\mathbf{z} = [0, \ldots, 0, \underset{\substack{\uparrow \\ r\text{-th}}}{s_k}, 0, \ldots, 0]^T \tag{2}$$

where r denotes the index of the active receive antenna, $s_k \in \mathcal{S}$, $k \in \{1, 2, \ldots, M\}$, is the k-th modulated symbol which satisfies $E[|s_k|^2] = 1$, and $\mathcal{S}$ denotes an M-ary phase-shift keying (PSK) or quadrature-amplitude modulation (QAM) constellation set. After pre-coding and transmission, the received signal $\mathbf{y} \in \mathbb{C}^{N_r \times 1}$ of PSM can be expressed as

$$\begin{aligned} \mathbf{y} &= \mathbf{Hx} + \mathbf{n} \\ &= \mathbf{HPz} + \mathbf{n}. \end{aligned} \tag{3}$$

where $\mathbf{P} = \beta \mathbf{H}^H \left(\mathbf{H}\mathbf{H}^H\right)^{-1}$ is the ZF pre-coding matrix and $\beta = \sqrt{N_r / \mathrm{Tr}[\left(\mathbf{H}\mathbf{H}^H\right)^{-1}]}$ is the normalized factor. When the MMSE pre-coding method is used, we have $\mathbf{P} = \beta \left(\mathbf{H}^H\mathbf{H} + N_r\sigma_n^2\mathbf{I}_{N_t}\right)^{-1}\mathbf{H}^H$, where $\beta = \sqrt{N_r / \mathrm{Tr}\left(\left(\mathbf{H}^H\mathbf{H} + N_r\sigma_n^2\mathbf{I}_{N_t}\right)^{-2}\mathbf{H}^H\mathbf{H}\right)}$. At the receiver, the optimal maximum likelihood (ML) detector can be given by [9]

$$\langle \hat{r}, \hat{k} \rangle = \underset{\mathbf{z} \in \mathcal{Z}}{\arg\min} \|\mathbf{y}/\beta - \mathbf{z}\|^2, \tag{4}$$

where $\mathcal{Z}$ represents the alphabet of super-symbol $\mathbf{z}$.

A basic restriction of the conventional PSM system is that the number of transmit antennas needs to be equal or greater than the number of receive antennas since the ZF pre-coding matrix is obtained from the right inverse of $\mathbf{H}$. That is, the ZF pre-coding matrix only exists when $N_t \geq N_r$ [9]. Thus, the total number of bits transmitted by the conventional PSM system can be expressed as

$$\eta_{\mathrm{PSM}} = \log_2 [\min(N_t, N_r)] + \log_2 M, \tag{5}$$

where N_t and N_r are the integer powers of two.

B. THE PROPOSED SCHEME

1) TRANSMITTER

The system model of the proposed group PSM scheme is depicted in Fig. 1, where $N_r \geq N_t$ and N_r is the integer powers of two. We divide receive antennas into G equal size groups, and each group contains N_g receive antennas, i.e., we have $G = N_r/N_g$. In this paper, we take $N_g \leq N_t$ to facilitate the linear pre-coding operation.

In the proposed scheme, one out of G receive antenna groups is chosen according to a part of input data during each time slot. In other words, the index of the selected receive antenna group is used for conveying additional information. Then, we operate a small dimension of PSM over the link between the transmit antennas and receive antennas of the selected group. Hence, we can notice that the PSM scheme is a special case of the proposed scheme if $N_g = N_r$.

More specifically, the input bit stream is split into three parts. The first portion, which is of length $\log_2 G$, is used to select a receive antenna group. The second portion, which is of length $\log_2 N_g$, is used to activate one receive antenna out of N_g receive antennas. And the last portion with $\log_2 M$ bits is mapped to a conventional M-ary modulated symbol.

Thus, the total number of bits transmitted by the proposed group PSM scheme can be given by

$$\begin{aligned} \eta_{\text{group PSM}} &= \log_2 G + \log_2 N_g + \log_2 M \\ &= \log_2 N_r + \log_2 M. \end{aligned} \tag{6}$$

Notice from (6) that the transmit rate of our group PSM scheme is higher than that of the PSM scheme.

We let $i \in \{1, 2, \ldots, G\}$ be the index of the selected receive antenna group and $j \in \left\{1, 2, \ldots, N_g\right\}$ be the index of the active receive antenna of the selected group. The transmitter generates G pre-coding matrixes for the links between transmit antennas and receive antenna groups. Then, the resultant signal after the pre-coding operation can be expressed as

$$\mathbf{x} = \mathbf{P}_i \mathbf{e}_j s_k, \tag{7}$$

where the spatial domain symbol $\mathbf{e}_j = [0, \ldots, 0, 1, 0, \ldots, 0]^T$ is a $N_g \times 1$ vector with all of its elements are zero except the j-th element which is equal to one, and $\mathbf{P}_i \in \mathbb{C}^{N_t \times N_g}$ is the linear matrix corresponding to the i-th receive antenna group. When the ZF pre-coding method is used, we can have

$$\mathbf{P}_i = \beta_i \mathbf{H}_i^H \left(\mathbf{H}_i \mathbf{H}_i^H\right)^{-1}, \tag{8}$$

$$\beta_i = \sqrt{\frac{N_g}{\mathrm{Tr}\left[\left(\mathbf{H}_i \mathbf{H}_i^H\right)^{-1}\right]}}, \tag{9}$$

where $\mathbf{H}_i$ represents the channel matrix from the transmitter to the i-th receive antenna group, β_i denotes the normalisation factor [4].

2) RECEIVER

In our group PSM scheme, the signal observed at N_r receive antennas can be expressed as

$$\mathbf{y} = \mathbf{H}\mathbf{P}_i \mathbf{e}_j s_k + \mathbf{n}. \tag{10}$$

Based on the ML criterion, the joint detection of the selected antenna group, the active receive antenna and the conventional modulated symbol can be formulated as

$$\begin{aligned} \langle \hat{i}, \hat{j}, \hat{k} \rangle &= \arg\min_{i,j,k} \|\mathbf{y} - \mathbf{H}\mathbf{P}_i \mathbf{e}_j s_k\|^2 \\ &= \arg\min_{i,j,k} |s_k|^2 \|\mathbf{H}\mathbf{p}_{i,j}\|^2 - 2\Re\{\mathbf{y}^H \mathbf{H}\mathbf{p}_{i,j} s_k\}, \end{aligned} \tag{11}$$

where $\mathbf{p}_{i,j}$ represents the j-th column of the pre-coding matrix $\mathbf{P}_i$. The optimal ML detector, however, imposes prohibitive computational complexity when the number of antennas is large.

To tackle this problem, we now introduce a sub-optimal low complexity detection method based on the structure characteristic of the received group PSM signal. By defining $\mathbf{y} = [\mathbf{y}^1, \ldots, \mathbf{y}^G]^T$ and following the principle of ZF pre-coding, the received signal in (10) can be rewritten as

$$\begin{cases} \mathbf{y}^i = \beta_i \mathbf{e}_j s_k + \mathbf{n}_i \\ \mathbf{y}^{\tilde{i}} = \mathbf{H}_{\tilde{i}} \mathbf{P}_i \mathbf{e}_j s_k + \mathbf{n}_{\tilde{i}}, \forall\, \tilde{i} \neq i, \end{cases} \tag{12}$$

where $\mathbf{y}^i \in \mathbb{C}^{N_g \times 1}$ ($\mathbf{y}^{\tilde{i}} \in \mathbb{C}^{N_g \times 1}$) and $\mathbf{n}_i$ ($\mathbf{n}_{\tilde{i}}$) represent the observation vector and the noise vector at the i-th ($\tilde{i}$-th) receive antenna group, respectively. Unlike the conventional PSM scheme where the transmitter beamforms the modulated signal to only one out of N_r receive antennas, in our scheme, the observations at the receive antennas of unselected groups also contain the modulated symbol. Moreover, the observation of the i-th receive antenna group in (12) can be individually expressed as

$$\begin{cases} y_j^i = \beta_i s_k + n_{i,j} \\ y_{j'}^i = n_{i,j'}, j' = 1, \ldots, N_g \text{ and } j' \neq j, \end{cases} \tag{13}$$

From (12) and (13) we can find that for the selected receive antenna group, the transmitter is capable of focusing the transmit power to a single receive antenna. It is because only one pre-coding matrix is chosen at the transmitter. In other words, only one receive antenna can observe the efficient modulated symbol in the selected receive antenna group.

When $\mathbf{P}_i$ is employed, the observed signal at the i-th group of receive antennas is given in (13). Based on the structure characteristics of (13), we can reduce the detection complexity via splitting the receiving signal vector as (12). Consequently, the low complexity detector can be given as

$$\langle \hat{i}, \hat{j}, \hat{k} \rangle = \arg\min_{i,j,k} \|\mathbf{y}^i / \beta_i - \mathbf{c}_j^k\|^2, \tag{14}$$

where $\mathbf{c}_j^k = \mathbf{e}_j s_k$.

Let us now discuss the complexity of the proposed sub-optimal detection method in comparison to the optimal ML detection method. In the optimal ML detection method, the number of multiplications required in the first term of (11) is $N_r(N_r N_t + N_r) + M$, because the computation of $|s_k|^2$ is shared by all i and j. Furthermore, the second term of (11) requires $N_r M(N_r + 1)$ multiplications due to the fact that $\mathbf{H}\mathbf{p}_{i,j}$ is already computed in the first term. As a result, the total number of multiplications required in the ML detection method is $N_r{}^2(N_t + M + 1) + M(N_r + 1)$. On the other hand, the total number of multiplications required for the proposed sub-optimal detection method in (14) is $N_r N_g M$, which is significantly smaller than the number of multiplications required in the ML method. Additionally, we can observe that the detection complexity of ML method is not related to the grouping mode of the receive antennas. On the contrary, the detection complexity of the sub-optimal method increases linearly with N_g. In order to make the comparison more intuitive, the complexities of the low complexity detection method and the optimal ML detection method are shown in Table. 1.

TABLE 1. Complexity comparison of the low complexity detection method and the optimal ML detection method.

Methods	$N_t = 5, N_r = 8, M = 4$	$N_t = 8, N_r = 16, M = 4$
Low complexity	64	128
Optimal ML	676	3396

C. PERFORMANCE ANALYSIS

In this subsection, we will derive the BER performance of our group PSM scheme based on the optimal ML detector and the low complexity detector, respectively. Based on the union bound technique [23], the BER performance of group PSM under the optimal ML detector [3] can be formulated as

$$P \leq \frac{1}{\eta 2^\eta} \sum_{\langle i,j,k \rangle} \sum_{\langle \hat{i},\hat{j},\hat{k} \rangle \neq \langle i,j,k \rangle} D(\langle i,j,k \rangle, \langle \hat{i},\hat{j},\hat{k} \rangle) \times \mathbb{E}_{\mathbf{H}} \left\{ p_r(\langle i,j,k \rangle \to \langle \hat{i},\hat{j},\hat{k} \rangle | \mathbf{H}) \right\}, \tag{15}$$

where $\eta = \eta_{\text{group PSM}}$, $D(\langle i,j,k \rangle, \langle \hat{i},\hat{j},\hat{k} \rangle)$ denotes the Hamming distance between the signals $\langle i,j,k \rangle$ and $\langle \hat{i},\hat{j},\hat{k} \rangle$, $p_r(\langle i,j,k \rangle \to \langle \hat{i},\hat{j},\hat{k} \rangle | \mathbf{H})$ is the conditioned pairwise error probability (PEP) of detecting $\langle \hat{i},\hat{j},\hat{k} \rangle$ when the signal $\langle i,j,k \rangle$ is transmitted.

For a given channel matrix $\mathbf{H}$, the PEP of utilizing the ML method can be derived as

$$\begin{aligned} p_r(\langle i, j, k\rangle &\rightarrow \langle \hat{i}, \hat{j}, \hat{k}\rangle|\mathbf{H}) \\ &= Q\left(\sqrt{\frac{1}{2\sigma_n^2}\|\mathbf{H}(\mathbf{P}_i\mathbf{e}_j s_k - \mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}})\|^2}\right), \end{aligned} \quad (16)$$

where $Q(\cdot)$ represents the Gaussian Q-function which is defined as $Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^{\infty} e^{-\frac{t^2}{2}} dt$ [24].

When the low complexity detection method in (14) is adopted, the PEP in (15) can be formulated as

$$\begin{aligned} &p_r(\langle i, j, k\rangle \rightarrow \langle \hat{i}, \hat{j}, \hat{k}\rangle|\mathbf{H}) \\ &= p_r(\|\mathbf{y}^i - \beta_i\mathbf{e}_j s_k\|^2 > \|\mathbf{y}^{\hat{i}} - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2) \\ &= p_r(\|\mathbf{n}_i\|^2 > \|\mathbf{y}^{\hat{i}} - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2) \\ &= p_r\Big[\|\mathbf{n}_i\|^2 > \|\mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2 + \|\mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k\|^2 \\ &\quad + \|\mathbf{n}_{\hat{i}}\|^2 + \|\beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2 - 2\Re\left(\mathbf{n}_{\hat{i}}^H(\mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}})\right)\Big]. \end{aligned} \quad (17)$$

At a reasonably high SNR region, the impacts of $\mathbf{n}_i$ and $\mathbf{n}_{\hat{i}}$ can be negligible. Hence, by ignoring the norm of the noise, (17) can be further approximated as

$$\begin{aligned} p_r(\langle i, j, k\rangle &\rightarrow \langle \hat{i}, \hat{j}, \hat{k}\rangle|\mathbf{H}) \\ &\approx p_r\Big(2\Re\{\mathbf{n}_{\hat{i}}^H(\beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}} - \mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k)\} \\ &\qquad > \|\mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2\Big) \\ &= p_r\left(\Re\left(\mathbf{n}_{\hat{i}}^H\Psi\right) > \frac{A}{2}\right), \end{aligned} \quad (18)$$

where $A = \|\mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k - \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2$ and $\Psi = \beta_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}} - \mathbf{H}_{\hat{i}}\mathbf{P}_i\mathbf{e}_j s_k$. Since the entries of $\mathbf{n}_{\hat{i}}$ are i.i.d variables obeying $\mathcal{CN}(0, \sigma_n^2)$, the left side in the brackets of the last line is a Gaussian noise vector whose entries obeying $\mathcal{CN}(0, \|\Psi\|^2\sigma_n^2/2)$. Consequently, the PEP can be expressed as

$$p_r(\langle i, j, k\rangle \rightarrow \langle \hat{i}, \hat{j}, \hat{k}\rangle|\mathbf{H}) = Q\left(\sqrt{\frac{A}{2\sigma_n^2}}\right). \quad (19)$$

Finally, we have

$$P \leq \frac{1}{\eta 2^{\eta}} \sum_{\langle i,j,k\rangle} \sum_{\langle \hat{i},\hat{j},\hat{k}\rangle \neq \langle i,j,k\rangle} D(\langle i, j, k\rangle, \langle \hat{i}, \hat{j}, \hat{k}\rangle) \times \mathbb{E}_{\mathbf{H}}\left\{Q\left(\sqrt{\frac{A}{2\sigma_n^2}}\right)\right\}. \quad (20)$$

Since it is difficult to obtain the statistical distribution of the power normalization factor β_i and the entry of the pre-coding matrix exactly, the closed form of (15) and (20) could not be derived straightforward. Alternatively, we can solve this problem by the similar approach that introduced in [19], where the Gamma approximation method [25] was applied. More explicitly, we first employ a tight upper bound to substitute the Q-function [19]. Then, we regard the conditioned PEP of (16) or (19) as a Gamma distributed random variable. The shape parameter and scale parameter of the probability density function (PDF) of the Gamma random variable can be obtained by carrying out the Monte-Carlo simulation. After obtaining the PDF of the Gamma random variable, the approximations of the expectation term of (15) and (20) can be obtained by some integral operations.

III. POWER ALLOCATED GROUP PSM

Extensive researches have revealed that the power allocation method is an efficient way to against the adverse effect of channels and attain the significant performance gain. As mentioned in the section II, the transmit power corresponding to each receive antenna group are the same and time-invariant for the proposed scheme. In order to further improve the system performance, in this section, we will present a power allocation method for the group PSM scheme.

The proposed power allocation method intends to optimize the power allocation coefficients of each receive antenna group based on maximizing the minimum Euclidean distance [13] among received group PSM signals, leading to a non-convex problem. To solve this problem, the linearized approach is used to convert the non-convex problem to the convex problem, and it can be iteratively solved by a suitable optimization method.

Let $\mathbf{a} = [a_1, \ldots, a_i, \ldots, a_G]^T$ represent the power allocation vector, where a_i is the power coefficient of the i-th receive antenna group. In this case, the Euclidean distance between two received group PSM signals can be written as

$$d = \|a_i\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k - a_{\hat{i}}\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2. \quad (21)$$

Consequently, the optimization problem of the power allocated group PSM scheme under the transmit power constraint can be formulated as

$$\begin{aligned} &\max_{\mathbf{a}} \min_{\forall\langle i,j,k\rangle \neq \langle \hat{i},\hat{j},\hat{k}\rangle} \|a_i\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k - a_{\hat{i}}\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2, \\ &\text{s.t. } \|\mathbf{a}\|^2 \leq G, \\ &\quad a_i > 0, \quad i = 1, \ldots, G. \end{aligned} \quad (22)$$

We now transform the above optimization problem by introducing an auxiliary variable t [26] and applying some matrix manipulations [27], then we have

$$\begin{aligned} &\max_{\mathbf{a}} \quad t \\ &\text{s.t. } \mathbf{a}^T\mathbf{T}_{i,\hat{i}}\mathbf{a} \geq t, \forall\langle i, j, k\rangle \neq \langle \hat{i}, \hat{j}, \hat{k}\rangle, \\ &\quad \|\mathbf{a}\|^2 \leq G, \\ &\quad a_i > 0, i = 1, \ldots, G, \end{aligned} \quad (23)$$

In (23),

$$(\mathbf{T}_{i,\hat{i}})_{m,n} = \begin{cases} (\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k)^H\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k, & \text{if } m = i, n = i \\ -(\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k)^H\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}, & \text{if } m = i, n = \hat{i} \\ (\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}})^H\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}, & \text{if } m = \hat{i}, n = \hat{i} \\ -(\mathbf{H}\mathbf{P}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}})^H\mathbf{H}\mathbf{P}_i\mathbf{e}_j s_k, & \text{if } m = \hat{i}, n = i \\ 0, & \text{otherwise,} \end{cases} \quad (24)$$

where $i \neq \hat{i}$, and $(\mathbf{X})_{m,n}$ represents the element of the m-th row and n-th column of $\mathbf{X}$. When $i = \hat{i}$, $\mathbf{T}_{i,\hat{i}}$ is given by

$$(\mathbf{T}_{i,\hat{i}})_{m,n} = \begin{cases} \|\mathbf{HP}_i\mathbf{e}_j s_k - \mathbf{HP}_{\hat{i}}\mathbf{e}_{\hat{j}} s_{\hat{k}}\|^2, & \text{if } m = i, n = i \\ 0, & \text{otherwise.} \end{cases} \quad (25)$$

Obviously, (23) constructs a non-convex problem because of the first constraint. Thus, we linearize the quadratic constraint function in (23) based on the Taylor expansion, which can be given by

$$\mathbf{a}^T\mathbf{T}_{i,\hat{i}}\mathbf{a} = \Re(2\mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a} - \mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a}_n) \quad (26)$$

where $\mathbf{a}_n$ represents the optimal solution toward the n-th iteration. Therefore, the problem of optimizing the power allocation coefficient at n-th iteration can be reformulated as

$$\begin{aligned} &\max_{\mathbf{a}} t \\ &\text{s.t. } \Re(2\mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a} - \mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a}_n) \geq t, \forall\langle i,j,k\rangle \neq \langle \hat{i},\hat{j},\hat{k}\rangle \\ &\quad \|\mathbf{a}\|^2 \leq G \\ &\quad a_i > 0, \quad i = 1,\ldots,G. \end{aligned} \quad (27)$$

Problem (27) is convex because $\Re(2\mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a} - \mathbf{a}_n^T\mathbf{T}_{i,\hat{i}}\mathbf{a}_n)$ is an affine function of $\mathbf{a}$, which can be solved by the CVX tool [28]. Finally, the solution of (23) can be obtained by iteratively solving the problem (27) until satisfying the convergence condition $\|\mathbf{a}_{n+1} - \mathbf{a}_n\| \leq \delta$.

IV. GENERALIZATION

In this section, we extend and generalize the proposed group PSM scheme, where only a single receive antenna is selected. The generalized group PSM scheme allows several receive antenna groups to be selected simultaneously, which improves the multiplexing gain and flexibility of group PSM.

To be specific, we assume that N_a receive antenna groups are selected during each time slot, which means that there are N_a spatial domain signals and modulated signals transmitted simultaneously during each time slot. Furthermore, let $|\mathcal{C}_g| = \binom{G}{N_a}$ denotes the set size of all the combinations of choosing N_a receive antenna groups from G receive antenna groups. For example, in a system where $G = 4$ and $N_a = 2$, the number of legitimate group patterns $|\mathcal{C}_g|$ is equal to 6 and the possible group patterns are given by $\mathcal{C}_g = \{[1,2],[1,3],[1,4],[2,3],[2,4],[3,4]\}$. In which, $N_s = 2^{\lfloor \log_2(|\mathcal{C}_g|)\rfloor}$ group patterns can be used for transmission.

Hence, the transmit rate of the generalized group PSM scheme can be given by

$$\gamma = \log_2 N_s + N_a\left(\log_2 N_g + \log_2 M\right), \quad (28)$$

From (28), we can see that the generalized group PSM scheme offers superior bandwidth efficiency compared with group PSM. We also let $\mathbf{s}_l = [s_{l_1}, s_{l_2}, \ldots, s_{l_{N_a}}]^T$ and $\mathbf{b}_n = [\mathbf{e}_{n_1}, \mathbf{e}_{n_2}, \ldots, \mathbf{e}_{n_{Na}}]^T$, where s_{l_r} and $\mathbf{e}_{n_r}$ are the modulated symbol and the spatial domain symbol corresponding to the r-th receive antenna group. By combining the $\mathbf{s}_l$ and $\mathbf{b}_n$, we can obtain

$$\Omega_l^n = [s_{l_1}\mathbf{e}_{n_1}, \ldots, s_{l_{N_a}}\mathbf{e}_{n_{N_a}}]^T. \quad (29)$$

Furthermore, we denote $\mathbf{Q}_m = [\mathbf{P}_{m_1}, \mathbf{P}_{m_2}, \ldots, \mathbf{P}_{m_{N_a}}]$, where $m = 1, 2, \ldots, N_s$ and $\mathbf{P}_{m_r}$ is the pre-coding matrix corresponding to the r-th receive antenna group. After the pre-coding process, the resultant transmit signal can be given by

$$\begin{aligned} \mathbf{x} &= \mathbf{Q}_m\Omega_l^n \\ &= [\mathbf{P}_{m_1}, \ldots, \mathbf{P}_{m_{N_a}}][s_{l_1}\mathbf{e}_{n_1}, \ldots, s_{l_{N_a}}\mathbf{e}_{n_{N_a}}]^T. \end{aligned} \quad (30)$$

Consequently, the observation at the receiver side can be written as

$$\begin{aligned} \mathbf{y} &= \mathbf{Hx} + \mathbf{n} \\ &= \frac{1}{\sqrt{Na}}\mathbf{H}[\mathbf{P}_{m_1}, \ldots, \mathbf{P}_{m_{N_a}}][s_{l_1}\mathbf{e}_{n_1}, \ldots, s_{l_{N_a}}\mathbf{e}_{n_{N_a}}]^T + \mathbf{n}. \end{aligned} \quad (31)$$

The receiver reconstructs the transmitted signal by using the ML detector which can be given by

$$< \hat{\mathbf{Q}}_m, \hat{\Omega}_l^n >= \arg\min_{\mathbf{Q}_m, \Omega_l^n} \|\mathbf{y} - \frac{1}{\sqrt{Na}}\mathbf{HQ}_m\Omega_l^n\|^2. \quad (32)$$

Now, we extend our power allocation method to the generalized group PSM scheme for the sake of achieving further improvement in the system performance. For the generalized group PSM scheme, we provide a similar optimization problem to the group PSM scheme. In group PSM, the power allocation coefficients of each receive antenna group are optimized while the power allocation coefficients of each group pattern are optimized in generalized group PSM.

More specifically, we consider that the antennas in each receive antenna group pattern have the same power allocation coefficient. Let $\widetilde{\mathbf{a}} = \left[\widetilde{a}_1, \ldots, \widetilde{a}_m, \ldots, \widetilde{a}_{N_s}\right]^T$, where $\widetilde{a}_m$ represents the power allocation coefficient of the m-th group pattern. Then, the optimization problem can be given by

$$\begin{aligned} &\max_{\widetilde{\mathbf{a}}} t \\ &\text{s.t. } \widetilde{\mathbf{a}}^T\widetilde{\mathbf{T}}_{m,\hat{m}}\widetilde{\mathbf{a}} \geq t, \forall\langle m,n,l\rangle \neq \langle \hat{m},\hat{n},\hat{l}\rangle, \\ &\quad \|\widetilde{\mathbf{a}}\|^2 \leq N_s, \\ &\quad \widetilde{a}_m > 0, \quad m = 1,\ldots,N_s, \end{aligned} \quad (33)$$

where the entries of $\widetilde{\mathbf{T}}_{m,\hat{m}}$ are given by

$$(\widetilde{\mathbf{T}}_{m,\hat{m}})_{u,v} = \begin{cases} (\mathbf{HQ}_m\Omega_l^n)^H\mathbf{HQ}_m\Omega_l^n, & \text{if } u = m, v = m \\ -(\mathbf{HQ}_m\Omega_l^n)^H\mathbf{HQ}_{\hat{m}}\Omega_{\hat{l}}^{\hat{n}}, & \text{if } u = m, v = \hat{m} \\ (\mathbf{HQ}_{\hat{m}}\Omega_{\hat{l}}^{\hat{n}})^H\mathbf{HQ}_{\hat{m}}\Omega_{\hat{l}}^{\hat{n}}, & \text{if } u = \hat{m}, v = \hat{m} \\ -(\mathbf{HQ}_{\hat{m}}\Omega_{\hat{l}}^{\hat{n}})^H\mathbf{HQ}_m\Omega_l^n, & \text{if } u = \hat{m}, v = m \\ 0, & \text{otherwise} \end{cases} \quad (34)$$

for $m \neq \hat{m}$, and when $m = \hat{m}$ we have

$$(\widetilde{\mathbf{T}}_{m,\hat{m}})_{u,v} = \begin{cases} \|\mathbf{HQ}_m\Omega_l^n - \mathbf{HQ}_{\hat{m}}\Omega_{\hat{l}}^{\hat{n}}\|^2, & \text{if } u = m, v = m \\ 0, & \text{otherwise.} \end{cases} \quad (35)$$

Notice that the problem (33) can be iteratively solved by the approach derived in Section III.

V. SIMULATION RESULTS

In this section, we provide simulation results to evaluate the performance of the proposed group PSM scheme and the generalized group PSM scheme. For comparison, we also provide the performance of the conventional PSM scheme and the JSM scheme. The M-QAM modulation is used in our simulations.

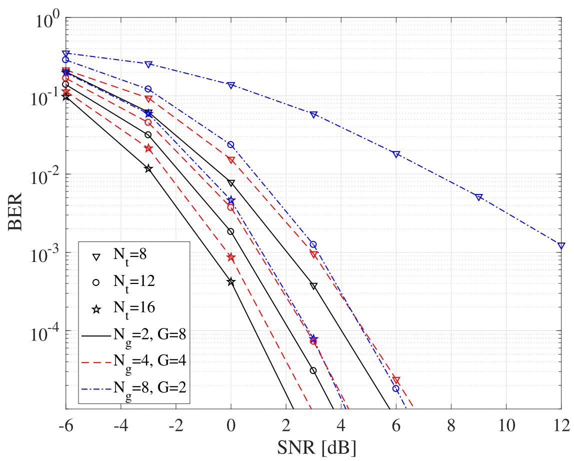

FIGURE 2. BER performance of group PSM with optimal ML detector, where $(N_t, N_r, N_g, M) = (\{8, 12, 16\}, 16, \{2, 4, 8\}, 4)$.

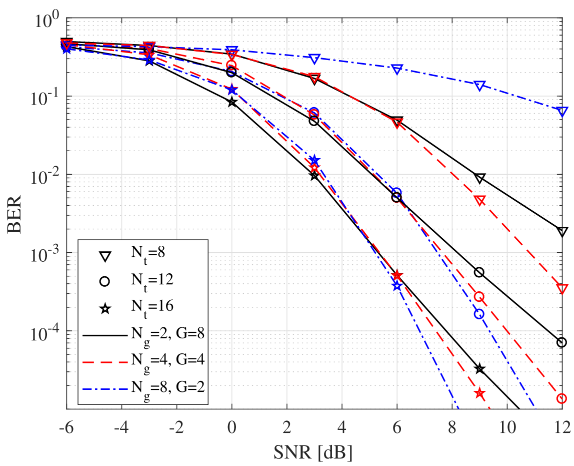

FIGURE 3. BER performance of group PSM with low complexity detector, where $(N_t, N_r, N_g, M) = (\{8, 12, 16\}, 16, \{2, 4, 8\}, 4)$.

In Fig. 2 and Fig. 3, we discuss the parameter selection for group PSM by varying (N_g, G) under a fixed $N_r = 16$ with the optimal ML detector and the low complexity detector, where $(N_t, N_r, N_g, M) = (\{8, 12, 16\}, 16, \{2, 4, 8\}, 4)$. As shown in Fig. 2, the BER performance improves with the number of transmit antennas N_t increases. This is because the transmit diversity increases with N_t. Furthermore, we can also find that the BER performance improves with the decrease of N_g, which implies that the optimal system setup is $N_g = 2$ in the case of using the optimal ML detector. It is because, in this case, the detection reliability of the index of antennas in the receive antenna group dominates the performance of the group PSM scheme.

From Fig. 3 we can observe that, in the case of $N_g = 8$, the BER performance improves drastically with the number of transmit antennas increases from 8 to 12 and 16. Furthermore, in the case of $N_t = 12$ and $N_t = 16$, $N_g = 2$ yields the best performance at the low SNR region while $N_g = 8$ yields the best performance at the high SNR region, which means that as N_g increases, the system achieves a better BER performance under the sufficient transmit diversity and transmit power. This results from the fact that when we use the low complexity detection method, the detection reliability of the index of receive antenna group becomes dominant owing to the interference between receive antenna groups.

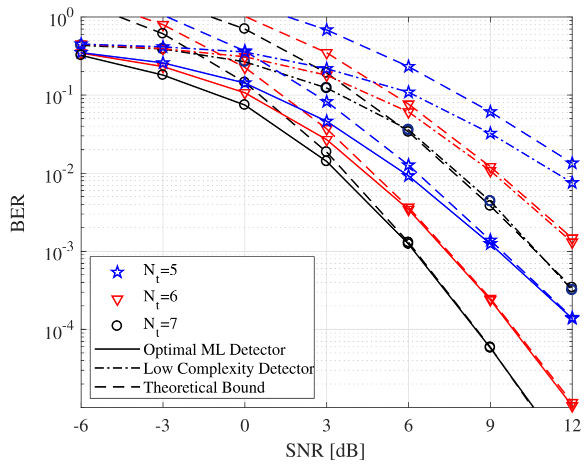

FIGURE 4. BER performance comparison of different detector, where $(N_t, N_r, N_g, M) = (\{5, 6, 7\}, 8, 4, 4)$.

Fig. 4 compares the BER performances of our group PSM under the optimal ML detector and the low complexity detector, respectively. In this case, we set $(N_t, N_r, N_g, M) = (\{5, 6, 7\}, 8, 4, 4)$. The theoretical ABEP curves of (15) and (20) are also drawn for verification. It can be observed that the simulation results verify the accuracy of theoretical bounds. In particular, the theoretical bounds are asymptotically accurate with an increasing number of transmit antennas. Additionally, the optimal ML detection method achieves better BER performance than the low complexity detention method.

In Fig. 5, we show the performances of the conventional PSM scheme and the group PSM scheme with the low complexity detection method under symmetric MIMO systems. In the case of $N_t = N_r = 8$, we take $N_g = 4$. It is shown that the group PSM scheme significantly outperforms the conventional PSM scheme when the ZF pre-coding method is adopted. Moreover, the BER performance of our scheme is almost the same under two pre-coding methods. For the MMSE pre-coding method, the performance of conventional PSM is superior to the proposed scheme. It is because the

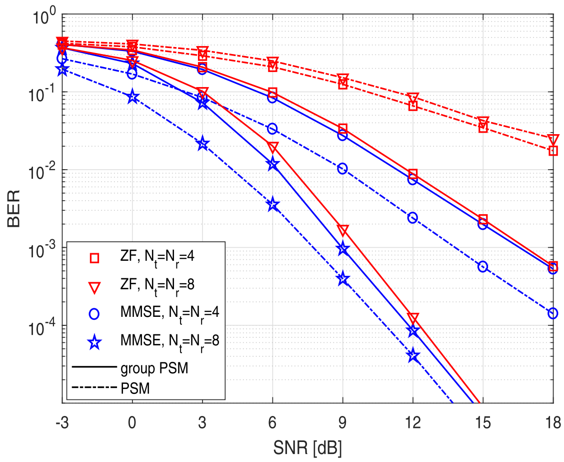

FIGURE 5. Comparison between the BER performance of our group PSM with the low complexity detector and conventional PSM under symmetric MIMO systems, where $M = 4$.

interference between receive antenna groups is more dominant than the noise when we use the MMSE pre-coding method.

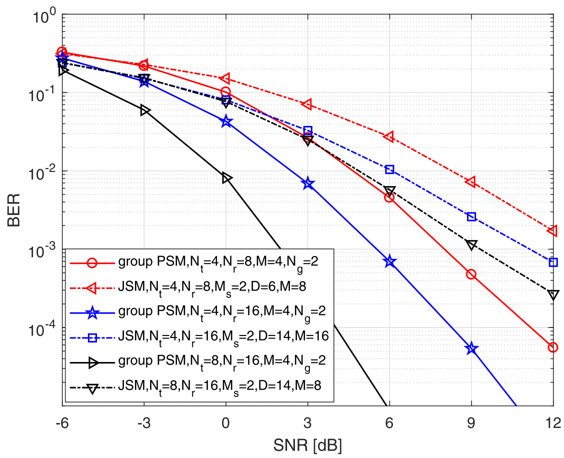

FIGURE 6. Comparison between the BER performance of group PSM and JSM over-determined MIMO systems.

In Fig. 6, we compare the BER performance of the proposed scheme with that of the JSM scheme over 4×8, 4×16 and 8×16 MIMO channels with $\eta = 5$, $\eta = 6$ and $\eta = 6$, respectively. We assume $M_s = 2$ in the JSM scheme, where M_s is the number of antennas of each transmit pattern. Additionally, D denotes the number of diversity antennas. It is shown that the proposed group PSM scheme achieves better performance than the JSM scheme.

Fig. 7 shows the BER performances of our generalized group PSM scheme under different N_g, where $N_t = 8$ and $N_r = 16$. In Fig. 7, the system throughput $\eta = 10$ when $N_a = 1, 2$ and $\eta = 14$ when $N_a = 3$, where the case of $N_a = 1$ represents the group PSM scheme. We can find that the BER performances of our generalized group PSM

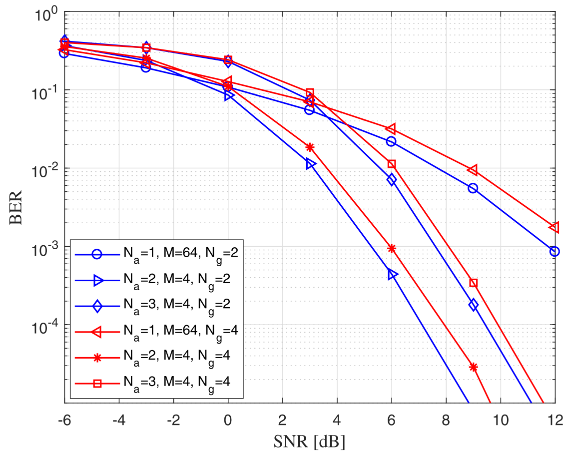

FIGURE 7. BER performance of the generalized group PSM scheme, where $N_t = 8, N_r = 16$.

is better than that of the group PSM scheme under the same system throughput. Furthermore, the BER performance of the generalized group PSM system is also improved with N_g decreases.

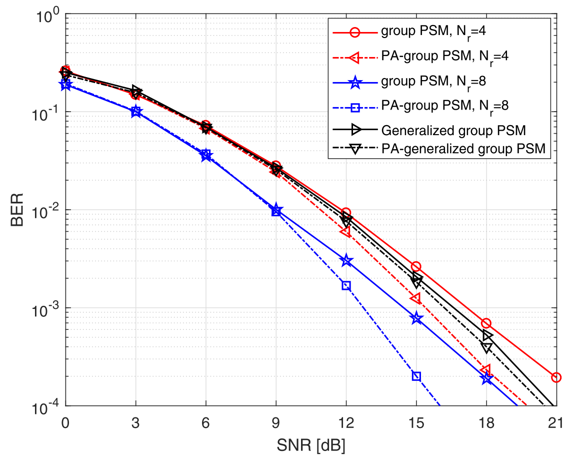

FIGURE 8. BER performance of PA-group PSM and PA-generalized group PSM, where $(N_t, N_r, N_g, M, N_a) = (2, \{4, 6, 8\}, 2, \{2, 4\}, 2)$.

Fig. 8 shows the BER performances of power allocated group PSM and power allocated generalized group PSM (which we called PA-group PSM and PA-generalized group PSM in the following) and compare them with group PSM and generalized group PSM, where $(N_t, N_r, N_g, M) = (2, \{4, 8\}, 2, 4)$ for group PSM schemes and $(N_t, N_r, N_g, M, N_a) = (2, 6, 2, 2, 2)$ for generalized group PSM schemes. As can be seen in Fig. 8, the power allocated schemes outperform the group PSM scheme and the generalized group PSM scheme. In particularly, PA-generalized group PSM has a slightly better BER performance than generalized group PSM. Besides, an additional SNR gain of almost 2dB can be observed between PA-group PSM and group PSM when $N_r = 8$ at the BER of 10^{-3}.

VI. CONCLUSION

In this paper, we proposed a group PSM scheme for over-determined MIMO systems. By employing the proposed group PSM scheme, the incoming bits are mapped to the index of the selected receive antenna group, the index of the activated antenna of the selected group and the conventional modulated symbol. In order to reduce the receiver complexity, a low complexity detection method was presented. Furthermore, we generalized the proposed group PSM to improve system throughput and flexibility. We also investigated the optimization of the power allocation coefficient for group PSM and generalized group PSM. Simulation results showed that the proposed group PSM scheme and the generalized group PSM scheme achieve a good BER performance and also exhibit a high flexibility for implementation.

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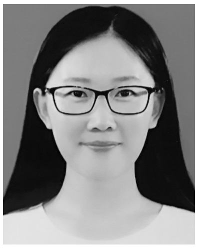

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