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Power Allocation-Aided Spatial Modulation for Limited-Feedback MIMO Systems

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***Abstract*—Adaptive power allocation (PA) algorithms based on optimization of the minimum distance $d_{\min}$ between signal points at the receiver side are investigated in spatial modulation (SM) systems. First, a closed-form solution of the optimal maximum-$d_{\min}$-aided PA algorithm is derived in the case of two transmit antennas (TAs). Moreover, for a higher number of TAs, we propose a numerical approach, which appropriately splits the power between the specific TA pair associated with $d_{\min}$ to increase this distance. Furthermore, our PA-aided SM systems may be readily combined with adaptive modulation (AM) to further improve the system's performance. Our numerical results show that the proposed algorithms provide beneficial system performance improvements compared with both the equal-gain PA-based SM and the identical-throughput PA-aided spatial multiplexing systems.**

***Index Terms*—Limited-feedback systems, multiple-input–multiple-output (MIMO), power allocation (PA), spatial modulation (SM).**

I. Introduction

Spatial modulation (SM), which maps information both to a carefully designed combination of antenna indices and to the classic amplitude and phase modulation (APM) constellation, constitutes a promising low-complexity multiple-input–multiple-output (MIMO) transmission technique [1]–[5]. The SM-based systems are capable of outperforming some of the classic MIMO techniques [6] even in the presence of channel estimation errors; however, they can only offer receive diversity [7], [8].

To overcome this problem, link adaptation (LA) schemes have been proposed in [9]–[15], where the transmit parameters are dynamically adapted to the channel conditions. Specifically, the effects of power imbalance [9], the issues of achieving transmit diversity [10], the particular choice of the constellation used [11], and the impact of cooperation have been researched [12]. However, most of the aforementioned LA schemes considered only a special case of SM, i.e., space-shift keying [3], which exclusively employs the antenna indices for data modulation. In [14], the Kronecker model was used to characterize a correlated SM-MIMO channel, and a beamforming codebook design algorithm was proposed to optimize the bit-error-ratio (BER) performance based on the rather limited knowledge of the channel envelope's spatial correlation. In [15], a power-scaling-assisted SM scheme was proposed, where a scaling factor (SF) was invoked for weighting the modulated symbols before their transmission. However, the related design algorithm of SF was not provided. Recently, we have proposed an adaptive modulation (AM)-aided SM (ASM) scheme [16], [17] to improve the attainable system performance. In ASM, the receiver requests the most suitable modulation order to be used by the transmitter for each transmit antenna (TA). However, constant-power ASM may not be capable of fully exploiting the available spatial-domain grade of freedom offered by MIMO channel.

Power allocation (PA) techniques are capable of alleviating the adverse effects of channel fading to achieve either an increased data rate or a reduced BER. Indeed, PA has been lavishly researched in the context of spatial multiplexing systems [18], [19]. As a new MIMO technique, SM may be also beneficially combined with PA for adjusting the transmission parameters for the sake of accommodating time-varying channels. However, since only a single TA is activated in each time slot in SM-based schemes, the PA approaches designed for spatial multiplexing-based MIMO systems may not be directly suitable for SM systems.

Against this background, the novel contributions of this paper are as follows.

- We investigate the benefits of adaptive PA based on the maximum-free distance (max-FD) $d_{\min}$ between the pairs of signal constellation points at the receiver side. An optimal $d_{\min}$ PA precoder is derived for BPSK-modulated (2×1)-element SM. Then, this result is extended to M-PSK/M-ary quadrature amplitude modulation (M-QAM) $(2 \times N_r)$-element PA-aided SM.
- To deal with the case of $N_t > 2$, we propose a numerical approach, termed as the "worse-case-first"-based PA (WCF-PA) algorithm, which appropriately splits the power between the specific TA pair associated with $d_{\min}$. As a further benefit, our PA-aided SM systems may be readily combined with AM techniques for the sake of maximizing the FD, hence improving the system's performance.

The organization of this paper is as follows. Section II presents the system model of the PA-aided SM. In Section III, we introduce our PA algorithms, whereas our simulation results and performance comparisons are presented in Section IV. Finally, Section V concludes this paper.

Notation: $(\cdot)^*$, $(\cdot)^T$, and $(\cdot)^H$ denote conjugate, transpose, and Hermitian transpose, respectively. Furthermore, $\|\cdot\|$ stands for the Frobenius norm, and $Re\{\cdot\}$ denotes the real part of a complex variable. $\lceil x \rceil$ denotes the smallest integer higher than or equal to x.

II. System Model

A. Transceiver

Let us consider a flat-fading MIMO channel associated with N_t TAs and N_r receiver antennas, represented by an $(N_r \times N_t)$-element matrix $\mathbf{H}$. The entries of $\mathbf{H}$ are assumed to be independent identically

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distributed and obeying $h_{vu} \sim \mathcal{CN}(0,1)$. Then, the transmitted PA-aided SM symbol $\mathbf{x} \in \mathbb{C}^{N_t \times 1}$ is given as

$$\mathbf{x} = s_l^q \mathbf{e}_q = [0, \ldots, \underset{\uparrow\, qth}{s_l^q}, \ldots, 0]^T \tag{1}$$

where s_l^q is the APM symbol assigned to the qth TA, such as r^q-QAM, which is associated with $d^q = \log_2(r^q)$ input bits, whereas $\mathbf{e}_q \in \mathbb{C}^{N_t \times 1}$ is selected from the N_t-dimensions standard basis vectors (i.e., $\mathbf{e}_1 = [1, 0, \ldots, 0]^T$), according to the $\log_2(N_t)$ input bits.

At the receiver, the corresponding $(N_r \times 1)$-element received signal vector is given by

$$\mathbf{y} = \mathbf{HPx} + \mathbf{n} \tag{2}$$

where the elements of the N_r-element noise vector $\mathbf{n}$ are Gaussian random variables obeying $\mathcal{CN}(0, N_0)$, and the diagonal matrix $\mathbf{P}$ allocates the total power P_T to the different TAs, yielding

$$\mathbf{P} = \begin{pmatrix} \sqrt{p_1} & & 0 \\ & \ddots & \\ 0 & & \sqrt{pN_t} \end{pmatrix} \tag{3}$$

where the diagonal elements are limited by the total power constraint as $\sum_{q=1}^{N_t} p_q = P_T$.

B. Performance Metric

Given the channel matrix $\mathbf{H}$, the pairwise error probability (PEP) of the SM system using the maximum-likelihood (ML) detector is given by [19]

$$P(\mathbf{x}_i \to \mathbf{x}_j | \mathbf{H}) \approx \lambda \cdot Q\left(\sqrt{\frac{1}{2N_0} d_{\min}(\mathbf{H})}\right) \tag{4}$$

where $Q(x) = (1/\sqrt{2\pi}) \int_x^\infty e^{-(y^2/2)} dy$, and λ is the number of neighbor constellation points [19] having the FD $d_{\min}(\mathbf{H})$, which is defined as

$$\begin{aligned} d_{\min}(\mathbf{H}) &= \min_{\substack{\mathbf{x}_i, \mathbf{x}_j \in \mathbb{X}, \\ \mathbf{x}_i \neq \mathbf{x}_j}} \|\mathbf{HP}(\mathbf{x}_i - \mathbf{x}_j)\|_F^2 \\ &= \min_{\mathbf{e}_{ij} \in \mathbb{E}} \|\mathbf{HP}\mathbf{e}_{ij}\|_F^2 \end{aligned} \tag{5}$$

where $\mathbb{X}$ is the set of all legitimate transmit symbols, whereas $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$ denotes the error vector and $\mathbb{E}$ is a set of error vectors. The complexity for calculating the FD of our SM-based system in (5) is relatively modest because of the following two reasons.

1) In (5), the collinear distance vectors generate the same distance. Hence, although the cardinality of the set $\mathbb{E}$ may be large, only a representative subset of those collinear error vectors has to be considered.
2) Unlike in conventional MIMO techniques, the transmit vectors of SM-MIMO schemes are sparsely populated since they have mostly zero values. This property can be exploited to further reduce the complexity of evaluating (5), as detailed in [17].

C. Optimization Criterion

Since the error events mainly arise from the nearest neighbors, the maximization of the FD in (5) directly reduces the probability of error, particularly at high-signal-to-noise ratios (SNRs) [19]. Note that the PA solution based on the FD is an attractive LA regime, which has been vastly researched in the context of spatial multiplexing systems. However, these PA approaches designed for spatial multiplexing-based MIMO systems may not be directly suitable for SM systems [20]–[22] because only a single TA is activated in each time slot. Based on (5), we propose a PA-aided SM system, which adapts the PA matrix $\mathbf{P}$ to maximize the FD under the transmit power constraint as

$$\begin{aligned} &\mathbf{P}_{\text{opt}} = \arg\max_{\mathbf{P}} d_{\min}(\mathbf{H}) \\ &\text{s.t.} \quad tr(\mathbf{P}^2) = P_T \end{aligned} \tag{6}$$

where $tr(\cdot)$ denotes the matrix trace.

In most of the PA algorithms conceived for VBLAST, the power is shared among the different TAs (space-only PA). This principle has been also adopted in our PA schemes, and hence, a fixed total power constraint is imposed on all TAs in (6). Since only a single TA is activated in each time slot in SM schemes, unlike in the PA algorithms designed for VBLAST, the PA of our proposed SM scheme beneficially exploits the time domain for maintaining the total power constraint. Under the assumption that all TAs are selected for transmission with equal probability, the average of the transmit power is fixed.

Moreover, it is noted that there are two main differences of ASM [16] and the proposed PA-aided SM schemes.

- They exploit different properties of the MIMO channels. Specifically, ASM dynamically adapts the modulation order assigned to TAs, whereas the PA-aided SM adapts the power assigned to these antennas.
- The modulation orders of ASM are selected from a discrete set, whereas the PA parameters are chosen from the real-valued field.

III. PA Algorithms

Here, we first derive a closed-form solution of (6) for BPSK-modulated (2×1)-element SM and then extend the method to the more general M-PSK/M-QAM modulated $(2 \times N_r)$-element PA-aided SM scenario. In the case of large TAs and high modulation order, deriving a closed-form solution to (6) remains an open challenge since the solution is obtained by exhaustive search for large search space constituted by all legitimate candidate transmit symbols or error vectors. Hence, a numerical search method is proposed.

A. Optimal-FD PA Matrix for a BPSK-modulated 2 × 1 SM

For BPSK-modulated SM associated with $N_t = 2$ and $N_r = 1$, the symbols belong to the set $\{1, -1\}$, and all possible error vectors $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$ are listed as follows: $\{[-2, 0]^T, [2, 0]^T, [0, -2]^T, [0, 2]^T, [-1, 1]^T, [-1, -1]^T, [1, -1]^T$, and $[1, 1]^T\}$. Since some of the vectors are collinear, the set to be studied is reduced to $\{\mathbf{e}_1, \mathbf{e}_2, \mathbf{e}_3, \mathbf{e}_4\} = \{[2, 0]^T, [0, 2]^T, [1, -1]^T, [1, 1]^T\}$. Given the (1×2) channel matrix $\mathbf{H} = [h_1, h_2]$, the received constellation point distances are given by

$$\begin{cases} d_1 = \|\mathbf{HP}\mathbf{e}_1\|^2 = 4p_1 \|h_1\|^2 \\ d_2 = \|\mathbf{HP}\mathbf{e}_2\|^2 = 4p_2 \|h_2\|^2 \\ d_3 = \|\mathbf{HP}\mathbf{e}_3\|^2 = \left\|\sqrt{p_1} h_1 - \sqrt{p_2} h_2\right\|^2 \\ d_4 = \|\mathbf{HP}\mathbf{e}_4\|^2 = \left\|\sqrt{p_1} h_1 + \sqrt{p_2} h_2\right\|^2. \end{cases} \tag{7}$$

Based on (7), the optimization problem of (6) can be simplified to

$$\begin{aligned} &\mathbf{P}_{\text{opt}} = \arg\max_{\mathbf{P}} \{\min\{d_1, d_2, d_3, d_4\}\} \\ &\text{s.t.} \quad p_1 + p_2 = P_T. \end{aligned} \tag{8}$$

As indicated in (8) and shown in Fig. 1, d_1 and d_2 are linear functions of parameter p_1, whereas d_3 and d_4 are convex or concave

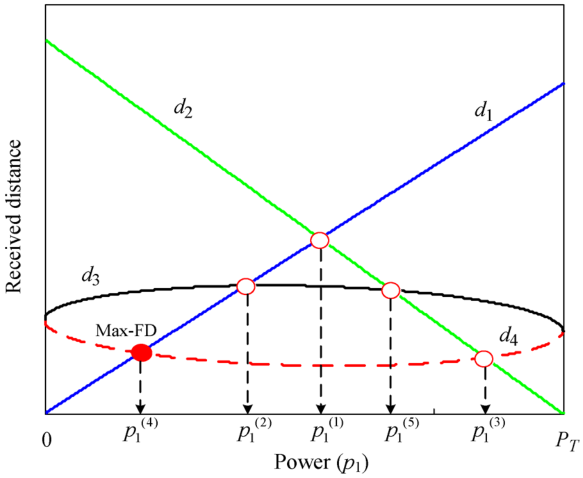

Fig. 1. Optimal-FD-aided PA for BPSK-modulated 2×1 SM.

functions and represented by an ellipse as a function of the power p_1. Hence, the max-FD solution according to p_1 is one of the intersections between these received distances $d_i (i = 1,2,3,4)$. More specifically, we can obtain the power assigned to the TA by finding these intersections and then selecting the one having the maximum FD as the final solution. As a result, the received FD is maximized, and then, the error performance is improved. To be specific, observe from (8) and for the total power constraint that the power $p_1^{(1)}$ associated with the first intersection in Fig. 1 satisfies

$$\begin{cases} d_1 = 4p_1^{(1)} \|h_1\|^2 = d_2 = 4p_2^{(1)} \|h_2\|^2 \\ p_1^{(1)} + p_2^{(1)} = p_T. \end{cases} \tag{9}$$

Upon introducing the shorthand of $a = \|h_1\|^2$, $b = \|h_2\|^2$, and $c = h_1^* h_2 + h_2^* h_1 = 2Re\{h_1^* h_2\}$ for a given channel matrix $\mathbf{H}$ and using (9), we obtain

$$\begin{cases} p_1^{(1)} = b/(a+b)P_T \\ p_2^{(1)} = a/(a+b)P_T. \end{cases} \tag{10}$$

Then, the power $p_1^{(2)}$ associated with second intersection of $d_1 = d_3$ in Fig. 1 is given by

$$\begin{cases} 4p_1^{(2)} \|h_1\|^2 = \left\| \sqrt{p_1^{(2)}} h_1 - \sqrt{p_2^{(2)}} h_2 \right\|^2 \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{11}$$

To elaborate a little further, (11) can be simplified to

$$\begin{cases} 3ap_1^{(2)} - bp_2^{(2)} + c\sqrt{p_1^{(2)}}\sqrt{p_2^{(2)}} = 0 \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{12}$$

Then, (12) can be viewed as a quadratic function of $\sqrt{p_1^{(2)}}$, which can be solved

$$\begin{cases} \sqrt{p_1^{(2)}} = \frac{-c+\sqrt{c^2+12ab}}{6a} \sqrt{p_2^{(2)}} \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{13}$$

From (13), we can then evaluate the power as

$$\begin{cases} p_1^{(2)} = \frac{c^2+6ab-c\sqrt{c^2+12ab}}{18a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T \\ p_2^{(2)} = \frac{18a^2}{18a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{14}$$

Similar to the evaluation process of $p_1^{(2)}$, we can obtain the candidate power $p_1^{(3)}$ associated with $d_1 = d_4$, the power $p_1^{(4)}$ associated with $d_2 = d_3$, and the power $p_1^{(5)}$ associated with $d_2 = d_4$ step by step, which are given by

$$\begin{cases} p_1^{(3)} = \frac{c^2+6ab+c\sqrt{c^2+12ab}}{18a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_1^{(4)} = \frac{c^2+6ab+c\sqrt{c^2+12ab}}{2a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_1^{(5)} = \frac{c^2+6ab-c\sqrt{c^2+12ab}}{2a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{15}$$

Then, based on the fixed total power constraint, the corresponding power assigned to the second TA is given by

$$\begin{cases} p_2^{(3)} = \frac{18a^2}{18a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_2^{(4)} = \frac{2a^2}{2a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_2^{(5)} = \frac{2a^2}{2a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{16}$$

Additionally, the solution $p_1^{(6)}$ associated with $d_3 = d_4$ satisfies $p_1^{(6)} p_2^{(6)} = 0$. Since the activation of the TAs conveys the information bits, the PA solution of $p_1^{(6)} = 0$ or $p_1^{(6)} = P_T$ $(p_2^{(6)} = 0)$ is not considered as a legitimate one. In conclusion of the algorithm, the FDs of these PA solutions are generated, and we select the one having the largest FD as our final result. Next, the aforementioned method is extended to M-PSK/M-QAM modulated PA-aided $(2 \times N_r)$-element SM. Here, the value of N_r is an arbitrary positive integer.

The detailed max-FD-aided PA algorithm is summarized in two steps as follows.

- Step 1: Compute all legitimate error vectors $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$, and eliminate the redundant collinear elements. Calculate all legitimate received constellation distances $d_l (1 = 1, \ldots, L)$ with the aid of the channel matrix $\mathbf{H}$ and $\mathbf{e}_{ij}$, which are either linear or nonlinear but convex functions of power p_1.
- Step 2: Find all possible intersections between the received constellation distances d_i and d_j $(i, j \in \{1, \ldots, L\})$, and calculate both the corresponding power matrix $P = \text{diag}\{\sqrt{p_1}, \sqrt{p_2}\}$ and the corresponding FD. Select the one having the largest FD as our final result.

Therefore, the allocated power to TA can be decided as a closed-form solution by the aforementioned steps with low complexity. Note that the restriction to $2 \times N_r$-element SM is imposed by the difficulty of the FD optimization, and the solution of the general problem remains an open challenge. Indeed, the determination of the PA matrix that maximizes the FD of (5) is difficult for two reasons: First, the solution depends on both the channel matrix and on the symbol alphabet, and the space of solutions is excessive. Hence, for a higher throughput, we propose a simple numerical approach for this difficult optimization problem.

B. WCF-PA

To deal with the case of $N_t > 2$ and high modulation orders, the conventional greedy algorithm-based PA (GA-PA) of [17] can be further developed for our SM systems. To be specific, at each step of the GA-PA algorithm, a small fraction Δp of the total power is allocated to that specific TA, which maximizes the FD. By contrast, the power of all the other TAs remains unchanged. As the total power P_T is gradually allocated, the final PA matrix $\mathbf{P}$ is approached. However, the GA-PA algorithm has to tentatively allocate power to all possible TAs in each iteration, which imposes high complexity.

TABLE I
PROPOSED WCF-PA ALGORITHM

Step1: Initialize $\mathbf{P}$ as a diagonal matrix with equal values $p_q = 1$, $q = 1, \cdots, N_t$.
Set the PA granularity Δp to a small number divisible by $\tilde{P}_T = 1$, such as $\Delta p = 0.05\tilde{P}_T$.
Let the consumed power of a single antenna be $\tilde{P}_u$. Set $k = 0$, $\mathbf{P}_{\text{ave}} = \mathbf{P}$, $\mathbf{P}^a = \mathbf{P}$ and $\mathbf{P}^b = \mathbf{P}$.
Set all entries of the $(N_t \times 1)$-element FD vectors $\mathbf{d}_{\min}(q)$ to zero.

Step2: calculate the $d_{\min}(\mathbf{H})$ of the PA matrix $\mathbf{P}$ as $d^1_{\min}$ and find the indices of the transmit antenna pair (m, n) achieved the value $d^1_{\min}$.

Step3: gradual power allocation between two TAs.
(a) If $m = n$, the distance $d^1_{\min}$ is achieved by one of the TAs, whereas two different signal constellation points are adopted.
while $\tilde{P}_u < \tilde{P}_T$,
Let $\Delta\hat{p} = \min\{\Delta p, \tilde{P}_T - \tilde{P}_u\}$ and $k = k + 1$;
Compute the norm $\|\mathbf{h}_q\|_F$ for $q = 1, \cdots, N_t$, find the TA u associated with the maximum norm as $u = \arg\max_q \|\mathbf{h}_q\|_F$ and update $p_u = \sqrt{1 - \Delta\hat{p}}$.
Then find the column m associated with the minimum norm as $m = \arg\min_q \|\mathbf{h}_q\|_F$ and update $p_m = \sqrt{1 + \Delta\hat{p}}$. Calculate the corresponding FD of $\mathbf{d}_{\min}(k) = \min_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}\mathbf{e}_{ij}\|^2_F$.
end
(b) elseif $m \neq n$, $d^1_{\min}$ is computed for different TAs.
while $\tilde{P}_u < \tilde{P}_T$,
Let $\Delta\hat{p} = \min\{\Delta p, \tilde{P}_T - \tilde{P}_u\}$ and $k = k + 1$;
Update $\mathbf{P}^a(n, n) = \sqrt{1 - \Delta\hat{p}}$ and $\mathbf{P}^a(m, m) = \sqrt{1 + \Delta\hat{p}}$;
Update $\mathbf{P}^b(n, n) = \sqrt{1 + \Delta\hat{p}}$ and $\mathbf{P}^b(m, m) = \sqrt{1 - \Delta\hat{p}}$;
Calculate the FD $d^a_{\min} = \min_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}^a\mathbf{e}_{ij}\|^2_F$ and $d^b_{\min} = \min_{\mathbf{e}_{ij} \in E} \left\|\mathbf{H}\mathbf{P}^b\mathbf{e}_{ij}\right\|^2_F$;
find the PA matrix associated with the maximum FD $\mathbf{P}_t$ in the set $\{\mathbf{P}^a, \mathbf{P}^b\}$ and achieve $\mathbf{d}_{\min}(k) = \min_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}_t\mathbf{e}_{ij}\|^2_F$.
end
(c) Find the index $k^* = \arg\max \mathbf{d}_{\min}(k)$ and select the corresponding PA $\mathbf{P}^*$ and add the consumed power $\Delta\hat{p}$ to $\tilde{P}_u$;
(d) Compare the FD $\mathbf{d}_{\min}(k^*)$ to the FD $d^{\text{ave}}_{\min} = \min_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}_{\text{ave}}\mathbf{e}_{ij}\|^2_F$ of the equal PA scheme and select the one having the larger FD as the final result.
End

To circumvent the aforementioned challenge, a WCF-PA algorithm is proposed for our PA-aided SM scheme, which reduces the search space by focusing its efforts on the specific TA pair (m, n) associated with the FD of (5) because this is associated with the most likely error event. Then the algorithm gradually assigns the appropriate portion of power to each of these TA pairs, whereas the power values of the remaining TAs remain unchanged with respect to their initial value.

The detailed WCF-PA algorithm is summarized as in Table I. Initially, we assume that the power is equally shared by all TAs. For a given channel matrix $\mathbf{H}$, the FD $d_{\min}(\mathbf{H})$ value associated with the initial PA matrix $\mathbf{P}$ is calculated as $d^1_{\min}$ in Step 1. Furthermore, the indices of the TA pair (m, n) achieving $d^1_{\min}$ are obtained. If $m = n$, the distance $d^1_{\min}$ is

$$d^1_{\min} = \|\mathbf{h}_m\|^2_F \, \bar{d}^2_{\min} \quad q \in \{1, \ldots, N_t\} \tag{17}$$

where $\mathbf{h}_m$ is the mth column of $\mathbf{H}$, and $\bar{d}_{\min}$ is the minimum distance in the APM constellation according to the modulation order, as shown in [19]. In (13), it is plausible that the TA m has the smallest channel gain $\|\mathbf{h}_m\|_F$. In this case, we deduct some power from the TA u, which has the largest channel gain and assign it to TA m; hence, $d_{\min}(\mathbf{H})$ may be increased due to the increased power assigned to the mth TA. Here, we define the achieved PA candidate as $\mathbf{P}_{\text{temp1}}$.

If the values of m and n are not the same, the value of $d^1_{\min}$ is computed for different TAs. To increase this FD, two possible PA strategies are considered. The first deducts some power from the TA m and assigns it to TA n, whereas the second deducts power from the TA n and assigns it to TA m. Then, the resultant PA candidates of these strategies can be represented as $\mathbf{P}_{\text{temp2}}$ and $\mathbf{P}_{\text{temp3}}$, respectively. Hence, the optimal PA matrix is formulated as

$$\mathbf{P}_{\text{opt}} = \begin{cases} \underset{\mathbf{P} \in \{\mathbf{P}_{\text{ave}}, \mathbf{P}_{\text{temp1}}\}}{\arg\max} \; d_{\min}(\mathbf{H}), & \text{if } m = n \\ \underset{\mathbf{P} \in \{\mathbf{P}_{\text{ave}}, \mathbf{P}_{\text{temp2}}, \mathbf{P}_{\text{temp3}}\}}{\arg\max} \; d_{\min}(\mathbf{H}), & \text{if } m \neq n. \end{cases} \tag{18}$$

As the $d_{\min}(\mathbf{H})$ value increases throughout the WCF-PA iterations, the proposed PA scheme provides a beneficial system performance improvement compared with the conventional SM. More importantly, this algorithm has low complexity because the greedy PA philosophy is adopted only for two TAs, regardless of the total number of TAs.

C. Joint AM and PA Techniques in SM

As shown in Section II, the PA and AM techniques may rely on different transmit parameters to achieve a BER improvement. To further exploit the associated grade of freedom, our PA-aided SM systems can be combined with AM technique. However, this hybrid scheme may become excessively complex, when aiming for jointly optimizing these parameters according to the near-instantaneous channel conditions. In this paper, we simplify the computations using a multistage adaptation strategy. First, the AM technique of [17] is invoked for choosing the optimal modulation constellations for the most correlated TA pair. Then their corresponding power is allocated. Although this approach may not be optimal for joint AM-PA-based systems, we will confirm with the aid of our simulations that this multistage adaptation strategy is capable of achieving a performance improvement compared with the ASM and PA-aided SM schemes. The efficient amalgamation of

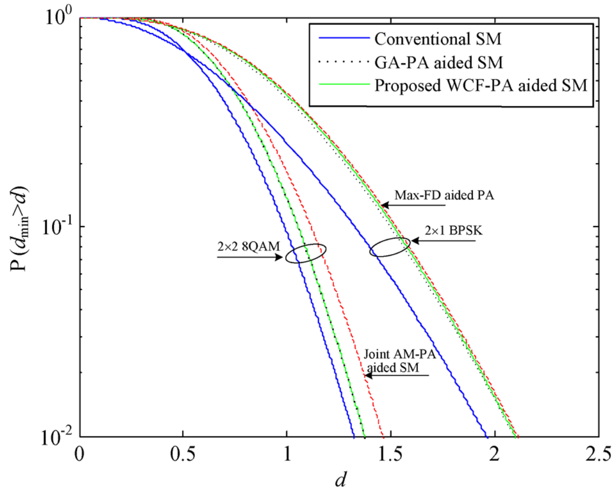

Fig. 2. Distribution of FD $d_{\min}$ of conventional and proposed PA schemes in 2×1 and 2×2 MIMO channels.

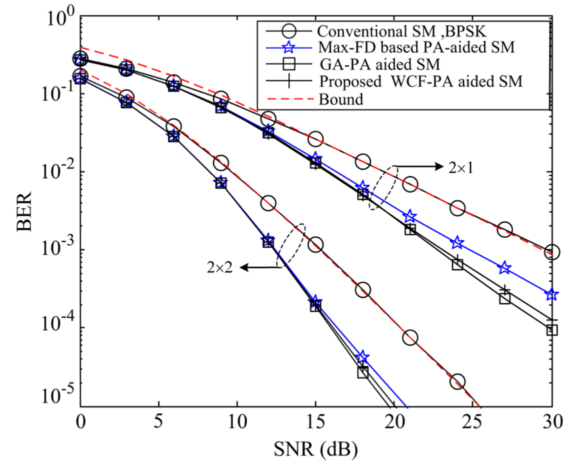

Fig. 3. BER comparison at $m_r = 2$ bits/symbol for the conventional BPSK-modulated SM and the PA-aided SM schemes.

AM and PA for the sake of exploiting all the benefits of the MIMO channels constitutes a challenging problem, which will be investigated in our further studies.

Fig. 2 portrays the complementary cumulative distribution functions of the FD recorded both for conventional SM and for the proposed PA-aided SM schemes in (2×1) and (2×2) MIMO channels. Observe in Fig. 2 that the PA-aided SM schemes are capable of beneficially increasing the FD. More specifically, as expected, the optimal max-FD-aided PA scheme has a higher FD than that of the GA-PA and the WCF-based PA schemes due to the fact that it is capable of finding the global optimal solution by using (8)–(16). Moreover, the GA-PA-aided SM achieves almost the same FD as that of the WCF-PA-aided SM. Hence, these two PA-aided schemes may achieve the same BER performance, as will be shown in Section IV. It is also shown in Fig. 2 that the joint AM-PA-aided SM may outperform the other PA-aided schemes for (2×2) MIMO channels because it has the highest FD among these PA-aided schemes.

D. Computational Complexity and Feedback Load

For each channel realization $\mathbf{H}$, the GA-PA algorithm has to conduct a full exhaustive search of the $\lceil (N_t/\Delta p) \rceil N_t$ number of PA matrix candidates, whereas the proposed WCF-PA algorithm only deals with $2\lceil (1/\Delta p) \rceil$ values. Here, the number of $d_{\min}(\mathbf{H})$ candidates to be evaluated is a good metric of quantifying the complexity of these algorithms. Moreover, in the proposed WCF-PA, we can first use the simplified calculation method of [16] and [17] for quantifying $d_{\min}(\mathbf{H})$ of the PA candidate $\mathbf{P}$. Then, the calculation of $d_{\min}(\mathbf{H})$ in Step 3 for the other candidates having the tentative PA only has to consider the updated TAs. Hence, the complexity can be further reduced. On the other hand, the GA-PA algorithm requires the receiver to feed back the index of the activated PA matrix to the transmitter, whereas the WCF-PA algorithm only has to feed back the index of the specific TA pair associated with the PA and their assigned power values.

IV. Simulation Results

Here, we evaluate the BER performance of the proposed PA-aided SM schemes over frequency-flat-fading channels. The simulation setup is based on 2–4 bits/symbol transmissions, and the number of modulated symbols is equal to $N_L = 30$ for each channel realization. For comparison, we consider the one-bit reallocation (OBRA)-ASM of [17], which is a simplified version of the ASM scheme of [16].

Fig. 3 shows the BER performance of the optimal max-FD-aided PA and the numerical search-aided PA schemes (the GA-PA-aided SM and the WCF-PA-aided SM schemes). In Fig. 3, the (2×1)-element and (2×2)-element MIMO channels using BPSK modulation are considered. For completeness, we add the theoretical upper bound of [14] for the conventional SM scheme. In Fig. 3, in the low-to-medium SNR regime, the numerical search-aided PA schemes achieve almost the same performance as the optimal max-FD-based PA-aided SM. Note that, although the optimal max-FD-based PA-aided SM is capable of achieving a higher FD than other PA-aided schemes, its number λ of the nearest neighbor constellation points may become doubled compared with the conventional SM due to the optimization process. By contrast, the numerical search-aided PA schemes, the solutions may be expected to be close to the optimal max-FD, and hence, parameter λ may not be doubled as that of the optimal max-FD algorithm. As indicated in (4), the FD and the corresponding number of nearest neighbors λ jointly determine the BER and having an increased value of λ may degrade the attainable BER performance. Hence, as shown in Fig. 3, at high SNRs, the optimal max-FD-based PA-aided SM may perform worse than numerical search-aided PA schemes associated with a lower λ[1]. To circumvent this problem, the combination of the max-FD and minimum-λ in PA-aided SM may be adopted, which has high complexity, as indicated in [23]. Moreover, we observe in Fig. 3 that the low-complexity WCF-PA-aided scheme attains a similar BER performance to that of the exhaustive-search-based GA-PA scheme.

In Fig. 4, the QPSK-modulated VBLAST scheme and its PA-aided counterpart associated with a zero-forcing successive interference cancelation (ZF-SIC) detector [20] are compared with our PA-aided schemes because their detection complexity values are similar [1], [5]. Observe in Fig. 4 for $m_r = 4$ that our PA-aided SM schemes outperform the PA-aided VBLAST arrangements relying on a ZF-SIC detector. Indeed, if a powerful ML detector is employed for the VBLAST system, we can achieve a better BER performance. However, designing PA algorithms for ML-based VBLAST systems is a challenge, and their detection complexity is high, as indicated in [21] and [22].

[1]Another reason for this result is that the max-FD-aided PA may achieve a lower Euclidean distance between the nonadjacent received constellation points than that of the WCF-PA and GA-PA schemes. Hence, based on the Q-function-aided PEP upper bound of [19], which depends on all received distances $d_{ij}(\mathbf{H}) = \|\mathbf{HP}(\mathbf{x}_i - \mathbf{x}_j)\|_F (i \neq j)$ of the received constellation points, the max-FD-based PA may not achieve the minimum BER performance compared with that of other PA schemes.

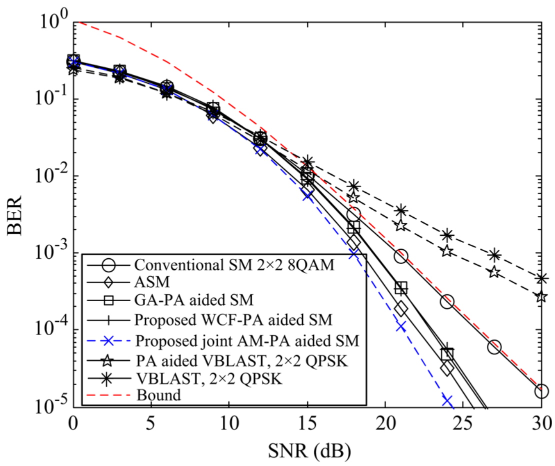

Fig. 4. BER comparison at $m_r = 4$ bits/symbol for the conventional 8-QAM-modulated SM and the PA-aided SM scheme in 2×2 MIMO channels. The corresponding BER results of the QPSK-modulated VBLAST scheme and its PA-aided scheme are calculated as the benchmarkers.

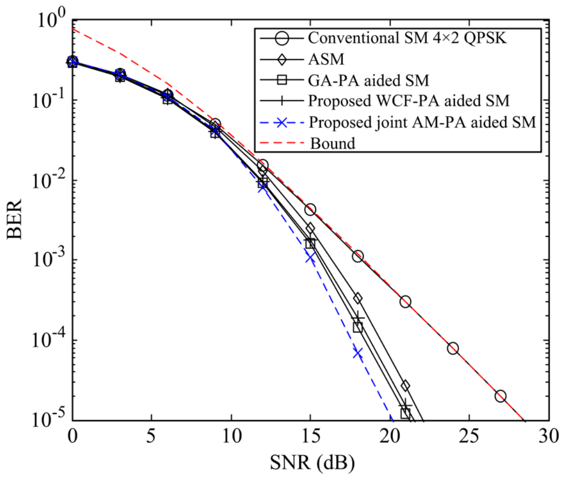

Fig. 5. BER comparison at $m_r = 4$ bits/symbol for the conventional QPSK-modulated SM, the ASM scheme, the proposed GA-PA-aided SM, the proposed WCF-PA SM, and the joint AM-PA-aided SM. Here, 4×2 MIMO channels are considered.

As indicated in Section II-C, our PA algorithm is to a degree reminiscent of the spatiotemporal PA scheme of [24], which is capable of achieving exponential diversity. However, they are different in the sense that the PA scheme in [24] relies on the long-term (time) average over which it maintains constant total power, whereas our scheme relies on the selection probability of the TAs for satisfying the total power constraint. Owing to this difference, the analysis method of [24] cannot be extended to our PA-aided SM scheme. However, since the power is allocated to both the spatial and temporal dimensions in our PA algorithm, more substantial performance gains may be expected than that of the pure spatial-domain PA-aided VBLAST scheme, as shown in Fig. 4. Deriving the explicit diversity order of this max-FD-based PA algorithm is based on the distribution of the FD distance. Since this FD depends both on the constellation and on the channel realization, its distribution is difficult to determine. Hence, the explicit diversity order of this max-FD-based PA algorithm is hard to characterize analytically. This challenge is also an open problem in the max-FD-aided PA algorithm of VBLAST [21], [22]. Nonetheless, the aforementioned challenge will be considered in our further research.

Fig. 5 compares the BER performances of the PA-aided schemes and of the conventional ASM schemes for $N_t = 4$. Upon comparing

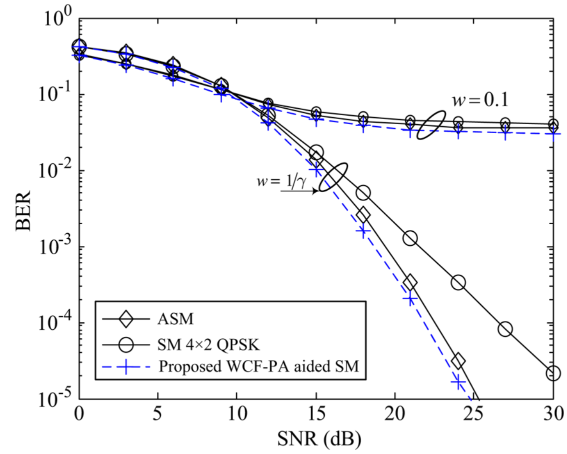

Fig. 6. BER comparison of the QPSK-aided SM, the ASM, and the proposed WCF-PA-aided SM schemes under 4×2 MIMO channels. We also considered the effects of CSI error associated with an equivalent channel estimation's noise variance of $w = 0.1$ and $(1/\gamma)$.

the results in Fig. 5 with the results in Fig. 4, we observe that the AM-aided and PA-assisted SM schemes exhibit different BER advantages for different numbers of TAs. This is because these techniques exploit different properties of the MIMO channels when aiming to maximize the FD, as indicated in Section II. As expected in Figs. 4 and 5, we observe that joint AM-PA-aided SM achieves the best BER performance among all the schemes.

Fig. 6 compares the achievable BER performance of the WCF-PA-aided SM in the presence of Gaussian-distributed channel state information (CSI) errors obeying $\mathcal{CN}(0, w)$ [7], [8] associated with $w = 0.1$ and $(1/\gamma)$, where γ is the average CSI estimation SNR at each receiver antenna. Observe in Fig. 6 that the BER performance of WCF-PA-aided SM is degraded upon introducing CSI estimation errors. However, this PA-based scheme still provides a considerable performance improvement over its nonadaptive counterparts with $w = 1/\gamma$.

V. Conclusion

In this paper, we have proposed the PA algorithms designed for limited-feedback SM-MIMO systems. Our simulation results confirm that the achievable performance is quite attractive. Our further work will be focused on the integration of space–time coding, channel coding, and space–time-shift keying [5] into the proposed schemes.

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Power Allocation-Aided Spatial Modulation for Limited-Feedback MIMO Systems

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***Abstract*—Adaptive power allocation (PA) algorithms based on optimization of the minimum distance $d_{\min}$ between signal points at the receiver side are investigated in spatial modulation (SM) systems. First, a closed-form solution of the optimal maximum-$d_{\min}$-aided PA algorithm is derived in the case of two transmit antennas (TAs). Moreover, for a higher number of TAs, we propose a numerical approach, which appropriately splits the power between the specific TA pair associated with $d_{\min}$ to increase this distance. Furthermore, our PA-aided SM systems may be readily combined with adaptive modulation (AM) to further improve the system's performance. Our numerical results show that the proposed algorithms provide beneficial system performance improvements compared with both the equal-gain PA-based SM and the identical-throughput PA-aided spatial multiplexing systems.**

***Index Terms*—Limited-feedback systems, multiple-input–multiple-output (MIMO), power allocation (PA), spatial modulation (SM).**

I. Introduction

Spatial modulation (SM), which maps information both to a carefully designed combination of antenna indices and to the classic amplitude and phase modulation (APM) constellation, constitutes a promising low-complexity multiple-input–multiple-output (MIMO) transmission technique [1]–[5]. The SM-based systems are capable of outperforming some of the classic MIMO techniques [6] even in the presence of channel estimation errors; however, they can only offer receive diversity [7], [8].

To overcome this problem, link adaptation (LA) schemes have been proposed in [9]–[15], where the transmit parameters are dynamically adapted to the channel conditions. Specifically, the effects of power imbalance [9], the issues of achieving transmit diversity [10], the particular choice of the constellation used [11], and the impact of cooperation have been researched [12]. However, most of the aforementioned LA schemes considered only a special case of SM, i.e., space-shift keying [3], which exclusively employs the antenna indices for data modulation. In [14], the Kronecker model was used to characterize a correlated SM-MIMO channel, and a beamforming codebook design algorithm was proposed to optimize the bit-error-ratio (BER) performance based on the rather limited knowledge of the channel envelope's spatial correlation. In [15], a power-scaling-assisted SM scheme was proposed, where a scaling factor (SF) was invoked for weighting the modulated symbols before their transmission. However, the related design algorithm of SF was not provided. Recently, we have proposed an adaptive modulation (AM)-aided SM (ASM) scheme [16], [17] to improve the attainable system performance. In ASM, the receiver requests the most suitable modulation order to be used by the transmitter for each transmit antenna (TA). However, constant-power ASM may not be capable of fully exploiting the available spatial-domain grade of freedom offered by MIMO channel.

Power allocation (PA) techniques are capable of alleviating the adverse effects of channel fading to achieve either an increased data rate or a reduced BER. Indeed, PA has been lavishly researched in the context of spatial multiplexing systems [18], [19]. As a new MIMO technique, SM may be also beneficially combined with PA for adjusting the transmission parameters for the sake of accommodating time-varying channels. However, since only a single TA is activated in each time slot in SM-based schemes, the PA approaches designed for spatial multiplexing-based MIMO systems may not be directly suitable for SM systems.

Against this background, the novel contributions of this paper are as follows.

- We investigate the benefits of adaptive PA based on the maximum-free distance (max-FD) $d_{\min}$ between the pairs of signal constellation points at the receiver side. An optimal $d_{\min}$ PA precoder is derived for BPSK-modulated (2×1)-element SM. Then, this result is extended to M-PSK/M-ary quadrature amplitude modulation (M-QAM) $(2 \times N_r)$-element PA-aided SM.
- To deal with the case of $N_t > 2$, we propose a numerical approach, termed as the "worse-case-first"-based PA (WCF-PA) algorithm, which appropriately splits the power between the specific TA pair associated with $d_{\min}$. As a further benefit, our PA-aided SM systems may be readily combined with AM techniques for the sake of maximizing the FD, hence improving the system's performance.

The organization of this paper is as follows. Section II presents the system model of the PA-aided SM. In Section III, we introduce our PA algorithms, whereas our simulation results and performance comparisons are presented in Section IV. Finally, Section V concludes this paper.

Notation: $(\cdot)^*$, $(\cdot)^T$, and $(\cdot)^H$ denote conjugate, transpose, and Hermitian transpose, respectively. Furthermore, $\|\cdot\|$ stands for the Frobenius norm, and $Re\{\cdot\}$ denotes the real part of a complex variable. $\lceil x \rceil$ denotes the smallest integer higher than or equal to x.

II. System Model

A. Transceiver

Let us consider a flat-fading MIMO channel associated with N_t TAs and N_r receiver antennas, represented by an $(N_r \times N_t)$-element matrix $\mathbf{H}$. The entries of $\mathbf{H}$ are assumed to be independent identically

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distributed and obeying $h_{vu} \sim \mathcal{CN}(0,1)$. Then, the transmitted PA-aided SM symbol $\mathbf{x} \in \mathbb{C}^{N_t \times 1}$ is given as

$$\mathbf{x} = s_l^q \mathbf{e}_q = [0, \ldots, \underset{\uparrow\, qth}{s_l^q}, \ldots, 0]^T \tag{1}$$

where s_l^q is the APM symbol assigned to the qth TA, such as r^q-QAM, which is associated with $d^q = \log_2(r^q)$ input bits, whereas $\mathbf{e}_q \in \mathbb{C}^{N_t \times 1}$ is selected from the N_t-dimensions standard basis vectors (i.e., $\mathbf{e}_1 = [1, 0, \ldots, 0]^T$), according to the $\log_2(N_t)$ input bits.

At the receiver, the corresponding $(N_r \times 1)$-element received signal vector is given by

$$\mathbf{y} = \mathbf{HPx} + \mathbf{n} \tag{2}$$

where the elements of the N_r-element noise vector $\mathbf{n}$ are Gaussian random variables obeying $\mathcal{CN}(0, N_0)$, and the diagonal matrix $\mathbf{P}$ allocates the total power P_T to the different TAs, yielding

$$\mathbf{P} = \begin{pmatrix} \sqrt{p_1} & & 0 \\ & \ddots & \\ 0 & & \sqrt{pN_t} \end{pmatrix} \tag{3}$$

where the diagonal elements are limited by the total power constraint as $\sum_{q=1}^{N_t} p_q = P_T$.

B. Performance Metric

Given the channel matrix $\mathbf{H}$, the pairwise error probability (PEP) of the SM system using the maximum-likelihood (ML) detector is given by [19]

$$P(\mathbf{x}_i \to \mathbf{x}_j | \mathbf{H}) \approx \lambda \cdot Q\left(\sqrt{\frac{1}{2N_0} d_{\min}(\mathbf{H})}\right) \tag{4}$$

where $Q(x) = (1/\sqrt{2\pi}) \int_x^\infty e^{-(y^2/2)} dy$, and λ is the number of neighbor constellation points [19] having the FD $d_{\min}(\mathbf{H})$, which is defined as

$$\begin{aligned} d_{\min}(\mathbf{H}) &= \min_{\substack{\mathbf{x}_i, \mathbf{x}_j \in \mathbb{X}, \\ \mathbf{x}_i \neq \mathbf{x}_j}} \|\mathbf{HP}(\mathbf{x}_i - \mathbf{x}_j)\|_F^2 \\ &= \min_{\mathbf{e}_{ij} \in \mathbb{E}} \|\mathbf{HPe}_{ij}\|_F^2 \end{aligned} \tag{5}$$

where $\mathbb{X}$ is the set of all legitimate transmit symbols, whereas $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$ denotes the error vector and $\mathbb{E}$ is a set of error vectors. The complexity for calculating the FD of our SM-based system in (5) is relatively modest because of the following two reasons.

1) In (5), the collinear distance vectors generate the same distance. Hence, although the cardinality of the set $\mathbb{E}$ may be large, only a representative subset of those collinear error vectors has to be considered.
2) Unlike in conventional MIMO techniques, the transmit vectors of SM-MIMO schemes are sparsely populated since they have mostly zero values. This property can be exploited to further reduce the complexity of evaluating (5), as detailed in [17].

C. Optimization Criterion

Since the error events mainly arise from the nearest neighbors, the maximization of the FD in (5) directly reduces the probability of error, particularly at high-signal-to-noise ratios (SNRs) [19]. Note that the PA solution based on the FD is an attractive LA regime, which has been vastly researched in the context of spatial multiplexing systems. However, these PA approaches designed for spatial multiplexing-based MIMO systems may not be directly suitable for SM systems [20]–[22] because only a single TA is activated in each time slot. Based on (5), we propose a PA-aided SM system, which adapts the PA matrix $\mathbf{P}$ to maximize the FD under the transmit power constraint as

$$\begin{aligned} \mathbf{P}_{\text{opt}} &= \arg\max_{\mathbf{P}} d_{\min}(\mathbf{H}) \\ \text{s.t.} \quad & tr(\mathbf{P}^2) = P_T \end{aligned} \tag{6}$$

where $tr(\cdot)$ denotes the matrix trace.

In most of the PA algorithms conceived for VBLAST, the power is shared among the different TAs (space-only PA). This principle has been also adopted in our PA schemes, and hence, a fixed total power constraint is imposed on all TAs in (6). Since only a single TA is activated in each time slot in SM schemes, unlike in the PA algorithms designed for VBLAST, the PA of our proposed SM scheme beneficially exploits the time domain for maintaining the total power constraint. Under the assumption that all TAs are selected for transmission with equal probability, the average of the transmit power is fixed.

Moreover, it is noted that there are two main differences of ASM [16] and the proposed PA-aided SM schemes.

- They exploit different properties of the MIMO channels. Specifically, ASM dynamically adapts the modulation order assigned to TAs, whereas the PA-aided SM adapts the power assigned to these antennas.
- The modulation orders of ASM are selected from a discrete set, whereas the PA parameters are chosen from the real-valued field.

III. PA Algorithms

Here, we first derive a closed-form solution of (6) for BPSK-modulated (2×1)-element SM and then extend the method to the more general M-PSK/M-QAM modulated $(2 \times N_r)$-element PA-aided SM scenario. In the case of large TAs and high modulation order, deriving a closed-form solution to (6) remains an open challenge since the solution is obtained by exhaustive search for large search space constituted by all legitimate candidate transmit symbols or error vectors. Hence, a numerical search method is proposed.

A. Optimal-FD PA Matrix for a BPSK-modulated 2 × 1 SM

For BPSK-modulated SM associated with $N_t = 2$ and $N_r = 1$, the symbols belong to the set $\{1, -1\}$, and all possible error vectors $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$ are listed as follows: $\{[-2, 0]^T, [2, 0]^T, [0, -2]^T, [0, 2]^T, [-1, 1]^T, [-1, -1]^T, [1, -1]^T$, and $[1, 1]^T\}$. Since some of the vectors are collinear, the set to be studied is reduced to $\{\mathbf{e}_1, \mathbf{e}_2, \mathbf{e}_3, \mathbf{e}_4\} = \{[2, 0]^T, [0, 2]^T, [1, -1]^T, [1, 1]^T\}$. Given the (1×2) channel matrix $\mathbf{H} = [h_1, h_2]$, the received constellation point distances are given by

$$\begin{cases} d_1 = \|\mathbf{HPe}_1\|^2 = 4p_1 \|h_1\|^2 \\ d_2 = \|\mathbf{HPe}_2\|^2 = 4p_2 \|h_2\|^2 \\ d_3 = \|\mathbf{HPe}_3\|^2 = \left\|\sqrt{p_1} h_1 - \sqrt{p_2} h_2\right\|^2 \\ d_4 = \|\mathbf{HPe}_4\|^2 = \left\|\sqrt{p_1} h_1 + \sqrt{p_2} h_2\right\|^2. \end{cases} \tag{7}$$

Based on (7), the optimization problem of (6) can be simplified to

$$\begin{aligned} \mathbf{P}_{\text{opt}} &= \arg\max_{\mathbf{P}} \{\min\{d_1, d_2, d_3, d_4\}\} \\ \text{s.t.} \quad & p_1 + p_2 = P_T. \end{aligned} \tag{8}$$

As indicated in (8) and shown in Fig. 1, d_1 and d_2 are linear functions of parameter p_1, whereas d_3 and d_4 are convex or concave

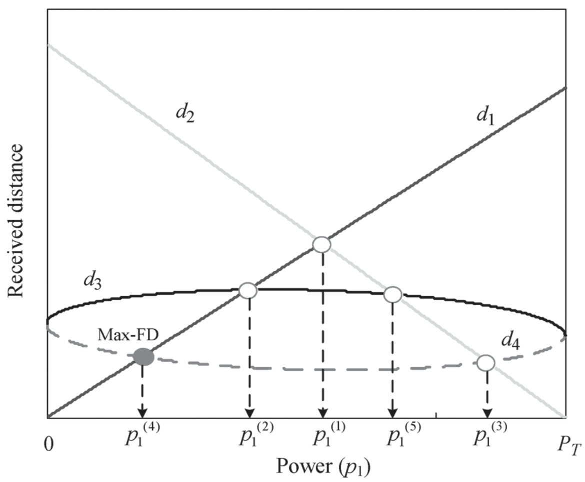

Fig. 1. Optimal-FD-aided PA for BPSK-modulated 2×1 SM.

functions and represented by an ellipse as a function of the power p_1. Hence, the max-FD solution according to p_1 is one of the intersections between these received distances $d_i (i = 1,2,3,4)$. More specifically, we can obtain the power assigned to the TA by finding these intersections and then selecting the one having the maximum FD as the final solution. As a result, the received FD is maximized, and then, the error performance is improved. To be specific, observe from (8) and for the total power constraint that the power $p_1^{(1)}$ associated with the first intersection in Fig. 1 satisfies

$$\begin{cases} d_1 = 4p_1^{(1)} \left\|h_1\right\|^2 = d_2 = 4p_2^{(1)} \left\|h_2\right\|^2 \\ p_1^{(1)} + p_2^{(1)} = p_T. \end{cases} \tag{9}$$

Upon introducing the shorthand of $a = \|h_1\|^2$, $b = \|h_2\|^2$, and $c = h_1^* h_2 + h_2^* h_1 = 2Re\{h_1^* h_2\}$ for a given channel matrix $\mathbf{H}$ and using (9), we obtain

$$\begin{cases} p_1^{(1)} = b/(a+b)P_T \\ p_2^{(1)} = a/(a+b)P_T. \end{cases} \tag{10}$$

Then, the power $p_1^{(2)}$ associated with second intersection of $d_1 = d_3$ in Fig. 1 is given by

$$\begin{cases} 4p_1^{(2)} \left\|h_1\right\|^2 = \left\| \sqrt{p_1^{(2)}} h_1 - \sqrt{p_2^{(2)}} h_2 \right\|^2 \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{11}$$

To elaborate a little further, (11) can be simplified to

$$\begin{cases} 3ap_1^{(2)} - bp_2^{(2)} + c\sqrt{p_1^{(2)}}\sqrt{p_2^{(2)}} = 0 \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{12}$$

Then, (12) can be viewed as a quadratic function of $\sqrt{p_1^{(2)}}$, which can be solved

$$\begin{cases} \sqrt{p_1^{(2)}} = \frac{-c+\sqrt{c^2+12ab}}{6a} \sqrt{p_2^{(2)}} \\ p_1^{(2)} + p_2^{(2)} = p_T. \end{cases} \tag{13}$$

From (13), we can then evaluate the power as

$$\begin{cases} p_1^{(2)} = \frac{c^2+6ab-c\sqrt{c^2+12ab}}{18a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T \\ p_2^{(2)} = \frac{18a^2}{18a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{14}$$

Similar to the evaluation process of $p_1^{(2)}$, we can obtain the candidate power $p_1^{(3)}$ associated with $d_1 = d_4$, the power $p_1^{(4)}$ associated with $d_2 = d_3$, and the power $p_1^{(5)}$ associated with $d_2 = d_4$ step by step, which are given by

$$\begin{cases} p_1^{(3)} = \frac{c^2+6ab+c\sqrt{c^2+12ab}}{18a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_1^{(4)} = \frac{c^2+6ab+c\sqrt{c^2+12ab}}{2a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_1^{(5)} = \frac{c^2+6ab-c\sqrt{c^2+12ab}}{2a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{15}$$

Then, based on the fixed total power constraint, the corresponding power assigned to the second TA is given by

$$\begin{cases} p_2^{(3)} = \frac{18a^2}{18a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_2^{(4)} = \frac{2a^2}{2a^2+c^2+6ab+c\sqrt{c^2+12ab}} P_T \\ p_2^{(5)} = \frac{2a^2}{2a^2+c^2+6ab-c\sqrt{c^2+12ab}} P_T. \end{cases} \tag{16}$$

Additionally, the solution $p_1^{(6)}$ associated with $d_3 = d_4$ satisfies $p_1^{(6)} p_2^{(6)} = 0$. Since the activation of the TAs conveys the information bits, the PA solution of $p_1^{(6)} = 0$ or $p_1^{(6)} = P_T$ $(p_2^{(6)} = 0)$ is not considered as a legitimate one. In conclusion of the algorithm, the FDs of these PA solutions are generated, and we select the one having the largest FD as our final result. Next, the aforementioned method is extended to M-PSK/M-QAM modulated PA-aided $(2 \times N_r)$-element SM. Here, the value of N_r is an arbitrary positive integer.

The detailed max-FD-aided PA algorithm is summarized in two steps as follows.

- Step 1: Compute all legitimate error vectors $\mathbf{e}_{ij} = \mathbf{x}_i - \mathbf{x}_j, i \neq j$, and eliminate the redundant collinear elements. Calculate all legitimate received constellation distances $d_l (1 = 1, \ldots, L)$ with the aid of the channel matrix $\mathbf{H}$ and $\mathbf{e}_{ij}$, which are either linear or nonlinear but convex functions of power p_1.
- Step 2: Find all possible intersections between the received constellation distances d_i and d_j $(i, j \in \{1, \ldots, L\})$, and calculate both the corresponding power matrix $P = \text{diag}\{\sqrt{p_1}, \sqrt{p_2}\}$ and the corresponding FD. Select the one having the largest FD as our final result.

Therefore, the allocated power to TA can be decided as a closed-form solution by the aforementioned steps with low complexity. Note AQ1 that the restriction to $2 \times N_r$-element SM is imposed by the difficulty of the FD optimization, and the solution of the general problem remains an open challenge. Indeed, the determination of the PA matrix that maximizes the FD of (5) is difficult for two reasons: First, the solution depends on both the channel matrix and on the symbol alphabet, and the space of solutions is excessive. Hence, for a higher throughput, we propose a simple numerical approach for this difficult optimization problem.

B. WCF-PA

To deal with the case of $N_t > 2$ and high modulation orders, the conventional greedy algorithm-based PA (GA-PA) of [17] can be further developed for our SM systems. To be specific, at each step of the GA-PA algorithm, a small fraction Δp of the total power is allocated to that specific TA, which maximizes the FD. By contrast, the power of all the other TAs remains unchanged. As the total power P_T is gradually allocated, the final PA matrix $\mathbf{P}$ is approached. However, the GA-PA algorithm has to tentatively allocate power to all possible TAs in each iteration, which imposes high complexity.

TABLE I
PROPOSED WCF-PA ALGORITHM

Step1: Initialize $\mathbf{P}$ as a diagonal matrix with equal values $p_q = 1$, $q = 1, \cdots, N_t$.
Set the PA granularity Δp to a small number divisible by $\tilde{P}_T = 1$, such as $\Delta p = 0.05\tilde{P}_T$.
Let the consumed power of a single antenna be $\tilde{P}_u$. Set $k = 0$, $\mathbf{P}_{\text{ave}} = \mathbf{P}$, $\mathbf{P}^a = \mathbf{P}$ and $\mathbf{P}^b = \mathbf{P}$.
Set all entries of the $(N_t \times 1)$-element FD vectors $\mathbf{d}_{\min}(q)$ to zero.

Step2: calculate the $d_{\min}(\mathbf{H})$ of the PA matrix $\mathbf{P}$ as $d_{\min}^1$ and find the indices of the transmit antenna pair (m, n) achieved the value $d_{\min}^1$.

Step3: gradual power allocation between two TAs.
(a) If $m = n$, the distance $d_{\min}^1$ is achieved by one of the TAs, whereas two different signal constellation points are adopted.
while $\tilde{P}_u < \tilde{P}_T$,
Let $\Delta\hat{p} = \min\{\Delta p, \tilde{P}_T - \tilde{P}_u\}$ and $k = k + 1$;
Compute the norm $\|\mathbf{h}_q\|_F$ for $q = 1, \cdots, N_t$, find the TA u associated with the maximum norm as $u = \arg\max\limits_q \|\mathbf{h}_q\|_F$ and update $p_u = \sqrt{1 - \Delta\hat{p}}$.
Then find the column m associated with the minimum norm as $m = \arg\min\limits_q \|\mathbf{h}_q\|_F$ and update $p_m = \sqrt{1 + \Delta\hat{p}}$. Calculate the corresponding FD of $\mathbf{d}_{\min}(k) = \min\limits_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}\mathbf{e}_{ij}\|_F^2$.
end
(b) elseif $m \neq n$, $d_{\min}^1$ is computed for different TAs.
while $\tilde{P}_u < \tilde{P}_T$,
Let $\Delta\hat{p} = \min\{\Delta p, \tilde{P}_T - \tilde{P}_u\}$ and $k = k + 1$;
Update $\mathbf{P}^a(n, n) = \sqrt{1 - \Delta\hat{p}}$ and $\mathbf{P}^a(m, m) = \sqrt{1 + \Delta\hat{p}}$;
Update $\mathbf{P}^b(n, n) = \sqrt{1 + \Delta\hat{p}}$ and $\mathbf{P}^b(m, m) = \sqrt{1 - \Delta\hat{p}}$;
Calculate the FD $d_{\min}^a = \min\limits_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}^a\mathbf{e}_{ij}\|_F^2$ and $d_{\min}^b = \min\limits_{\mathbf{e}_{ij} \in E} \left\|\mathbf{H}\mathbf{P}^b\mathbf{e}_{ij}\right\|_F^2$;
find the PA matrix associated with the maximum FD $\mathbf{P}_t$ in the set $\{\mathbf{P}^a, \mathbf{P}^b\}$ and achieve $\mathbf{d}_{\min}(k) = \min\limits_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}_t\mathbf{e}_{ij}\|_F^2$.
end
(c) Find the index $k^* = \arg\max \mathbf{d}_{\min}(k)$ and select the corresponding PA $\mathbf{P}^*$ and add the consumed power $\Delta\hat{p}$ to $\tilde{P}_u$;
(d) Compare the FD $\mathbf{d}_{\min}(k^*)$ to the FD $d_{\min}^{\text{ave}} = \min\limits_{\mathbf{e}_{ij} \in E} \|\mathbf{H}\mathbf{P}_{\text{ave}}\mathbf{e}_{ij}\|_F^2$ of the equal PA scheme and select the one having the larger FD as the final result.
End

To circumvent the aforementioned challenge, a WCF-PA algorithm is proposed for our PA-aided SM scheme, which reduces the search space by focusing its efforts on the specific TA pair (m, n) associated with the FD of (5) because this is associated with the most likely error event. Then the algorithm gradually assigns the appropriate portion of power to each of these TA pairs, whereas the power values of the remaining TAs remain unchanged with respect to their initial value.

The detailed WCF-PA algorithm is summarized as in Table I. Initially, we assume that the power is equally shared by all TAs. For a given channel matrix $\mathbf{H}$, the FD $d_{\min}(\mathbf{H})$ value associated with the initial PA matrix $\mathbf{P}$ is calculated as $d_{\min}^1$ in Step 1. Furthermore, the indices of the TA pair (m, n) achieving $d_{\min}^1$ are obtained. If $m = n$, the distance $d_{\min}^1$ is

$$d_{\min}^1 = \|\mathbf{h}_m\|_F^2 \, \bar{d}_{\min}^2 \quad q \in \{1, \ldots, N_t\} \tag{17}$$

where $\mathbf{h}_m$ is the mth column of $\mathbf{H}$, and $\bar{d}_{\min}$ is the minimum distance in the APM constellation according to the modulation order, as shown in [19]. In (13), it is plausible that the TA m has the smallest channel gain $\|\mathbf{h}_m\|_F$. In this case, we deduct some power from the TA u, which has the largest channel gain and assign it to TA m; hence, $d_{\min}(\mathbf{H})$ may be increased due to the increased power assigned to the mth TA. Here, we define the achieved PA candidate as $\mathbf{P}_{\text{temp1}}$.

If the values of m and n are not the same, the value of $d_{\min}^1$ is computed for different TAs. To increase this FD, two possible PA strategies are considered. The first deducts some power from the TA m and assigns it to TA n, whereas the second deducts power from the TA n and assigns it to TA m. Then, the resultant PA candidates of these strategies can be represented as $\mathbf{P}_{\text{temp2}}$ and $\mathbf{P}_{\text{temp3}}$, respectively. Hence, the optimal PA matrix is formulated as

$$\mathbf{P}_{\text{opt}} = \begin{cases} \arg\max\limits_{\mathbf{P} \in \{\mathbf{P}_{\text{ave}}, \mathbf{P}_{\text{temp1}}\}} d_{\min}(\mathbf{H}), & \text{if } m = n \\ \arg\max\limits_{\mathbf{P} \in \{\mathbf{P}_{\text{ave}}, \mathbf{P}_{\text{temp2}}, \mathbf{P}_{\text{temp3}}\}} d_{\min}(\mathbf{H}), & \text{if } m \neq n. \end{cases} \tag{18}$$

As the $d_{\min}(\mathbf{H})$ value increases throughout the WCF-PA iterations, the proposed PA scheme provides a beneficial system performance improvement compared with the conventional SM. More importantly, this algorithm has low complexity because the greedy PA philosophy is adopted only for two TAs, regardless of the total number of TAs.

C. Joint AM and PA Techniques in SM

As shown in Section II, the PA and AM techniques may rely on different transmit parameters to achieve a BER improvement. To further exploit the associated grade of freedom, our PA-aided SM systems can be combined with AM technique. However, this hybrid scheme may become excessively complex, when aiming for jointly optimizing these parameters according to the near-instantaneous channel conditions. In this paper, we simplify the computations using a multistage adaptation strategy. First, the AM technique of [17] is invoked for choosing the optimal modulation constellations for the most correlated TA pair. Then their corresponding power is allocated. Although this approach may not be optimal for joint AM-PA-based systems, we will confirm with the aid of our simulations that this multistage adaptation strategy is capable of achieving a performance improvement compared with the ASM and PA-aided SM schemes. The efficient amalgamation of

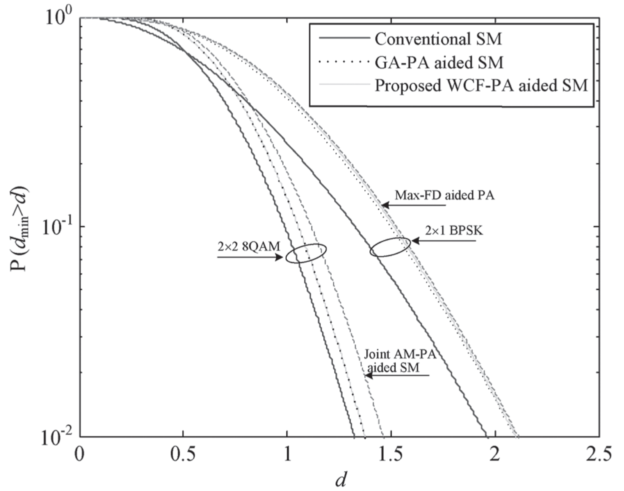

Fig. 2. Distribution of FD $d_{\min}$ of conventional and proposed PA schemes in 2×1 and 2×2 MIMO channels.

AM and PA for the sake of exploiting all the benefits of the MIMO channels constitutes a challenging problem, which will be investigated in our further studies.

Fig. 2 portrays the complementary cumulative distribution functions of the FD recorded both for conventional SM and for the proposed PA-aided SM schemes in (2×1) and (2×2) MIMO channels. Observe in Fig. 2 that the PA-aided SM schemes are capable of beneficially increasing the FD. More specifically, as expected, the optimal max-FD-aided PA scheme has a higher FD than that of the GA-PA and the WCF-based PA schemes due to the fact that it is capable of finding the global optimal solution by using (8)–(16). Moreover, the GA-PA-aided SM achieves almost the same FD as that of the WCF-PA-aided SM. Hence, these two PA-aided schemes may achieve the same BER performance, as will be shown in Section IV. It is also shown in Fig. 2 that the joint AM-PA-aided SM may outperform the other PA-aided schemes for (2×2) MIMO channels because it has the highest FD among these PA-aided schemes.

D. Computational Complexity and Feedback Load

For each channel realization $\mathbf{H}$, the GA-PA algorithm has to conduct a full exhaustive search of the $\lceil (N_t/\Delta p) \rceil N_t$ number of PA matrix candidates, whereas the proposed WCF-PA algorithm only deals with $2\lceil (1/\Delta p) \rceil$ values. Here, the number of $d_{\min}(\mathbf{H})$ candidates to be evaluated is a good metric of quantifying the complexity of these algorithms. Moreover, in the proposed WCF-PA, we can first use the simplified calculation method of [16] and [17] for quantifying $d_{\min}(\mathbf{H})$ of the PA candidate $\mathbf{P}$. Then, the calculation of $d_{\min}(\mathbf{H})$ in Step 3 for the other candidates having the tentative PA only has to consider the updated TAs. Hence, the complexity can be further reduced. On the other hand, the GA-PA algorithm requires the receiver to feed back the index of the activated PA matrix to the transmitter, whereas the WCF-PA algorithm only has to feed back the index of the specific TA pair associated with the PA and their assigned power values.

IV. Simulation Results

Here, we evaluate the BER performance of the proposed PA-aided SM schemes over frequency-flat-fading channels. The simulation setup is based on 2–4 bits/symbol transmissions, and the number of modulated symbols is equal to $N_L = 30$ for each channel realization. For comparison, we consider the one-bit reallocation (OBRA)-ASM of [17], which is a simplified version of the ASM scheme of [16].

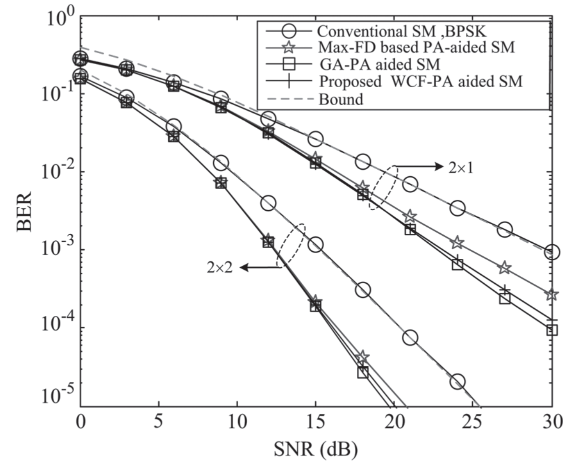

Fig. 3. BER comparison at $m_r = 2$ bits/symbol for the conventional BPSK-modulated SM and the PA-aided SM schemes.

Fig. 3 shows the BER performance of the optimal max-FD-aided PA and the numerical search-aided PA schemes (the GA-PA-aided SM and the WCF-PA-aided SM schemes). In Fig. 3, the (2×1)-element and (2×2)-element MIMO channels using BPSK modulation are considered. For completeness, we add the theoretical upper bound of [14] for the conventional SM scheme. In Fig. 3, in the low-to-medium SNR regime, the numerical search-aided PA schemes achieve almost the same performance as the optimal max-FD-based PA-aided SM. Note that, although the optimal max-FD-based PA-aided SM is capable of achieving a higher FD than other PA-aided schemes, its number λ of the nearest neighbor constellation points may become doubled compared with the conventional SM due to the optimization process. By contrast, the numerical search-aided PA schemes, the solutions may be expected to be close to the optimal max-FD, and hence, parameter λ may not be doubled as that of the optimal max-FD algorithm. As indicated in (4), the FD and the corresponding number of nearest neighbors λ jointly determine the BER and having an increased value of λ may degrade the attainable BER performance. Hence, as shown in Fig. 3, at high SNRs, the optimal max-FD-based PA-aided SM may perform worse than numerical search-aided PA schemes associated with a lower λ[1]. To circumvent this problem, the combination of the max-FD and minimum-λ in PA-aided SM may be adopted, which has high complexity, as indicated in [23]. Moreover, we observe in Fig. 3 that the low-complexity WCF-PA-aided scheme attains a similar BER performance to that of the exhaustive-search-based GA-PA scheme.

In Fig. 4, the QPSK-modulated VBLAST scheme and its PA-aided counterpart associated with a zero-forcing successive interference cancelation (ZF-SIC) detector [20] are compared with our PA-aided schemes because their detection complexity values are similar [1], [5]. Observe in Fig. 4 for $m_r = 4$ that our PA-aided SM schemes outperform the PA-aided VBLAST arrangements relying on a ZF-SIC detector. Indeed, if a powerful ML detector is employed for the VBLAST system, we can achieve a better BER performance. However, designing PA algorithms for ML-based VBLAST systems is a challenge, and their detection complexity is high, as indicated in [21] and [22].

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[1]Another reason for this result is that the max-FD-aided PA may achieve a lower Euclidean distance between the nonadjacent received constellation points than that of the WCF-PA and GA-PA schemes. Hence, based on the Q-function-aided PEP upper bound of [19], which depends on all received distances $d_{ij}(\mathbf{H}) = \|\mathbf{HP}(\mathbf{x}_i - \mathbf{x}_j)\|_F (i \neq j)$ of the received constellation points, the max-FD-based PA may not achieve the minimum BER performance compared with that of other PA schemes.

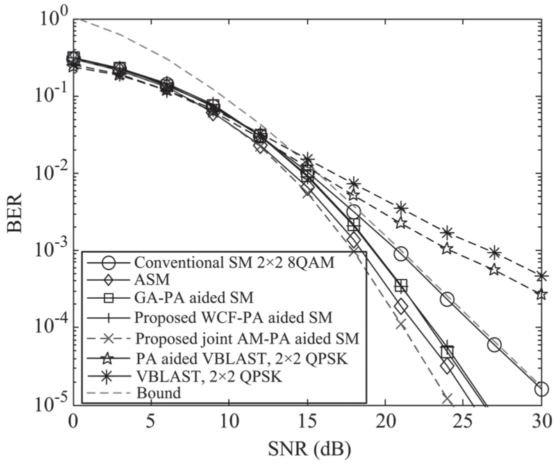

Fig. 4. BER comparison at $m_r = 4$ bits/symbol for the conventional 8-QAM-modulated SM and the PA-aided SM scheme in 2×2 MIMO channels. The corresponding BER results of the QPSK-modulated VBLAST scheme and its PA-aided scheme are calculated as the benchmarkers.

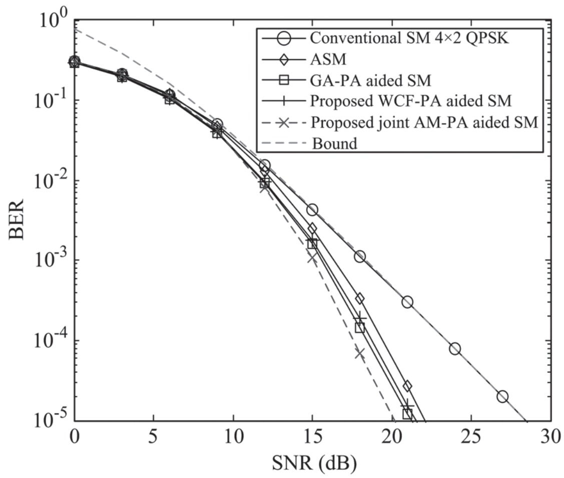

Fig. 5. BER comparison at $m_r = 4$ bits/symbol for the conventional QPSK-modulated SM, the ASM scheme, the proposed GA-PA-aided SM, the proposed WCF-PA SM, and the joint AM-PA-aided SM. Here, 4×2 MIMO channels are considered.

As indicated in Section II-C, our PA algorithm is to a degree reminiscent of the spatiotemporal PA scheme of [24], which is capable of achieving exponential diversity. However, they are different in the sense that the PA scheme in [24] relies on the long-term (time) average over which it maintains constant total power, whereas our scheme relies on the selection probability of the TAs for satisfying the total power constraint. Owing to this difference, the analysis method of [24] cannot be extended to our PA-aided SM scheme. However, since the power is allocated to both the spatial and temporal dimensions in our PA algorithm, more substantial performance gains may be expected than that of the pure spatial-domain PA-aided VBLAST scheme, as shown in Fig. 4. Deriving the explicit diversity order of this max-FD-based PA algorithm is based on the distribution of the FD distance. Since this FD depends both on the constellation and on the channel realization, its distribution is difficult to determine. Hence, the explicit diversity order of this max-FD-based PA algorithm is hard to characterize analytically. This challenge is also an open problem in the max-FD-aided PA algorithm of VBLAST [21], [22]. Nonetheless, the aforementioned challenge will be considered in our further research.

Fig. 5 compares the BER performances of the PA-aided schemes and of the conventional ASM schemes for $N_t = 4$. Upon comparing

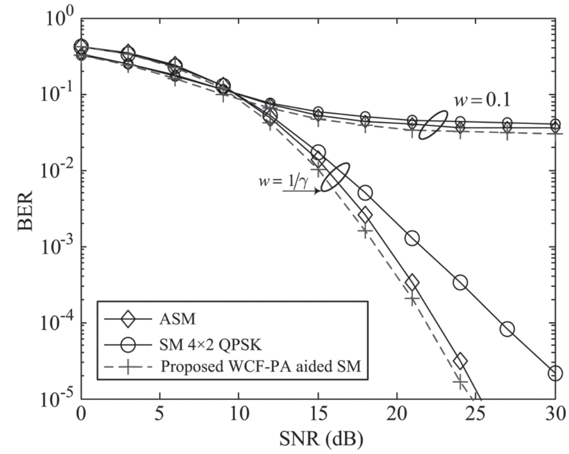

Fig. 6. BER comparison of the QPSK-aided SM, the ASM, and the proposed WCF-PA-aided SM schemes under 4×2 MIMO channels. We also considered the effects of CSI error associated with an equivalent channel estimation's noise variance of $w = 0.1$ and $(1/\gamma)$.

the results in Fig. 5 with the results in Fig. 4, we observe that the AM-aided and PA-assisted SM schemes exhibit different BER advantages for different numbers of TAs. This is because these techniques exploit different properties of the MIMO channels when aiming to maximize the FD, as indicated in Section II. As expected in Figs. 4 and 5, we observe that joint AM-PA-aided SM achieves the best BER performance among all the schemes.

Fig. 6 compares the achievable BER performance of the WCF-PA-aided SM in the presence of Gaussian-distributed channel state information (CSI) errors obeying $\mathcal{CN}(0, w)$ [7], [8] associated with $w = 0.1$ and $(1/\gamma)$, where γ is the average CSI estimation SNR at each receiver antenna. Observe in Fig. 6 that the BER performance of WCF-PA-aided SM is degraded upon introducing CSI estimation errors. However, this PA-based scheme still provides a considerable performance improvement over its nonadaptive counterparts with $w = 1/\gamma$.

V. Conclusion

In this paper, we have proposed the PA algorithms designed for limited-feedback SM-MIMO systems. Our simulation results confirm that the achievable performance is quite attractive. Our further work will be focused on the integration of space–time coding, channel coding, and space–time-shift keying [5] into the proposed schemes.

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AQ1 = Please check if changes made in this sentence are appropriate.
AQ2 = Please check if "SIC" was properly defined.

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