Abstract—This paper proposes a new precoding algorithm for orthogonalized spatial multiplexing (OSM) systems over flat-fading multiple-input multiple-output (MIMO) channels. The OSM scheme was recently introduced for closed-loop MIMO systems which allows single symbol decodability for maximum likelihood detection. To further improve the performance in OSM systems, we propose a new precoding method. For identifying the parameters of a precoder, we introduce a partitioning approach on the minimum Euclidean distance between constellation points in the effective channel. Also, it is shown that two real value parameters and one bit are required for feedback information in 4-QAM systems. Simulation results demonstrate that our precoding algorithm allows us to significantly improve the system performance with small increase of feedback values. We also confirm through simulations that the performance of the proposed scheme is the same as the optimum closed-loop MIMO systems.

I. INTRODUCTION

Communication over multiple-input multiple-output (MIMO) channels has been the subject of intense research over the past several years. The MIMO channels can offer much higher spatial multiplexing (SM) gains over their single-input single-output (SISO) counterpart [1][2][3]. Recently, orthogonalized spatial multiplexing (OSM) has been proposed which achieves orthogonality between transmitted symbols by applying phase rotation at the transmitter [4]. The OSM scheme is a new closed-loop MIMO system, which does not rely on the SVD. By applying rotation operations to the transmitted symbols, the OSM system allows a simple transmission method, the OSM scheme exhibits a good system performance overhead. Thus, compared to conventional SVD-based transmission methods, the OSM scheme exhibits a good system performance gain with lower complexity and feedback overhead.

In this paper, we introduce a new precoding scheme for OSM systems which further improve the performance of the OSM. First, we consider a criterion based on the Euclidean distance between constellation points in the effective channel, since the minimum Euclidean distance accounts for the symbol error probability. To efficiently identify the parameters of the proposed precoder, we introduce a partitioning approach. Our derivation indicates that two real value parameters and one bit are required for feedback information in 4-QAM systems. The simulation results show that our optimal precoding obtains 9dB and 7.5dB gains over the conventional OSM case at a bit error rate (BER) of $10^{-4}$ for 4-QAM and 16-QAM, respectively. We verify that the performance of the proposed system based on symbol by symbol detection is identical to that of the system in [5] based on joint ML detection.

II. SYSTEM DESCRIPTIONS

In this section, we consider a SM system with $M_t$ transmit and $M_r$ receive antennas in a flat fading channel. Throughout this paper, normal letters represent scalar quantities, boldface letters indicate vectors and boldface uppercase letters designate matrices. With a bar accounting for complex variables, for any complex notation $\tau$, we denote the real and imaginary parts of $\tau$ by $\Re[\tau]$ and $\Im[\tau]$, respectively.

We consider the complex channel output as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (1)$$

where $\mathbf{x} \in \mathbb{C}^{M_t \times 1}$ is the complex transmitted signal, $\mathbf{y} \in \mathbb{C}^{M_r \times 1}$ indicates the complex received signal, $\mathbf{H} \in \mathbb{C}^{M_r \times M_t}$ represents the complex channel matrix with the $(i,j)$th element $h_{ij}$ denoting the fading coefficient between the $j$th transmit and the $i$th receive antenna, and $\mathbf{n} \sim \mathcal{N}(0, \sigma_n^2 \mathbf{I}_{M_r})$ is the circularly symmetric complex Gaussian noise. Here, $\mathbf{I}_{M_r}$ denotes an identity matrix of size $M_r$. We assume that the elements of the MIMO channel matrix $\mathbf{H}$ are obtained from an independent and identically distributed (i.i.d.) complex Gaussian distribution with zero mean and unit variance. Each channel realization is assumed to be known at the receiver.

In what follows, we give a brief review on the OSM scheme [4]. The transmitter structure diagram of the OSM is given in [4]. The OSM orthogonalizes a channel by applying the rotation operation at the transmitter and transmits two independent data streams. Then, a single symbol decodable receiver is employed at the receiver which greatly decreases the detection complexity. To simplify the presentation, we focus on $M_t = 2$. Note that the original OSM scheme in [4] can be applied to any system with $M_t \geq 2$.

To orthogonalize the channel, the OSM precodes two transmitted symbols as $\mathbf{F}(\mathbf{x}, \theta) = \text{diag}\{1, \exp(j\theta)\}\mathbf{s}(\mathbf{x})$ where $\theta$...
is the rotation phase angle applied to the second antenna and
$s(x) \triangleq [\Re[x_1] + j\Re[x_2] \ \Im[x_1] + j\Im[x_2]]^T = [s_1(x) \ s_2(x)]^T$.

Employing the above precoding, equation (1) can be rewritten as

$$
y = \overline{H}F(x, \theta) + n = \overline{H}os(x) + n \tag{2}
$$

where $\overline{H}_o$ accounts for the effective channel matrix for $s(x)$, represented by $\overline{H}_o = \overline{H} \times \text{diag}\{1, \exp(j\theta)\}$.

Equivalently, the real-valued representation of the system (2) is given as [4]

$$
y = \overline{H}_o s(x) + n = \begin{bmatrix} \Re[y] \\ \Im[y] \end{bmatrix} = \begin{bmatrix} \Re[\overline{H}_o] \\ \Im[\overline{H}_o] \end{bmatrix} \begin{bmatrix} \Re[s(x)] \\ \Im[s(x)] \end{bmatrix} + \begin{bmatrix} \Re[\overline{H}] \\ \Im[\overline{H}] \end{bmatrix} = \begin{bmatrix} h_1^o \ h_2^o \ h_3^o \ h_4^o \end{bmatrix} s(x) + n \tag{3}
$$

where the real column vector $h_i^o$ of length $2M_r$ denotes the $i$th column of the effective real-valued channel matrix $\overline{H}_o$, $s(x)$ represents $[\Re[x_1] \ \Im[x_1] \ \Re[x_2] \ \Im[x_2]]^T$, and $n$ indicates $[\Re[n] \ \Im[n]]^T$.

From the real-valued representation of the channel matrix in (3), it is easy to see that the column vectors $h_1^o$ and $h_2^o$ are orthogonal to $h_3^o$ and $h_4^o$, respectively, regardless of $\theta$. We also notice that we have $h_1^o \cdot h_1^o = -h_2^o \cdot h_2^o$ for all $\theta$, where $\cdot$ denotes the inner product between vectors $a$ and $b$. In this case, $\overline{H}_o$ becomes orthogonal if and only if $h_1^o \perp h_4^o$ and $h_2^o \perp h_3^o$.

Then, the rotation angle $\theta = \theta_o$ for the orthogonality between $h_1^o$ and $h_4^o$ (or $h_2^o$ and $h_3^o$) can be written as [4]

$$
\theta_o = \tan^{-1}\left(\frac{B}{A}\right) = \frac{\pi}{2}
$$

where $A = \sum_{m=1}^{M_r} ||\overline{h}_m|| \sin(\angle\overline{h}_m - \angle\overline{h}_{m1})$ and $B = \sum_{m=1}^{M_r} ||\overline{h}_m|| \cos(\angle\overline{h}_m - \angle\overline{h}_{m1})$. This rotation angle makes $h_1^o$ and $h_2^o$ (or $h_3^o$ and $h_4^o$) orthogonal to each other.

Utilizing this orthogonality, the ML estimate of transmitted symbol $\hat{x}_1$ and $\hat{x}_2$ can be obtained as [4]

$$
\hat{x}_1 = \arg\min_{\tau \in Q} \left|\y - [h_1^o \ h_2^o] \begin{bmatrix} \Re[\tau_1] \\ \Im[\tau_1] \end{bmatrix} \right|^2 \tag{4}
$$

and

$$
\hat{x}_2 = \arg\min_{\tau \in Q} \left|\y - [h_3^o \ h_4^o] \begin{bmatrix} \Re[\tau_2] \\ \Im[\tau_2] \end{bmatrix} \right|^2 \tag{5}
$$

where $Q$ is a signal constellation of size $M_c$. As a result, the complexity of the ML detection of the OSM reduces from $M_r^2$ to $M_c$.

III. PRECODING SCHEMES FOR OSM

First we discuss the necessity of the precoding for OSM systems. We can see from equations (4) and (5) that a solution for the OSM system is transformed into two SISO equations. Here, $h_1^o$, $h_2^o$, $h_3^o$ and $h_4^o$ represent the channel column vectors corresponding to the components of two transmitted symbols. It was shown in [4] that the subspace spanned by $h_1^o$ and $h_2^o$ is orthogonal to that spanned by $h_3^o$ and $h_4^o$ in the

![Schematic diagram of the transmitter structure for the proposed scheme](image)

OSM system. It is easy to see that we have $||h_1^o|| = ||h_2^o||$, $||h_3^o|| = ||h_4^o||$, and $h_1^o \cdot h_2^o = h_3^o \cdot h_4^o$. Because of this fact, one may think that a precoding operation is not necessary. However the column vectors $h_1^o$ and $h_3^o$ are not orthogonal to $h_2^o$ and $h_4^o$, respectively. Also, $||h_1^o||$ is not equal to $||h_2^o|| = ||h_3^o||$. As a result, the channel energies corresponding to the inphase and quadrature components are still different and the symbol with smaller channel gain can degrade the performance.

Motivated by this observation, we present a new precoding algorithm for OSM systems which maximizes the minimum distance denoted by $d_{min}$, since the performance of the optimum ML receiver depends on the minimum Euclidean distance in the received signal constellation [6]. In this section, we introduce the optimal precoding for the OSM system. To facilitate the illustration, we focus on the $M_r = 2$ case. Note that the proposed scheme can be extended to systems with $M_r > 2$ using the similar approach in [4].

To construct the precoding for OSM systems, we can rewrite the system model in (3) as

$$
y = \overline{H}_o P s(x) + n \tag{6}
$$

where $P$ denotes a 4-by-4 real precoding matrix. Then, in order to preserve the orthogonality between the transmitting symbols, we can choose $P$ as a block orthogonal matrix,

$$
P = \begin{bmatrix} P_1 & 0 \\ 0 & P_2 \end{bmatrix} \tag{7}
$$

where $P_1$ and $P_2$ are 2-by-2 real matrices. Then, $P_1$ and $P_2$ controls the channel gain of $\tau_1$ and $\tau_2$, respectively. Since the channel quality in the first and second columns of $\overline{H}_o$ is identical to that in the third and fourth columns of $\overline{H}_o$ in terms of $d_{min}$, $P_1$ and $P_2$ within the precoding matrix are the same ($P_1 = P_2$). Now, the precoder design problem reduces to identifying a 2-by-2 real matrix $P_1$ which improves the OSM performance. In this paper, we decompose $P_1$ into three 2-by-2 real matrices as

$$
P = \overline{H}_o R_{\theta_1} D' R_{\theta_2} = \begin{bmatrix} R_{\theta_1} & 0 \\ 0 & R_{\theta_2} \end{bmatrix} \begin{bmatrix} D & 0 \\ 0 & D \end{bmatrix} \begin{bmatrix} R_{\theta_2} & 0 \\ 0 & R_{\theta_1} \end{bmatrix}
$$
By substituting the precoding in (8) into the system model in (6), we have

\[ y = H_\theta R'_\theta D'R'_\theta s(x) + n = H_\theta^1 R'\theta_2 s(x) + n = H_\theta^2 s(x) + n \]

where \( H_R^\theta = H_\theta R'_\theta = [h_1^R h_2^R h_3 \ldots h_n^R] \) and \( H_p = H_\theta R'_\theta D'R'_\theta = [h_1^p h_2^p h_3 h_4^p] \). We assume that the total transmit power is constrained to be \( \mathbb{E}(\text{tr}(Ps(x)s(x)^T)) = P_T \) which equals \( \mathbb{E}(\text{tr}(s(x)s(x)^T)) = P_T \).

The structure of the proposed precoding system is shown in Fig. 1.

Now, by transforming Equation (4) into the equation with the precoding (7), the ML solution for \( \pi_1 \) in our scheme can be rewritten as

\[ \pi_1 = \arg \min_{\pi \in \mathbb{Q}} \left\| y - [h_1^p h_2^p] \begin{bmatrix} H_1 \pi_1 \\ H_2 \pi_1 \end{bmatrix} \right\|^2 \]

Similarly, \( \pi_2 \) can be obtained by using \( h_4^p \) and \( h_2^p \). Since the channel gains of \( \pi_1 \) and \( \pi_2 \) are the same in terms of \( d_{\min} \), we will consider one ML solution in (9) from now on.

To obtain the precoding matrix which maximizes \( d_{\min} \), we need to compute three parameters \( \theta_1 \), \( \theta_2 \) and \( p \), for (8). Because it is very complex to jointly optimize these three parameters, we simplify the computation by prefixing \( \theta_1 \) with the inner rotation introduced in [7]. Even though this approach may be non-optimal for OSM systems, we can confirm through the simulation results that the proposed precoding achieves the performance identical to that of the optimum closed-loop general MIMO systems. The orthogonality between \( h_{1R} \) and \( h_{2R} \) is achieved with the following angle [7]

\[ \theta_1 = \tan^{-1} \left( \frac{C + \sqrt{C^2 + 4D^2}}{2D} \right) \]

where \( C = ||h_{2R}^1||^2 - ||h_1^1||^2 \) and \( D = (h_1^1)^T h_2^1 \). Here, \( ||\theta_1^R|| \) is maximized for \( \theta_1 = \tan^{-1} \left( \frac{C + \sqrt{C^2 + 4D^2}}{2D} \right) \), whereas \( ||\theta_2^R|| \) is maximized for \( \theta_1 = \tan^{-1} \left( \frac{C - \sqrt{C^2 + 4D^2}}{2D} \right) \). Because both two solutions make the same \( d_{\min} \), we consider only the first case \( \theta_1 = \tan^{-1} \left( \frac{C + \sqrt{C^2 + 4D^2}}{2D} \right) \). Then, the magnitudes of \( h_{1R}^1 \) and \( h_{2R}^1 \) equal the first and the second singular value of the matrix \( \mathbf{H} \), respectively.

Now, the other two parameters \( \theta_2 \) and \( p \) need to be determined. In what follows, we will show that our scheme requires only a finite set of \( p \) and \( \theta_2 \) for feedback information. Thus just a few bits are sufficient for an index of the chosen set of \( p \) and \( \theta_2 \).
area $0 \leq p \leq \sqrt{2}$ and $0 \leq \theta_2 \leq \pi/4$ is divided into three partitions as shown in Fig. 2. Three partitions are made in Fig. 2 such that the number of candidates of $d_{min}$ is minimized. Defining $k$ as $\|h_{1R}^\theta\|^2/\|h_{2R}^\theta\|^2$, we will find $p$ and $\theta_2$ which maximize $d_{min}$ in each partition. Since we choose $\theta_1$ so as to maximize $\|h_{1R}^\theta\|$, we have $\|h_{1R}^\theta\|^2 \geq \|h_{2R}^\theta\|^2$. Thus, this leads to $k \geq 1$. Then, after finding the maximum value of $d_{min}$ in each partition, we can identify the maximum $d_{min}$ for the whole region. We start with the partition $A_2^\theta$.

1) Partition $A_2^\theta \left( \sqrt{2}/(k+1) \leq p \leq \sqrt{6}/(k+3), 0 \leq \theta_2 \leq \pi/4 \right)$

In this partition, it is easy to see that $\|h_{1R}^\theta\|^2/p^2 \geq \|h_{2R}^\theta\|^2/(2-p^2)$. Using this and equation (12), we have the following relations: $\|h_{1p}^\theta\|^2 \geq \|h_{2p}^\theta\|^2$, $\|h_{1p}^\theta - h_{2p}^\theta\|^2 \geq \|h_{1}^\theta + h_{2p}^\theta\|^2$ and $\|h_{1p}^\theta + h_{2p}^\theta\|^2 \geq \|h_{2p}^\theta\|^2$. This means that $d_{min}$ is equal to $\|h_{2p}^\theta\|$ in this partition, regardless of $p$ and $\theta_2$. For given $\theta_2$, $\|h_{2p}^\theta\|^2$ is a monotonically increasing function of $p$ for $\sqrt{2}/(k+1) \leq p \leq \sqrt{6}/(k+3)$. Also, for given $p$, $\|h_{2p}^\theta\|^2$ is a monotonically increasing function of $\theta_2$ for $0 \leq \theta_2 \leq \pi/4$. As a result, to maximize $d_{min}$, $p$ and $\theta_2$ must be $\sqrt{6}/(k+3)$ and $\pi/4$, respectively. Then, for these values, we have $d_{min} = 4k\|h_{2R}^\theta\|^2/(k+3)$.

2) Partition $A_3^\theta \left( \sqrt{6}/(k+3) \leq p \leq \sqrt{2}, 0 \leq \theta_2 \leq \pi/4 \right)$

In this partition, we have $\|h_{1R}^\theta\|^2 \geq \|h_{2p}^\theta\|^2$ and $\|h_{1R}^\theta - h_{2p}^\theta\|^2 \geq \|h_{1p}^\theta + h_{2p}^\theta\|^2$. Then, the candidates of $d_{min}$ are $\|h_{2p}^\theta\|$ and $\|h_{1p}^\theta + h_{2p}^\theta\|$. We can show that the case of $\|h_{2p}^\theta\|^2 = \|h_{1p}^\theta + h_{2p}^\theta\|^2$ maximizes $d_{min}$ in this partition, and the proof is presented in [8]. By solving this equality, $\theta_2$ can be expressed as follows

$$\theta_2 = \sin^{-1} \sqrt{3k-1} p^2 - (2 - p^2) \sqrt{(3k-1) p^2 - (2 - p^2)^2 - 5k^2} \sqrt{5(kp^2 - (2 - p^2)^2).} \quad (13)$$

Substituting Eq. (13) into Eq. (12) yields

$$\|h_{2R}^\theta\|^2 = \frac{3(k-1)p^2 + 6}{5} \sqrt{(3k-1) p^2 - (2 - p^2)^2 - 5k^2} \sqrt{5(kp^2 - (2 - p^2))}. \quad (14)$$

To identify the maximum value of Equation (14), we need to find the derivative of (14) with respect to $p$. The derivative of $\|h_{2p}^\theta\|^2$ is given by

$$\frac{d}{dp} \|h_{2p}^\theta\|^2 = \frac{2p}{5} \left( 3(k-1) - \frac{(4k^2 + 12k + 4) p^2 - (12k + 8) \sqrt{(4k^2 + 12k + 4) p^4 - 2(12k + 8) p^2 + 16}}{(16)} \right). \quad (15)$$

Then, rearranging Equation (15) becomes

$$\frac{(k^2 - 6k + 1)(k^2 + 3k + 1)^2}{4} - 2(3k + 2)(k^2 - 6k + 1) p^2 + 4(-6k + 1) = 0. \quad (16)$$

Equation (16) is a quadratic equation of $p^2$ and the root of the equation can be obtained as

$$p^2 = \frac{(3k+2)(k^2 - 6k + 1) \pm \sqrt{9k^2(k^2 - 6k + 1)(k-1)^2}}{(k^2 - 6k + 1)(k^2 + 3k + 1)} \quad (17).$$

Using Equations (16) and (17), we can recognize that $p$ must be on the boundary of this partition to maximize $d_{min}$, which is equal to $\|h_{2p}^\theta\|^2$. Thus, $p$ must be either $\sqrt{6}/(k+3)$ or $\sqrt{2}$ to maximize Equation (14). Correspondingly, $\theta_2$ is either $\pi/4$ or $0.464$ and $d_{min}$ is either $4k\|h_{2R}^\theta\|^2/(k+3)$ or $2k\|h_{2R}^\theta\|^2/5$, respectively. It is interesting to note that $\theta_2$ is independent of $k$.

3) Partition $A_4^\theta \left( 0 \leq p \leq \sqrt{2}/(k+1), 0 \leq \theta_2 \leq \pi/4 \right)$

In this partition, since the magnitude of the first column in $H_R^\theta$ is made smaller than that of the second column for the range of $p$ in $A_4^\theta$, we have a smaller maximum value of $d_{min}$ than the other partitions. Thus, we do not consider this partition.

Now that all three partitions are computed, we are ready to obtain $p_{opt}$ and $\theta_{2opt}$ for the proposed precoding.

**TABLE I**

<table>
<thead>
<tr>
<th>Mod</th>
<th>Case</th>
<th>$p_{opt}$</th>
<th>$\theta_{2opt}$</th>
</tr>
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<tbody>
<tr>
<td>4-QAM</td>
<td></td>
<td></td>
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<tr>
<td></td>
<td>$1 \leq k \leq 2$</td>
<td>$\pi/4$</td>
<td></td>
</tr>
<tr>
<td>16-QAM</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$1 \leq k \leq 7, 59$</td>
<td>$\pi/4$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$8 \leq k \leq 33, 1$</td>
<td>$0.464$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$33.11 \leq k \leq 101.03$</td>
<td>$0.345$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$k \geq 101.03$</td>
<td>$0.245$</td>
<td></td>
</tr>
</tbody>
</table>

This result indicates that there are only two candidate sets of $p$ and $\theta_2$ for 4-QAM. Thus, after the value $k$ is determined at the receiver, one of two sets in (18) is employed depending on the threshold value of $k = 7$.

Similar to 4-QAM, the optimal parameters for 16-QAM can be obtained as

$$p_{opt} = \frac{\sqrt{6}/(k+3)}{k}, \quad \theta_{2opt} = \pi/4 \quad \text{if } 1 \leq k < 7.59$$

$$p_{opt} = \frac{\sqrt{6}/(k+3)}{k+21}, \quad \theta_{2opt} = 0.489 \quad \text{if } 7.59 \leq k < 43.11$$

$$p_{opt} = \frac{\sqrt{182}/(k+91)}{k+91}, \quad \theta_{2opt} = 0.345 \quad \text{if } 43.11 \leq k < 101.03$$

$$p_{opt} = \frac{\sqrt{2}}{k+91}, \quad \theta_{2opt} = 0.245 \quad \text{if } k > 101.03$$

The detailed derivation of this result is presented in [8]. It is clear that the partitioning approach presented in this section is effective for identifying the maximum of $d_{min}$. We note that the boundary of the partition is critical, and that $\|h_{2p}^\theta\| = \|h_{1p}^\theta + nh_{2p}^\theta\|$ for $n = 1, 2, \ldots, \log_2 M-1$ in M-QAM systems are an important equation to determine the maximum of $d_{min}$. From this fact, the same approach made
in this section can be extended to higher modulation systems such as 64-QAM.

In the proposed precoding systems, two real values $\theta_1$ and $k$ are needed for feedback additionally in comparison to the original OSM. For solutions (18) and (19), $k$ should be reported to the transmitter. In what follows, simplified methods are presented to further reduce the feedback information without any performance loss. We first recognize that the variation in $p_{opt}$ is small in Equations (18) and (19) in terms of $k$. Thus, we can replace the variable $p$ with 1 or $\sqrt{2}$. The final result is depicted in Table I. Note that the case $p = 1$ indicates no power loading in (8) ($D=I$). With this simplification, instead of feeding back the value $k$ to the transmitter, only one bit and two bits are required for feedback information in 4-QAM and 16-QAM systems, respectively. In the simulation section, it will be shown that this simplification causes no performance loss.

IV. SIMULATION RESULTS

In this section, we provide simulation results to demonstrate the effectiveness of the proposed precoding algorithm for OSM systems, and compare with the original OSM systems.

In Figure 3, we compare the bit-error-rate (BER) performance of various systems for 4-QAM and 16-QAM constellations. First, we note that the performance of the proposed precoding is the same as that of the optimal precoding presented in [5]. It was shown in [5] that the precoding system in [5] is optimal in terms of $d_{min}$ in closed-loop MIMO systems. Comparing this optimum precoding with the proposed algorithm, we confirm that the proposed precoding attains the optimal performance. Note that the optimum precoding in [5] was presented in only BPSK and 4-QAM systems. Also, the receiver structure of the precoding in [5] requires joint ML detection which makes it difficult to be applied for higher modulation systems due to high computational complexity. On the contrary, our precoding method is based on the symbol-by-symbol detection for arbitrary modulation levels. The proposed algorithm requires two real feedback information and one bit for 4-QAM, which is also less than the system in [5]. It is interesting to note that even though the original OSM does not rely on the SVD operation, the optimized OSM with our proposed precoder exhibits the performance identical to the optimum precoding based on SVD in [5].

Also, we can see that the proposed precoding algorithm provides 9dB and 7.5dB gains at a BER of $10^{-4}$ over the original OSM with 4-QAM and 16-QAM in Fig. 3, respectively. Compared to the original OSM, our precoding method requires one additional feedback value for $\theta_1$ and one bit. Also, we depict the power loading for OSM which we previously proposed in [9]. Simulation results show that the performance improvement of our proposed scheme is 4dB and 2.5dB in comparison with power loading precoding for OSM in 4-QAM and 16-QAM systems, respectively.

V. CONCLUSION

In this paper, we have presented a new precoding algorithm for the OSM scheme in MIMO systems, which maximizes the minimum Euclidean distance to enhance the system performance. To determine the optimal precoding parameters, we have illustrated a partitioning approach on the minimum Euclidean distance between the constellation points in the effective channel in OSM systems. The derivation shows that our scheme requires two real values and one bit for feedback for 4-QAM. The simulation results confirm that the proposed algorithm for the OSM is optimal in terms of $d_{min}$ in closed-loop MIMO systems. It is straightforward to extend the proposed algorithm to higher level modulations. Also, the proposed scheme can be applied to systems which support more than two data streams by employing a method presented in [10].

REFERENCES