ADAPTIVE HYBRID SPATIAL MODULATION SCHEME FOR MIMO SYSTEMS

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Spatial modulation (SM) is an emerging communication technology applied in MIMO systems, which can further exploit the potential of MIMO systems. A novel hybrid spatial modulation (HSM) policy is developed in this paper to better match the changing channel conditions in MIMO systems. In HSM, the transmission policy switches between different SM schemes for each channel realization according to the value of the minimum Euclidean distance. In addition, by employing antenna selection mechanisms (specifically, EDAS and COAS) in HSM, improved versions named HSM-EDAS, and HSM-COAS are proposed. In order to reduce the high computation complexity occurred in HSM-EDAS, two low-complexity algorithms are further presented. Simulation results verify the theoretical analysis that the presented HSM, HSM-EDAS and HSM-COAS schemes all decrease the pairwise error probability (PEP) and HSM-EDAS outperforms HSM-COAS whereas the latter achieves better tradeoff between performance and capacity with much lower computation complexity.

Keywords: spatial modulation, MIMO systems, antenna selection

1. Introduction

Spatial modulation (SM) has emerged in recent years as an attractive technology to convey information by exploiting jointly the traditional digital modulation symbols and antenna indices for wireless multiple input multiple output (MIMO) systems. The application of SM can significantly simplify the hardware implementation structure with lower signal processing complexity and higher flexibility for link configuration. Therefore, SM is expected to provide efficient and reliable solutions for future mobile MIMO communication systems.

Under SM scheme, only one transmit antenna is allowed to transmit data per time-slot, so that the problems of inter-channel interference (ICI) and inter-antenna synchronization (IAS) existed in traditional MIMO systems are avoided effectively. However, the main drawback of SM is that the utilization rate of antennas is low. The mechanism of allowing only one transmit antenna to send signals in SM leads to lower system capacity and worse bit error rate (BER)

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performance compared to other MIMO systems where all antennas are allowed to send [1-3]. Although the transmit antenna serial number can carry some data, the transmission efficiency of SM system is still poor when there have not enough antennas. In addition, the modulation mode of most SM systems is binary and the number of transmit antennas must be in power of two, which means plenty of antennas are needed when high spectral efficiency is required [4]. In order to further improve antenna utilization rate and transmission efficiency, a Quadrature Spatial Modulation (QSM) system model was proposed in literature [5] and [6]. The essential difference between QSM and SM is that SM system uses one antenna to send data whereas QSM divides data symbols into real and virtual parts and utilizes two antennas to transmit these two components respectively. On one hand, due to the orthogonality of the two components, it can be guaranteed that the two signals will not interfere with each other when transmitted simultaneously. On the other hand, the transmission efficiency under QSM is effectively improved compared with the traditional SM since two antennas are employed to transmit signals. In order to be able to use more transmit antennas in SM system, an idea of Generalized-SM (GSM) was proposed [7-9] in which part of antennas are selected to send signals. The growth in the number of available antennas not only improves the information transmission rate in time domain but also increases the number of bits that carry information. For the purpose of maximizing the utilization of antenna resources, extended spatial modulation (ESM) mechanism was put forward in [10-11] who allows all transmit antennas to send information in one time-slot including those inactive ones. The spectral efficiency under ESM has a linear growth relationship with the number of transmit antennas rather than a Logarithmic growth, which enables ESM to achieve better spectral efficiency with a certain number of antennas.

The above-mentioned strategies based on SM system are all belong to structural improvement to achieve higher transmission efficiency and better antenna utilization. Each structure has its own advantages and should be selected properly according to the actual requirements in practical application. It should be noted that the approaches above also brings new problems. In MIMO systems, increasing the number of transmit antennas results in high hardware costs and computational burden. In MIMO systems, the application of transmit antenna selection has been proved to be able to reduce the hardware implementation complexity, attain a gain on signal-to-noise (SNR) ratio and effectively increase the system capacity [12]. Moreover, antenna selection also solves the problem of antenna resource shortage effectively by allocating antennas reasonably [13].

Following this motivation, a Euclidean distance optimized antenna selection named EDAS technique was presented in [14]. In EDAS, a subset of transmit antennas is selected to maximize the minimum instantaneous Euclidean distance who determines the performance of system using maximum likelihood

(ML) detection. EDAS can obtain optimal BER by traversing all the combinations of antenna serial number and modulation symbol, which inevitably brings high complexity especially for SM system with a mass of antennas [15]. In order to reduce the computation complexity, an antenna selection mechanism denoted by capacity optimized antenna selection (COAS) was put forward based on singular value decomposition [14]. In [16], the complexity involved in EDAS was reduced by searching only the largest least-square singular value. The authors in [17] show that EDAS-based algorithms perform better than COAS whereas COAS achieves better tradeoff between system capacity and BER with much lower computation complexity. In [18], QR decomposition is employed and performed on channels that can be decomposed into orthogonal amplitude modulation symbol sets, which reduces the number of calculating Hamming distances and thus reduces the complexity of the algorithm.

In recent years, a series of new solutions for spatial modulation and antenna selection have emerged, aiming at obtaining better performance in power control, information security and other aspects [19-23]. Provided that the channel states are available, the performance of SM systems can be enhanced by utilizing adaptive power allocation. The authors of [19] considered the power allocation optimization in SM systems from the perspective of information theory, and proposed two power allocation optimization algorithms to obtain the mutual information, which effectively improves SM performance and increases the capacity. In [20], Artificial noise (AN) was introduced into the GSM systems and a power minimization scheme of AN-aided GSM systems was presented to improve the power efficiency without affecting the interference effect produced by AN. The authors of [21] studied the security of AN-aided GSM systems. For the purpose of improving the jamming intensity of the traditional AN scheme, a Euclidean distance optimization Scheme (ED-AN) was proposed to minimize the Euclidean distance between transmitted and jamming signal, which brings better security. When eavesdroppers exist, security cannot be guaranteed if only optimal capacity of legitimate recipients is considered. In this case, employing physical-layer security technology (such as AN and antenna selection) can improve the security of the system. In order to maximize the noise to the eavesdropper, the AN vector was rearranged in [22] by utilizing beam forming technique so that it is collinear with the eavesdropper channel. Such physical-layer design can significantly reduce the BER performance of eavesdropper without affecting the legitimate receivers. Theoretical and simulation results show that the joint antenna selection and AN algorithm can also effectively enhance the physical-layer security of SM systems [23].

Different from the above algorithms developed in [19-23], this paper still focuses on the BER performance and implementation complexity of SM systems. Research showed that an adaptive mode which switches among different SM-based transmission modes can better adapt to the changing channel conditions [24]. Toward this direction, we propose a hybrid spatial modulation (HSM) scheme to enhance the BER performance of SM systems in this paper. Under HSM, a transmission mode switches between SM and ESM according to the value of the minimum Euclidean distance. Then, antenna selection mechanism EDAS is introduced, and an EDAS-based hybrid spatial modulation scheme named as HSM-EDAS is put forward, which increases the probability that the minimum Euclidean distance of ESM is greater than that of SM and this probability determines the BER performance. In order to reduce the complexity of HSM-EDAS, two reduced-complexity HSM-EDAS algorithms, referred respectively as RC-HSM-EDAS and FRC-HSM-EDAS are developed. Moreover, combined with COAS policy, a novel HSM-COAS scheme is proposed with much lower complexity than that of EDAS-based schemes. In terms of BER performance, it can be demonstrated that EDAS-based HSM schemes achieve lower BER whereas HSM-COAS can balance the capacity and performance better.

The rest of this paper can be divided into the following sections. System model is presented in section 2. Hybrid Spatial Modulation (HSM) Scheme is proposed in section 3. In section 4, the improved versions of HSM by employing antenna selection are proposed, i.e., HSM-EDAS and HSM-COAS. In addition, two reduced-complexity HSM-EDAS schemes are put forward to reduce the high complexity occurred in HSM-EDAS. Section 5 gives the analysis and comparison of computation complexity of the presented policies. In section 6 and section 7, simulation results and conclusion are presented, respectively.

Notations: we use uppercase letters and bold lowercase to denote matrices and column vectors respectively. $[\cdot]^T$, $(\cdot)^H$ and $\|\cdot\|$ represent respectively the transposition, Hermitian operation and Frobenius norm. We use $\binom{\cdot}{\cdot}$ to denote the binomial coefficient and $|\mathrm{A}|$ is cardinality of set A. The minimum square singular value of a matrix M is defined as σ(M).

2. System Model

In this paper, there exist N_t transmit antennas and N_r receive antennas in our SM and ESM systems. In SM scheme, the input bits sequence **b** is mapped into two carriers, the activated antenna and the modulated symbol. SM allows one antenna at the transmitter to be activated at one time slot so that only one element in the transmit vector **x** is nonzero [1]. Unlike SM, ESM scheme can activate multiple transmit antennas according to the given incoming bits stream [10-11]. Then the transmit vector of ESM can be represented as $\mathbf{x} = [0,...,\mathrm{s}_m, 0,...,\mathrm{s}_m, 0,...]^T$, where $s_m (1 \le m \le M)$ denotes one of the symbols of the conventional modulation scheme employed, such as M-QAM.

Hence, for both the receivers of SM and ESM scheme, the N_r-dimension received signal can be given by

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

where $\mathbf{x} \in \mathbb{C}^{N_t \times 1}$ is the transmit vector of SM or ESM scheme and the channel matrix is $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_t}] \in \mathbb{C}^{N_r \times N_t}$, whose elements $h_{i,j}$ ($1 \le i \le N_r$, $1 \le j \le N_t$) are random variables independent of each other following the Gaussian distribution with a mean of 0 and a variance of 1, i.e., obeying $\mathcal{CN}(0,1)$, and $\mathbf{n} \in \mathbb{C}^{N_r}$ represents the noise whose components are Gaussian random variables $\mathcal{CN}(0, N_0)$.

To evaluate the transmit vector, maximum likelihood (ML)-based optimal detector [14] is applied, which is given by

$$\begin{aligned} \hat{\mathbf{x}} &= \arg\max_{\mathbf{x}_i \in \Lambda} p_Y(\mathbf{y} | \mathbf{x}_i, \mathbf{H}) \\ &= \arg\min_{\mathbf{x}_i \in \Lambda} \| \mathbf{y} - \mathbf{H}\mathbf{x}_i \|_F^2 \end{aligned} \tag{2}$$

where $\hat{\mathbf{x}}$ is the transmit vector estimated by the detector and Λ denotes the set of all possible transmit vectors, $\mathbf{x}_i$ ($1 \le i \le 2^{m_b}$, m_b is the spectral efficiency) is one of the possible transmit vector. $p_Y(\mathbf{y} | \mathbf{x}_i, \mathbf{H})$ in (2) represents the conditioned probability density of the received signal.

In case of high SNR, the approximation of the pairwise error probability (PEP) of SM or ESM using ML detection can be obtained as [25]

$$P\left(\mathbf{x}_i \to \mathbf{x}_j | \mathbf{H}\right) \approx \lambda \cdot Q\left(\sqrt{\frac{1}{2N_0} d_{min}^2(\mathbf{H})}\right) \tag{3}$$

where $P\left(\mathbf{x}_i \to \mathbf{x}_j | \mathbf{H}\right)$ denotes the conditioned PEP that codeword $\mathbf{x}_i$ is erroneously detected as $\mathbf{x}_j$, $Q(x)$ is a variable limit integral function given by

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{(-y^2/2)} dy$$

and λ denotes the number of spatial constellation neighbor points. In equation (3), minimum square Euclidean distance $d_{min}^2(\mathbf{H})$ is given as follows

$$d_{min}^2(\mathbf{H}) = \underset{\substack{\mathbf{x}_i, \mathbf{x}_j \in \Lambda \\ \mathbf{x}_i \neq \mathbf{x}_j}}{\arg\min} \| \mathbf{H}(\mathbf{x}_i - \mathbf{x}_j) \|_F^2 \quad (4)$$

where $\mathbf{x}_i$ and $\mathbf{x}_j$ are two possible transmit vectors.

3. Hybrid Spatial Modulation (HSM) Scheme

It has been mentioned above that SM avoids ICI and IAS perfectly by using only one antenna to convey information at one time-slot. However, it can be observed from [10] that ESM can almost save half of the transmit antennas than SM when the spectral efficiency and modulation order are both fixed. To combine the benefits of SM and ESM strategies, a Hybrid spatial modulation (HSM) scheme is proposed and it can be summarized as follows: we assume that the channel state is varying slowly, the receiver estimates channel state information for each channel realization and selects the optimal transmission scheme between SM and ESM, and then sends the information to the transmitter. The system model is shown in Fig. 1.

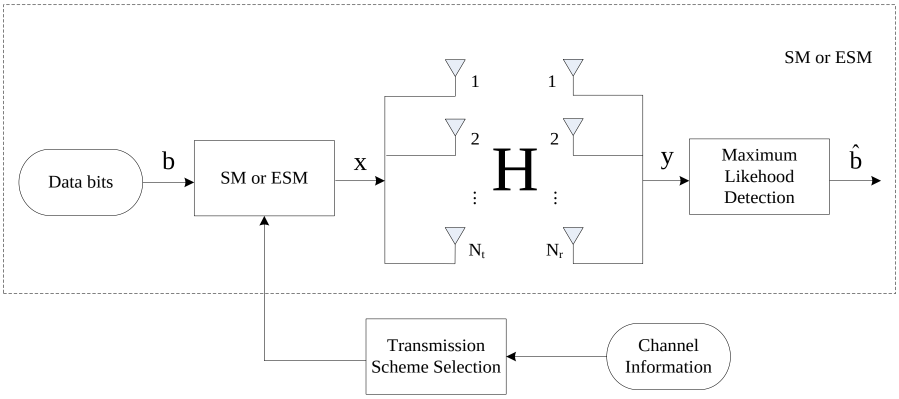

Fig. 1. The Hybrid spatial modulation model.

It is observed from (3) that the system performance mainly depends on the minimum square Euclidean distance and the conditioned PEP is a monotonically decreasing function about it, which means the system performance can be enhanced by maximizing the minimum square Euclidean distance. From the definition of minimum Euclidean distance given by (4), one can find that the value of $d_{min}^2(\mathbf{H})$ is mainly determined by the channel condition realization matrix $\mathbf{H}$ and the transmit vector $\mathbf{x}$. Hence, for a fixed channel realization,

$d_{min}^2(\mathbf{H})$ would change under different transmission schemes adaptively. Therefore, it is possible to exploit the channel freedom to improve the system performance by switching between ESM and SM schemes, which is the main motivation for proposing the HSM policy. The switching criterion can be formulated as follows:

Switching Criterion: For each channel realization, compute the minimum square Euclidean distance of SM denoted by $d_{min,SM}^2(\mathbf{H})$ and that of ESM denoted by $d_{min,ESM}^2(\mathbf{H})$ according to (4) and then choose the scheme with larger minimum Euclidean distance. This is briefly shown as

$$HSM = \begin{cases} SM, & d_{min,SM}^2(\mathbf{H}) > d_{min,ESM}^2(\mathbf{H}) \\ ESM, & d_{min,SM}^2(\mathbf{H}) \le d_{min,ESM}^2(\mathbf{H}) \end{cases}. \quad (5)$$

4. Hybrid Spatial Modulation Schemes based on Antenna Selection

As mentioned above, the antenna selection technique can be employed in spatial modulation systems to increase system capacity and obtain SNR gain. Therefore, we combine the HSM strategy with two antenna selection mechanisms, i.e., EDAS and COAS mentioned above and propose hybrid spatial modulation schemes based on antenna selection in this section. Hybrid spatial modulation scheme combined with antenna selection can be structured as Fig. 2.

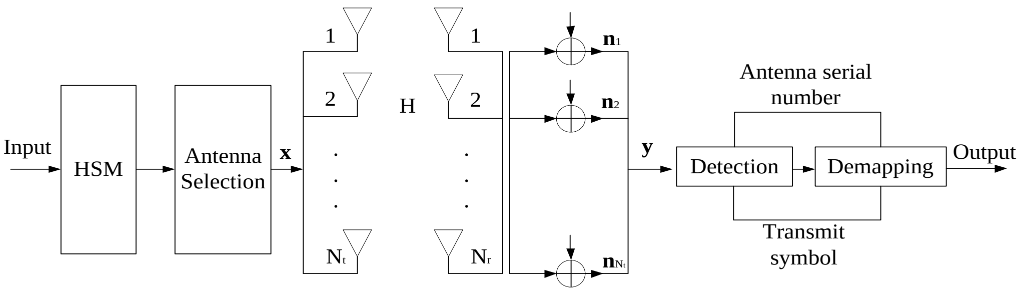

Fig. 2. Structure of HSM combined with antenna selection.

4.1 HSM-EDAS

It can be obtained from (5) that the system performance of HSM is determined by the values of $d_{min,SM}^2(\mathbf{H})$ and $d_{min,ESM}^2(\mathbf{H})$. Hence, an EDAS algorithm is proposed for ESM, referred as ESM-EDAS to increase $d_{min,ESM}^2(\mathbf{H})$. Let Γ represent the set of all possible combinations of N_t antennas chosen from

N_s transmit antennas. As a result, the specific antenna set who maximizes the minimum square Euclidean distance of all possible candidates is selected as [14]

$$l_{selected} = \arg\max_{l\in\Gamma}\{ \min_{\mathbf{x}_i \neq \mathbf{x}_j \in \Lambda} \| \mathbf{H}^{(l)}(\mathbf{x}_i - \mathbf{x}_j) \|_F^2 \} \tag{6}$$

where $\mathbf{H}^{(l)} = [\mathbf{h}_1^{(l)}, \mathbf{h}_2^{(l)}, \ldots, \mathbf{h}_{N_t}^{(l)}] \in \mathbb{C}^{N_r \times N_t}$ which has N_t columns given by l and Λ represents the set of all possible transmit vectors, $\mathbf{x}_i$ and $\mathbf{x}_j$ are two possible ESM transmit vectors. As is mentioned above that for ESM system, $\mathbf{x}_i$ and $\mathbf{x}_j$ can be expressed as $\mathbf{x}_i = [0,...,\mathrm{s}_1,0,...,\mathrm{s}_1,0,...]^T$ and $\mathbf{x}_j = [0,...,\mathrm{s}_2,0,...,\mathrm{s}_2,0,...]^T$ where s_1 and s_2 are random symbols selected from a M-QAM signal set S. Let us define $\Phi_i^{(l)}, \Phi_j^{(l)} \subset \Phi^{(l)}$ as the set of the indices of the element in $\mathbf{x}_i$ and $\mathbf{x}_j$ who is nonzero for a given l. For example, if $\mathbf{x}_i = [+1\ 0], \mathbf{x}_j = [+1\ +1]$, then we have $\Phi_i^{(l)} = \{1\}, \Phi_j^{(l)} = \{1,2\}$. Hence, (6) is rewritten as

$$\begin{aligned} l_{selected} &= \arg\max_{l\in\Gamma}\{ \min_{\substack{\Phi_i^{(l)},\Phi_j^{(l)} \in \Phi^{(l)} \\ s_1,s_2 \in S}} \| \sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} s_1 - \sum_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)} s_2 \|_F^2 \} \\ & s.t.\ (\Phi_i^{(l)} = \Phi_j^{(l)}) \cup (\mathrm{s}_1 = \mathrm{s}_2) \neq 0. \end{aligned} \tag{7}$$

Based on the above-mentioned switching criterion in HSM, to improve the system performance further, we combine ESM-EDAS with the conventional SM, namely, HSM-EDAS. It is described as follows: For each channel realization, compute the minimum square Euclidean distance of ESM using the channel matrix $\mathbf{H}^{(l_{selected})}$ selected from (7) and that of SM according to (4) and then compare the Euclidean distances with each other to choose the transmission scheme with the larger one. One can conclude from (7) that the complexity order imposed by computing $l_{selected}$ is $O(nN_t^2M^2)$. In the next section, it is shown that the complexity order can be reduced. It is explained here that the complexity order is considered as the number of times the optimization metric is iterated. Thus, minimizing the order can minimize the overall computational complexity.

4.2 Low-Complexity HSM-EDAS schemes

In view of the high complexity occurred in antenna selection algorithm EDAS, two low-complexity algorithms are presented to improve the applicability of HSM-EDAS in practical systems in this section.

4.2.1 Reduced-Complexity HSM-EDAS scheme (RC-HSM-EDAS)

A reduced-complexity HSM-EDAS scheme, referred as RC-HSM-EDAS is developed by simplifying (7). Consider an upper triangular matrix $D^{(l)} \in \mathbb{C}^{\phi\times\phi}$, where $\phi = \left|\Phi^{(l)}\right| = 2^{N_t} - 1$ whose entry $(i, j)^{th}$ is given by $D_{i,j}^{(l)}$

$$D_{i,j}^{(l)} = \begin{cases} \min\limits_{s_1,s_2\in S} \left\| \sum\limits_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} \right\|_F^2 \left|s_1 - s_2\right|^2, & i = j \quad (8.a) \\ \min\limits_{s_1,s_2\in S} \left\| \sum\limits_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} s_1 - \sum\limits_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)} s_2 \right\|_F, & i \neq j \quad (8.b) \end{cases} \tag{8}$$

then the computation of $l_{selected}$ can be rewritten as

$$l_{selected} = \arg\max_{l\in\Gamma}\{\min D^{(l)}\} \tag{9}$$

where $\min D^{(l)}$ represents the non-zero elements of matrix $D^{(l)}$.

For $i \neq j$, (8.b) can be written as

$$D_{i,j}^{(l)} = \min_{s_1,s_2\in S}\{\|\boldsymbol{\Delta}\|_F^2\} \tag{10}$$

where $\boldsymbol{\Delta} = \overline{\mathbf{H}}s$, in which

$$\overline{\mathbf{H}} = \sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} - \sum_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)}, \qquad s = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix}.$$

Taking QR decomposition of $\overline{\mathbf{H}}$ with $\mathbf{Q}$ and $\mathbf{R}$, where $\mathbf{Q}^H\mathbf{Q}$ equals the identity matrix $\mathbf{I}$ and $\mathbf{R}$ is an (4×4) upper triangle matrix. Then, (8.b) can be rewritten as

$$D_{i,j}^{(l)} = \min_{\substack{s_{1I},s_{2I}\in N_1-PAM \\ s_{1Q},s_{QI}\in N_2-PAM}} \left\| \mathbf{R} \begin{bmatrix} s_{1I} \\ s_{1Q} \\ -s_{2I} \\ -s_{2Q} \end{bmatrix} \right\|_F^2, i \neq j \tag{11}$$

where s_{iI} and s_{iQ} denotes respectively the real and imaginary part of s_i, the modulation order $M = N_1N_2$. Considering the symmetry of M-QAM constellation, (8) can be further simplified.

Firstly, in (8.a), the minimum distance is required for $s_1 \neq s_2$. The minimum distance between any two symbols for QAM constellation is $d_{\min}^{QAM} = 2$. Hence, equation (8.a) can be replaced by

$$D_{i,j}^{(l)} = (d_{\min}^{QAM})^2 \left\| \sum_{i \in \Phi_i^{(l)}} \mathbf{h}_p^{(l)} \right\|_F^2, i = j. \tag{12}$$

Secondly, in (11), the metric should be computed for all constellation points, i.e. $s_{1I}, s_{2I} \in N_1 - PAM$, $s_{1Q}, s_{2Q} \in N_2 - PAM$. However, if $s_{1I} = \alpha$ and $s_{1Q} = \beta$ are replaced by $s_{2I} = -\alpha$ and $s_{2Q} = -\beta$, the result should be equivalent since $\|\mathbf{s}\|_F^2 = \|-\mathbf{s}\|_F^2$. Thus, (11) can be rewritten as

$$D_{i,j}^{(l)} = \min_{\substack{s_{1I}, s_{2I} \in M_1 \\ s_{1Q}, s_{2Q} \in M_2}} \left\| \mathbf{R} \begin{bmatrix} s_{1I} \\ s_{1Q} \\ -s_{2I} \\ -s_{2Q} \end{bmatrix} \right\|_F^2, i \neq j \tag{13}$$

where M_1 and M_2 represent half of the PAM constellation sets N_1 and N_2, respectively. Hence, the computation complexity of the $(i, j)^{th}$ element of $D^{(l)}$ is decreased from $O(\hat{M})$ to $O(1)$ where $\hat{M} = \binom{M}{2}$ for $i = j$ and from $O(N_1 N_2)$ to $O(\frac{1}{2} N_1 N_2)$ for $i \neq j$.

4.2.2 Further-Reduced-Complexity HSM-EDAS scheme (FRC-HSM-EDAS)

In this subsection, we will show the complexity order in RC-ESM-EDAS can be further reduced. Since the two equations $\Phi_i^{(l)} = \Phi_j^{(l)}$ and $s_1 = s_2$ in (7) do not hold simultaneously, we may consider this situation in three scenarios. Hence, we distinguish three types of minimum square Euclidean distances, denoted as $d_{signal}^{(l)2}(\mathbf{H})$, $d_{spatial}^{(l)2}(\mathbf{H})$, $d_{joint}^{(l)2}(\mathbf{H})$. To be specific,

1. $\Phi_i^{(l)} = \Phi_j^{(l)}$ and $s_1 \neq s_2$

$$\begin{aligned} d_{signal}^{(l)2}(\mathbf{H}) &= \min_{\Phi_i^{(l)}\in\Phi^{(l)}} \left\| \sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} \right\|_F^2 \min_{s_1,s_2\in S} |s_1 - s_2|^2 \\ &= \min_{p=1,2,\ldots,N_t} \left\| \mathbf{h}_p^{(l)} \right\|_F^2 \min_{s_1,s_2\in S} |s_1 - s_2|^2 \end{aligned} \tag{14}$$

2. $\Phi_i^{(l)} \neq \Phi_j^{(l)}$ and $s_1 = s_2$

$$d_{spatial}^{(l)2}(\mathbf{H}) = \min_{\Phi_i^{(l)},\Phi_j^{(l)}\subset\Phi^{(l)}} \left\| \sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} - \sum_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)} \right\|_F^2 \min_{s_1\in S} |s_1|^2 \tag{15}$$

3. $\Phi_i^{(l)} \neq \Phi_j^{(l)}$ and $s_1 \neq s_2$

$$\begin{aligned} d_{joint}^{(l)2}(\mathbf{H}) &= \min_{\Phi_i^{(l)},\Phi_j^{(l)}\subset\Phi^{(l)}} \left\| \sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} s_1 - \sum_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)} s_2 \right\|_F^2 \\ &\geq \min_{\substack{p\in\Phi_i^{(l)},q\in\Phi_j^{(l)} \\ \Phi_i^{(l)},\Phi_j^{(l)}\subset\Phi^{(l)}}} \delta_{p,q}^2 \min_{s_1,s_2\in S} \left\| \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} \right\|_F^2 \end{aligned} \tag{16}$$

where we have

$$\min_{\substack{p\in\Phi_i^{(l)},q\in\Phi_j^{(l)} \\ \Phi_i^{(l)},\Phi_j^{(l)}\subset\Phi^{(l)}}} \delta_{p,q}^2 = \sigma\left(\sum_{p\in\Phi_i^{(l)}} \mathbf{h}_p^{(l)} - \sum_{q\in\Phi_j^{(l)}} \mathbf{h}_q^{(l)} \right).$$

Consequently, on can formulate the minimum Euclidean distance as follows

$$d_{min,ESM}^{(l)2}(\mathbf{H}) = \min\{d_{signal}^{(l)2}(\mathbf{H}), d_{spatial}^{(l)2}(\mathbf{H}), d_{joint_{bound}}^{(l)2}(\mathbf{H})\} \tag{17}$$

where $d_{joint_{bound}}^{(l)2}(\mathbf{H})$ can be obtained by (16). Therefore, FRC-HSM-EDAS selects the antenna subset with the largest minimum square Euclidean distance among all possible candidates. Hence,

$$l_{selected} = \arg\max_{l\in\Gamma}\{d_{min,ESM}^{(l)2}(\mathbf{H})\}. \tag{18}$$

4.3 Low-Complexity HSM-COAS schemes

It can be observed from (17) that even after the reduction of complexity, EDAS-based hybrid spatial modulation scheme still suffers high computation complexity which is relevant to both the number of transmit antennas and the constellation size. For example, we assume the constellation order is fixed and there exist N_t out of N_s antennas are selected. In order to select the best antenna combination, equation (17) has to be computed $\binom{N_s}{N_t}$ times and compared with each other. Thus, the computation complexity increases with the number of transmit antennas. Aiming to solve the complexity problem, combined with COAS technique, a COAS based ESM scheme, referred as ESM-COAS is proposed.

Under a given SNR and a fixed channel realization with N_t transmit antennas, ESM scheme has system capacity bounded as

$$C_{ESM} \geq \gamma = \frac{1}{N_t} \sum_{p=1}^{N_t} \log_2 \left(1 + \frac{E_s}{N_0} \left\| \mathbf{h}_p^{(l)} \right\|_F^2 \right) \tag{19}$$

where E_s is the mean power of transmit vector, N_0 is the power of the noise. It may be readily seen in (19) that one can increase the value of γ to enhance the capacity by choosing N_t antennas according to the largest channel norms out of the N_s transmit antennas. The set of antenna indices can be selected at the receiver and be fed back to the sender and can be obtained by

$$\begin{aligned} &l_{\text{selec}ted} = \left\{ p_1, p_2, \ldots, p_{N_t} \right\} \\ s.t. \quad &\| \mathbf{h}_{p_1} \|_F^2 > \| \mathbf{h}_{p_2} \|_F^2 > \ldots > \| \mathbf{h}_{p_{N_t}} \|_F^2 > \ldots > \| \mathbf{h}_{p_{N_s}} \|_F^2 . \end{aligned} \tag{20}$$

Finally, a novel transmission scheme (HSM-COAS) by combining ESM-COAS and SM is developed according to the value of Euclidean distance. The scheme can be described as follows: For each channel condition, we first choose N_t columns with the largest channel norms to establish a new sub-channel with N_r rows and N_t columns, then compute $d_{min,SM}^2(\mathbf{H})$ and $d_{min,ESM}^2(\mathbf{H})$ according to (4) to choose the scheme with larger minimum Euclidean distance.

5. Complexity Analysis

The purpose of this section is to investigate the computation complexity required for performing the EDAS and COAS based hybrid spatial modulation schemes. Here we consider each complex addition, subtraction, multiplication, division, or square root as one flop [26]. As we can observe that the complexities of SM in all HSM schemes are the same and we only need to compare the complexity of ESM. We first define

$$\Psi=\{\Phi_1,\Phi_2,...,\Phi_\omega\},\ \omega=\sum_{i=1}^{N_t}\binom{N_s}{i}$$

as the set of antenna indices for all *n* antenna combinations by deleting repeated ones. If $N_t=2, N_s=4,$ and then $\Psi=\{1,2,3,4,\{1,2\},\{1,3\},\{1,4\},\{2,3\},\{2,4\},\{3,4\}\}$.

Firstly, according to (7) for HSM-EDAS, the exhaustive search requires a number of flops as

$$\eta_{HSM-EDAS}=\sum_{\Phi_i,\Phi_j\subset\Psi}[(|\Phi_i|+|\Phi_j|+3)N_r-1] \tag{21}$$

where $|\Phi_i|$ and $|\Phi_j|$ denotes the cardinality of Φ_i and Φ_j, respectively.

Secondly, in RC-HSM-EDAS, based on (8), the main diagonal elements of the matrix *D* require

$$\sum_{\Phi_i^{(l)}\in\Phi^{(l)}}[(|\Phi_i^{(l)}|+1)N_r-1]$$

complex operations. The off-diagonal ones need QR decompositions, which needs

$$\sum_{\substack{\Phi_i^{(l)},\Phi_j^{(l)}\subset\Phi^{(l)}\\ \Phi_i^{(l)}\neq\Phi_j^{(l)}}}[(|\Phi_i^{(l)}|+|\Phi_j^{(l)}|+62)N_r-\frac{2}{3}64]$$

flops for each element. Hence, considering all the possible combinations of the transmit vectors, the number of flops required to operate RC-HSM-EDAS is

$$\eta_{RC-HSM-EDAS}=\sum_{\Phi_i\subset\Psi}\left[(|\Phi_i|+1)N_r-1\right]+\sum_{\substack{\Phi_i,\Phi_j\subset\Psi\\ \Phi_i\neq\Phi_j}}\left[(|\Phi_i|+|\Phi_j|+62)N_r-\frac{2}{3}64\right] \tag{22}$$

Thirdly, FRC-HSM-EDAS imposes a complexity as follows:

- The number of flops for the signal part is $N_s(2N_r-1)$;
- The number of flops for spatial part is

$$\sum_{\substack{\Phi_i,\Phi_j\subset\Psi\\ \Phi_i\neq\Phi_j}}[(|\Phi_i|+|\Phi_j|+1)N_r-1]$$

- Based on the Householder method [26, Eq. (31.4)] employed for performing singular value decomposition, the number of flops required for the joint part is then

$$\sum_{\substack{\Phi_i,\Phi_j\subset\Psi\\ \Phi_i\neq\Phi_j}}[(|\Phi_i|+|\Phi_j|+14)N_r-\frac{16}{3}].$$

Hence, the total number of flops of FRC-HSM-EDAS is

$$\eta_{FRC-HSM-EDAS} = N_s(2N_r - 1) + \sum_{\substack{\Phi_i,\Phi_j \subset \Psi \\ \Phi_i \neq \Phi_j}} \left[(2|\Phi_i| + 2|\Phi_j| + 15) N_r - \frac{19}{3} \right]. \quad (23)$$

Finally, for HSM-COAS, the computation complexity is

$$\eta_{HSM-COAS} = N_s(2N_r - 1). \quad (24)$$

In order to have a more intuitive understanding of the complexity of each algorithm, Table 1 shows the numbers of flops required for HSM-EDAS, RC-HSM-EDAS, FRC-HSM-EDAS and HSM-COAS with 4-QAM. As we can see, the reduction in the number of flops required for calculating RC-HSM-EDAS, FRC-HSM-EDAS and HSM-COAS is significant. Specifically, RC-HSM-EDAS or FRC-HSM-EDAS achieves a reduction of nearly 20% and 80% compared with exhaustive-search-based HSM-EDAS. Moreover, although HSM-COAS achieves a reduction of almost 100% in complexity, it suffers a serious performance loss compared with the other three schemes.

Table 1

Example of comparing the computational complexity with $N_s = 6$, $N_t = 3$, $N_r = 2$ and $M = 4$.

	HSM-EDAS	RC-HSM-EDAS	FRC-HSM-EDAS	HDSM-COAS
Number of flops	188480	148979	34758	18

6. Performance Simulation Analysis

Computer simulation experiments are performed in this section for evaluating the performances of the presented schemes HSM, EDAS-based HSM and HSM-COAS. In all cases, we assume that the power is assigned equally for each available transmit antenna and the channel state information of the Rayleigh flat-fading channel is totally known at the receiver. One hundred thousand independent channel realizations are utilized, and each element of MIMO channels is assumed to be independent identically distributed random variable that follows complex Gaussian distribution with zero mean and unit variance. In addition, the simulation operates in an environment with single user and the total number of receive antennas that are available is 2.

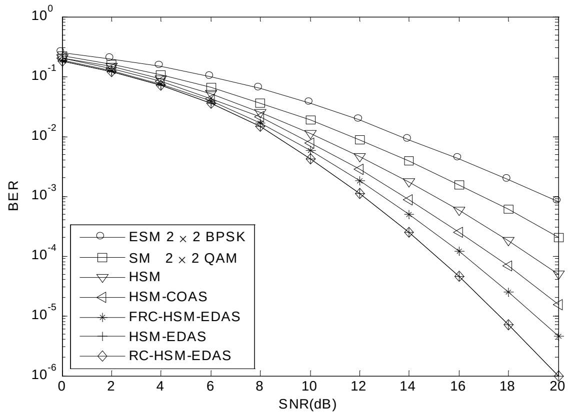

Fig. 3. Experimental results of Performances with spectral efficiency 3 b/s/Hz and 2 out of 4 transmit antennas.

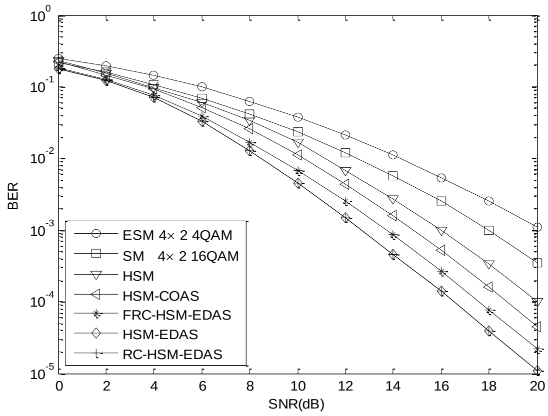

Fig. 4. Exp erimental results of performances with spectral efficiency 6 b/s/Hz and 4 out of 6 transmit antennas.

Fig. 3 and Fig. 4 show the BER comparison of the conventional ESM, SM and the proposed HSM policies under different spectral efficiency. As expected, HSM outperforms SM and ESM by 2 dB and 4 dB at 10^{-3} BER respectively which can be seen in Fig. 3 because the system performance decreases monotonically with the minimum Euclidean distance and HSM always selects larger minimum Euclidean distance between SM and ESM for each channel realization. In addition, EDAS-based hybrid schemes and HSM-COAS perform better than HSM as a result of the antenna selection increasing the value of $d^2_{\min,ESM}(\mathbf{H})$. Specially, proposed RC-HSM-EDAS performs almost the same as HSM-EDAS (the two curves in the figures almost coincide) and FRC-HSM-EDAS performs worse with lower complexity. It is worth noting that the EDAS-based hybrid schemes all outperform HSM-COAS scheme with at least 1 dB SNR gain at 10^{-3} BER. The reason for this result is that although HSM-COAS reduces the computational complexity as analyzed in the previous section, it suffers performance loss compared with EDAS-based hybrid schemes. Finally, comparing Fig. 3 and Fig. 4, one can find that increasing the spectral efficiency decreases the performance owing to the increased order of the constellation.

7. Conclusions

In this paper, a novel MIMO transmission policy named HSM is put forward, which combines ESM and SM according to the minimum Euclidean distance. HSM can select a suitable transmission scheme between ESM and SM according to the changing channel state information, which improves the system performance. Based on HSM, we present two improved versions by employing antenna selection, i.e., Euclidean distance antenna selection-based HSM (named HSM-EDAS) and capacity optimization antenna selection-based HSM (named HSM-COAS). Furthermore, in order to reduce the high computation complexity occurred in HSM-EDAS, low-complexity EDAS algorithms are then proposed for HSM scheme, named as RC-HSM-EDAS and FRC-HSM-EDAS. Theoretical analysis and simulation results demonstrate that the presented HSM scheme outperforms the existing SM or ESM and it performs better if combined with COAS or EDAS antenna selection policy. Additionally, the performance of HSM-EDAS may be influenced slightly after reducing complexity. On the other hand, all EDAS-based hybrid schemes outperform HSM-COAS whereas HSM-COAS achieves lower computation complexity and better tradeoffs between capacity and performance.

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