A Novel Spatial Modulation Scheme over Correlated Fading Channels

Peng Zhang, Dongfeng Yuan, Haixia Zhang
School of Information Science and Engineering, Shandong University, Jinan, P. R. China
Email: sduisezhangp@mail.sdu.edu.cn, sduzhangp@163.com, {dfyuan, haixia.zhang}@sdu.edu.cn

Abstract— **A novel spatial modulation (SM) scheme suitable for correlated Rician fading scenario is proposed with the help of the key idea of Ungerboeck's set partitioning. The motivation is based on the fact that the different error performance exists between the antenna domain and modulation signal domain in the traditional SM scheme. And the total system error performance is mainly determined by the worse case. We analyze the unequal error protection (UEP) performance and find that the error performance of the antenna selection is mainly determined by the correlation coefficient and Rician factor. If the correlation coefficient or Rician factor is large, the error performance will be bad. Considering the error performance of the modulation signal domain is mainly determined by the minimum Euclidean distance between pair of signal points, the error performance of two domains will differ significantly over the fading channels with strong correlation or large Rician factor. The main idea of the novel scheme is to establish the relationship between the antenna domain and modulation signal domain by expanding the signal constellation to carry all the input information bits. The expanded constellation is then partitioned into subsets by using Ungerboeck's set partitioning. The bits mapping to the antenna index are also used to select the partitioned subset, while the remaining bits are used to determine the transmit signal in the selected partitioned subset. In this way, the error performance of the bits mapping to the antenna index is improved, while the error performance of the other bits is preserved by maximizing the minimum Euclidean distance between any pair of signal points in the partitioned subset. Performance analysis and simulation results show that the novel SM scheme could improve the error system performance greatly when the correlation coefficient or Rician factor become larger.**

Index Terms— **MIMO, spatial modulation, set partitioning, spatial correlation, unequal error protection (UEP)**

I. Introduction

Spatial modulation (SM) is recently proposed to further improve data rates of Multiple-Input-Multiple-Output (MIMO) systems by joint exploiting the potential of both antenna space and signal space [1], [2]. With SM, at each time slot, only one transmit antenna is activated. Consequently, inter-channel interference (ICI) caused by multiple antennas can be avoided. Meanwhile, such a structure of SM makes the decoding at receiver much simple and Maximum-Likelihood (ML) decoding [3] can be employed to get optimal performance. SM is actually a novel 3-dim modulation (spatial plus two-dim modulation) concept for MIMO systems. As a special case, SM can be taken as Space Shift Keying (SSK) if transmitted signals are carried by only the transmit antenna index [4].

Since being proposed, most of the literatures concerning SM are mainly focused on the performance analysis over different channels. [2] analyzed the performance when MRRC (maximum receive ratio combining) detection is employed over Rayleigh fading channels. [3] proved that SM with MRRC is suboptimal and in the same time proposed an optimal ML detection method. In [3], an upper bound of bit error rate was also derived by using the union bound method. Other related work could be found in [5], [6] for SSK modulation, and in [7], [8] for SM schemes with only one receive antenna. Recently, [9] extended the performance analysis result of [7], [8] to the general SM schemes with multiple receive antennas and arbitrary complex modulation schemes. To the best of our knowledge, till now little attention has been paid on effect of a single domain on the whole performance except for [2], which has compared the performance of SM by assuming the antenna selection domain and modulation signal domain to be independent. Generally speaking, this is not correct, especially for the correlated scenarios. On the other hand, the previous works, such as [10], [11], have demonstrated that the UEP (unequal error protection) properties really exist between the antenna space and signal space, especially over correlated channel conditions. In order to tackle this unbalance, trellis coding has been proposed in [10], [11]. However, in [10] only the bits for antenna selection are convolutional encoded, which could provide no coding gain for the other source bits. As an alter scheme, [11] introduced the jointly trellis coding design method by considering antenna space and signal space together, but it is really hard work to design a good trellis code.

In this article, we develop a novel SM scheme from the perspective of spatial modulation only. The relationship between the antenna space and signal space is established, through which the two components could help each other to overcome the inherent UEP properties of the traditional SM scheme. More specially, the novel scheme is mainly used in the spatial correlated channel conditions, where the performance of antenna space will be much worse

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Correspondence should be addressed to Dongfeng Yuan, dfyuan@sdu.edu.cn.

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than that of the signal space. In order to improve the error performance of antenna space, we try to allocate the same bits used to select the transmit antenna onto the signal space, therefore, an expanded signal constellation is constructed compared with the traditional SM scheme. In our work, the novel expanded constellation will carry both the bits for antenna selection and the bits for traditional modulation signal. The question is as the constellation gets larger, and the minimum distance between pair of signals will get smaller. Hence, the next trick is attempting to keep the error performance of the bits corresponding to the modulation signal in the traditional SM scheme. This design intent coincides with the key ideas of the Ungerboeck's set partitioning rule, which is used for the design of trellis coded modulation (TCM) schemes [12]. Therefore, a novel SM scheme is naturally constructed based on the Ungerboeck's set partitioning rule, by which an improved error performance is achieved compared with the traditional SM scheme over correlated fading channels.

The rest of the paper is organized as follows: In Section II, the traditional SM approach is introduced and the general SM model is presented. Following which the general correlated/uncorrelated Rayleigh/Rician fading channel model is provided in Section III, and the novel SM scheme based on the Ungerboeck's set partitioning rule is proposed in Section IV. In Section V, the performance of the two SM schemes are comparably studied, and the UEP performance is also discussed. Simulation results are presented in Section VI. Finally, Section VII concludes the paper.

Notation: In this paper, N_t and N_r denote the number of transmit and receive antennas, respectively. Bold, lowercase and capital letters denote column vectors and matrices, respectively. $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ stand for complex conjugation, transposition and complex conjugation transposition, respectively. $\det(\mathbf{A})$ and $\mathbf{A}^{-1}$ denotes the determinant and inverse of $\mathbf{A}$, respectively. $E(\cdot)$ denotes the expectation operation. $\|\cdot\|_F$ accounts for the Frobenius norm. $\mathbb{C}_M$ represents the complex modulation signal space with the cardinality equal to M. $\Re(\cdot)$ is to get the real part of the expression.

II. Traditional Spatial Modulation

As shown in Fig. 1, we consider a SM system model with N_t transmit and N_r receive antennas. Moreover, we assume that the number of transmit antennas is an integer power of 2, i.e., $N_t = 2^n$. The MIMO channel matrix is given by

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1N_t} \\ h_{21} & h_{22} & \cdots & h_{2N_t} \\ \cdots & \cdots & \ddots & \cdots \\ h_{N_r1} & h_{N_r2} & \cdots & h_{N_rN_t} \end{bmatrix}, \tag{1}$$

where $h_{ij},\ i = 1, 2, \ldots, N_r,\ j = 1, 2, \ldots, N_t$ denotes the channel coefficient from the transmit antenna i to the receive antenna j. Let $\mathbf{h}_j$ represent the jth column of $\mathbf{H}$

$$\mathbf{h}_j = \begin{bmatrix} h_{1j} & h_{2j} & \cdots & h_{N_rj} \end{bmatrix}^T,\ j = 1, 2, \ldots, N_t, \tag{2}$$

which denotes the channel coefficients corresponding to the transmit antenna j.

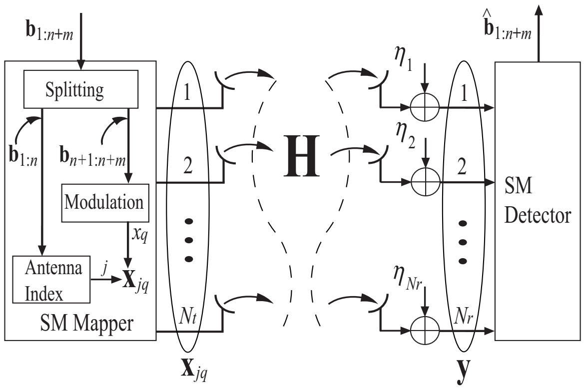

Figure 1. Spatial Modulation System Model

For the traditional SM scheme, we assume M-ary digital modulation is employed at the transmitter with $M = 2^m$, and the modulation set is denoted as $\mathbb{C}_M = \{x_1, x_2, \ldots, x_M\}$. And then the spectral efficiency provided by the SM system is

$$R = \log_2(N_t M) = n + m. \tag{3}$$

At transmitter, the information bits are split into groups, and each group includes $n + m$ bits. Then the two subgroups with n and m bits, respectively, are generated by further splitting each group. Finally, the subgroup with n bits is used to select antenna and the other subgroup with m bits is modulated onto one complex constellation point according to the specified digital modulation method.

For example, as shown in Fig. 1, the input information is splitted into groups with each one denoted as

$$\mathbf{b}_{1:n+m} = \begin{bmatrix} b_1 & \ldots & b_n & b_{n+1} & \ldots & b_{n+m} \end{bmatrix}^T, \tag{4}$$

which are further splitted into two subgroups, the fisrt n bits

$$\mathbf{b}_{1:n} = \begin{bmatrix} b_1 & b_2 & \ldots & b_n \end{bmatrix}^T, \tag{5}$$

and the rest m bits

$$\mathbf{b}_{n+1:n+m} = \begin{bmatrix} b_{n+1} & b_{n+2} & \ldots & b_{n+m} \end{bmatrix}^T. \tag{6}$$

The first subgroup is used to select the transmit antennas with the antenna index calculated as

$$j = \sum_{u=1}^{n} b_u 2^{n-u} + 1. \tag{7}$$

The second subgroup is help to select the modulation symbols with the symbol index calculated as

$$q = \sum_{u=1}^{m} b_{n+u} 2^{m-u} + 1. \tag{8}$$

Suppose that the jth transmit antenna and the modulation constellation point of $x_q \in \mathbb{C}_M$ are selected for transmission, the output of the SM mapper could be defined as

$$\mathbf{x}_{jq} \triangleq \begin{bmatrix} 0 & \dots & 0 & x_q & 0 & \dots & 0 \\ & & & \uparrow & & & \\ & & & j\text{th position} & & & \end{bmatrix}^T, \quad (9)$$

which is a N_t-dim vector with j denoting the activated antenna. And then the received signals are given by

$$\mathbf{y} = \sqrt{\gamma}\mathbf{H}\mathbf{x}_{jq} + \eta = \sqrt{\gamma}\mathbf{h}_j x_q + \eta, \quad (10)$$

where γ denots the average signal to noise ratio (SNR) at each receive antenna, and $\eta = \begin{bmatrix}\eta_1 & \eta_2 & \cdots & \eta_{N_r}\end{bmatrix}^T$ is an N_r-dim additive white Gaussian noise (AWGN), each element of which is of independent and identically Gaussian distribution with zero mean and unit variance.

According to [3], the optimal detector based on the ML (Maximum Likelihood) principle is given by

$$\begin{aligned}[\hat{j}_{\mathrm{ML}}, \hat{q}_{\mathrm{ML}}] &= \arg\max_{j,q} p_{\mathbf{Y}}\left(\mathbf{y}|\mathbf{x}_{jq}, \mathbf{H}\right) \\ &= \arg\min_{j,q} \|\mathbf{y} - \mathbf{H}\mathbf{x}_{jq}\|_F^2, \quad (11)\end{aligned}$$

since the PDF (probability density function) of $\mathbf{y}$ conditioned on $\mathbf{H}$ and $\mathbf{x}_{jq}$ is

$$p_{\mathbf{Y}}\left(\mathbf{y}|\mathbf{x}_{jq}, \mathbf{H}\right) \propto \exp\left(-\|\mathbf{y} - \mathbf{H}\mathbf{x}_{jq}\|_F^2\right). \quad (12)$$

Obviously, a number of advantages could be achieved by adopting the SM approach. Firstly, the ICI is avoided since only one antenna is activated. Correspondingly the decoding complexity at the receiver is also reduced. And then, for a fixed spectral efficiency, the modulation constellation size is reduced by utilizing antenna index to carry some information bits, and hereby the minimum distance between pair of constellation points is increased and the improved error performance is achieved.

For the traditional SM scheme, there are no relationship between the information bits for antenna space and the information bits for modulation space. From the following analysis, we will know that these two independent bits will appear different error performance. Especially over correlated fading channels, their performance will differ greatly, and the information bits for antenna selection will have a higher error rate than the information bits for modulation. To address this issue, trellis coded spatial modulation (TCSM) schemes have been proposed in [10], [11] by jointly designing the trellis encoder and the SM mapper together. In which, [10] have applied the key idea of TCM [12] to the antenna domain only, whereas in [11] the total SM bits are jointly encoded to pursue optimized diversity and coding gain.

In this paper, we propose a novel SM design method from the viewpoint of SM. In order to improve the error performance of the information bits for antenna selection in the traditional SM, one expanded modulation constellation is adopted to carry all the information bits, as a result, the information bits for antenna selection will be protected by the antenna selection and the modulation symbols together. Moreover, by using the Ungerboeck's set partitioning, the error performance the information bits for the modulation signal in the traditional SM is preserved as much as possible. Consequently, the total error performance of SM could be improved.

III. Generalized Rician fading channel model

As shown in [9], [10], the generalized Rician fading channel is modeled as

$$\mathbf{H} = \sqrt{\frac{K}{K+1}}\bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}}\tilde{\mathbf{H}}. \quad (13)$$

where $\sqrt{\frac{K}{K+1}}\bar{\mathbf{H}}$ represents the fixed LOS (line-of-sight) component, while $\sqrt{\frac{1}{K+1}}\tilde{\mathbf{H}}$ denotes the variable fading component caused by the NLOS (Non-line-of-sight) scatter signals. K is the Rician factor which reflects the power ratio of the fixed and variable parts. Each element of $\bar{\mathbf{H}}$ is equal to one to represent the LOS signal. The scatter component matrix $\tilde{\mathbf{H}}$ is modeled by using the Kronecker model [13], [14], i.e.,

$$\tilde{\mathbf{H}} = \mathbf{R}_r^{1/2}\breve{\mathbf{H}}\mathbf{R}_t^{1/2}, \quad (14)$$

where $\mathbf{R}_t$ and $\mathbf{R}_r$ denotes the correlation matrices at the transmitter and receiver, respectively. $\breve{\mathbf{H}}$ is a $N_r \times N_t$ matrix whose entries are independently distributed according to the complex Gaussian distribution with zero mean and unit variance. The transmit correlation matrix $\mathbf{R}_t$ is formed as

$$\mathbf{R}_t = \begin{bmatrix} \varphi_{11}^t & \varphi_{12}^t & \cdots & \varphi_{1N_t}^t \\ \varphi_{21}^t & \varphi_{22}^t & \cdots & \varphi_{2N_t}^t \\ \cdots & \cdots & \ddots & \cdots \\ \varphi_{N_t1}^t & \varphi_{N_t2}^t & \cdots & \varphi_{N_tN_t}^t \end{bmatrix} \quad (15)$$

with $\varphi_{ij}^t,\ i,\ j \in \{1, 2, \dots, N_t\}$ representing the cross correlation between the channel coefficients of the two transmit antennas indexed by i and j.

In a similar way, the receive correlation matrix $\mathbf{R}_r$ is written as

$$\mathbf{R}_r = \begin{bmatrix} \varphi_{11}^r & \varphi_{12}^r & \cdots & \varphi_{1N_r}^r \\ \varphi_{21}^r & \varphi_{22}^r & \cdots & \varphi_{2N_r}^r \\ \cdots & \cdots & \ddots & \cdots \\ \varphi_{N_r1}^r & \varphi_{N_r2}^r & \cdots & \varphi_{N_rN_r}^r \end{bmatrix}, \quad (16)$$

in which $\varphi_{ij}^r,\ i,\ j \in \{1, 2, \dots, N_r\}$ denoting the cross correlation between the channel coefficients related to the two receive antennas indexed by i and j.

By using the above generalized fading model of (13), the uncorrelated/correlated Rayleigh/Rician fading channels could be easily constructed via the appropriate selection of parameters. The channel correlation matrices in the Kronecker model are frequently calculated according to two common approaches. One is presented in [15] and the correlation matrices are computed based on a clustered channel model using the power azimuth spectrum (PAS) distribution and the array geometry. The other is the exponential correlation model of [16], where the correlation matrix entries are formed as

$$\begin{aligned}\varphi_{ij}^t &= \rho_t^{|i-j|},\ i, j \in \{1, 2, \dots, N_t\} \\ \varphi_{ij}^r &= \rho_r^{|i-j|},\ i, j \in \{1, 2, \dots, N_r\}, \quad (17)\end{aligned}$$

in which ρ_t stands for the correlation coefficients between two adjacent transmit antennas, and ρ_r denotes the correlation coefficients between two adjacent receive antennas.

IV. Novel Spatial Modulation

A. Novel spatial modulation design

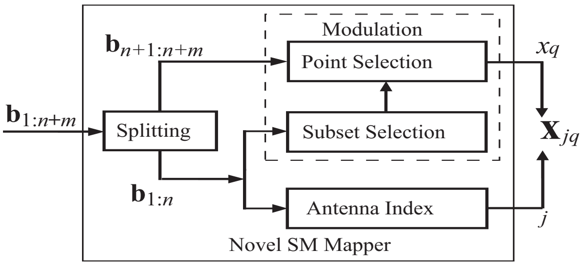

Figure 2. Novel Spatial Modulation Mapper

With the SM system model as shown in Fig. 1, we propose a novel spatial modulation scheme, in which the SM mapper is re-designed according to Fig. 2. The main difference between the newly designed modulation scheme and the traditional one lies in the different operation of the modulation block compared with the SM mapper as shown in Fig. 1.

For the novel SM mapper, similarly the input $n+m$ bits are splitted into two subgroups according to (5) and (6). Then, the antenna index is determined according to the bit sequence of $\mathbf{b}_{1:n}$ similar with the traditional SM operation. And the modulated output is determined by all the input bits $\mathbf{b}_{1:n+m}$, i.e, $x_q \in \mathbb{C}_{2^{n+m}}$, which is different from the traditional SM approach as shown in Fig. 1, where the modulation symbol is determined by $\mathbf{b}_{n+1:n+m}$, i.e., $x_q \in \mathbb{C}_{2^m}$. In our work, all the input bits are included in the modulation constellation and the bit-to-constellation mapping is based on the Ungerboeck's set partitioning rule [12]. After n-step Ungerboeck's set partitioning, 2^n subsets are produced. Elements of them are from the modulation constellation of $\mathbb{C}_{2^{n+m}}$. To modulate, one subset will first be chosen according to $\mathbf{b}_{1:n}$, and then the constellation point will be determined by $\mathbf{b}_{n+1:n+m}$. Finally, we will get the modulated x_q which is jointly determined by $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$.

In this paper, by applying the key idea of Ungerboeck's set partitioning to the bit-to-constellation mapping of x_q, we mainly try to fulfill the following two key objectives: Firstly, try to improve the error performance of $\mathbf{b}_{1:n}$ by allocating them as the antenna selection as well as selecting one of the partitioned subsets. Secondly, via comparing the proposed scheme with the traditional SM approach, the new SM mapper should strive to preserve the error performance of $\mathbf{b}_{n+1:n+m}$, hence they are allocated to a signal point in one partitioned subset, in which the minimum Euclidean distance between pair of signals is enlarged as much as possible by using Ungerboeck's set partitioning.

For the sake of brevity, we define the ith partitioned subset after n-steps partitioning operation as $\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}$, which is partitioned from $\mathbb{C}_{2^{n+m}}$, and i is the integer value of the n bits used to select the partitioned subset. For example, suppose that there are 4 transmit antennas, i.e., $N_t = 4$ $(n = 2)$, and 4 bps/Hz $(n+m = 4)$ spectrum efficiency is required. For the conventional SM mapper, as shown in Fig. 1, QPSK is employed. While for the new SM mapper shown in Fig. 2, 16QAM should be exploited to carry 4 information bits and the bit-to-constellation mapping is demonstrated as Fig. 3.

Obviously, with the novel spatial modulation operation, the performance of $\mathbf{b}_{n+1:n+m}$ is mainly determined by the minimum Euclidean distance between pair of signals in the partitioned subset of $\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}$. While the performance of $\mathbf{b}_{1:n}$ is mainly determined by both the transmit antenna selection and the partitioned subset selection. With respect to the transmit antenna selection, the probability density function of the transmit channel coefficients and the correlation between them will have main effect on the decoding performance, while for the the partitioned subset selection, the decoding performance is mainly determined by the the minimum Euclidean distance between pair of the partitioned subsets, that is equal to the minimum Euclidean distance between pair of signals in $\mathbb{C}_{2^{n+m}}$. So by the new design, the decoding performance of $\mathbf{b}_{1:n}$ is improved, especially in the case of correlated channels.

However, it should be noted that the minimum Euclidean distance between pair of signals in the partitioned subset of $\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}$ may be less than that of the conventional modulation constellation $\mathbb{C}_{2^m}$, which will have a negative impact on the required signal-to-noise ratio (SNR) for correct decoding. In other words, by using the novel SM mapper, the decoding performance of $\mathbf{b}_{n+1:n+m}$ may be worse than the traditional SM approach. But we should note that the total performance of SM is mainly determined by the worse one between $\mathbf{b}_{n+1:n+m}$ and $\mathbf{b}_{1:n}$. When the SM channels become strong correlation, the decoding performance of $\mathbf{b}_{1:n}$ will become very poor. For example, when the correlation coefficient is one, no correct decoding of $\mathbf{b}_{1:n}$ is achieved, while $\mathbf{b}_{n+1:n+m}$ could still be correctly decoded. Hence, we attempt to improve the decoding performance of $\mathbf{b}_{1:n}$ by allocating them in the modulation constellation. At the same time, we will try to enlarge the minimum Euclidean distance between pair of signals in the partitioned subset of $\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}$ in order to preserve the decoding performance of $\mathbf{b}_{n+1:n+m}$ compared to the traditional SM approach.

Compared with the conventional modulation schemes, the SNR loss caused by the Ungerboeck's set partitioning is defined as

$$\gamma_l(\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}) = 10\log_{10}\left(\frac{d^2_{min}(\mathbb{C}_{2^m})}{d^2_{min}(\mathbb{C}^i_{2^m \leftarrow 2^{n+m}})}\right), \quad (18)$$

where $d_{min}(\cdot)$ is to get the minimum Euclidean distance between pair of signals in the specified modulation constellation. Table I illustrates the corresponding SNR loss with some cases as examples. From which we could see

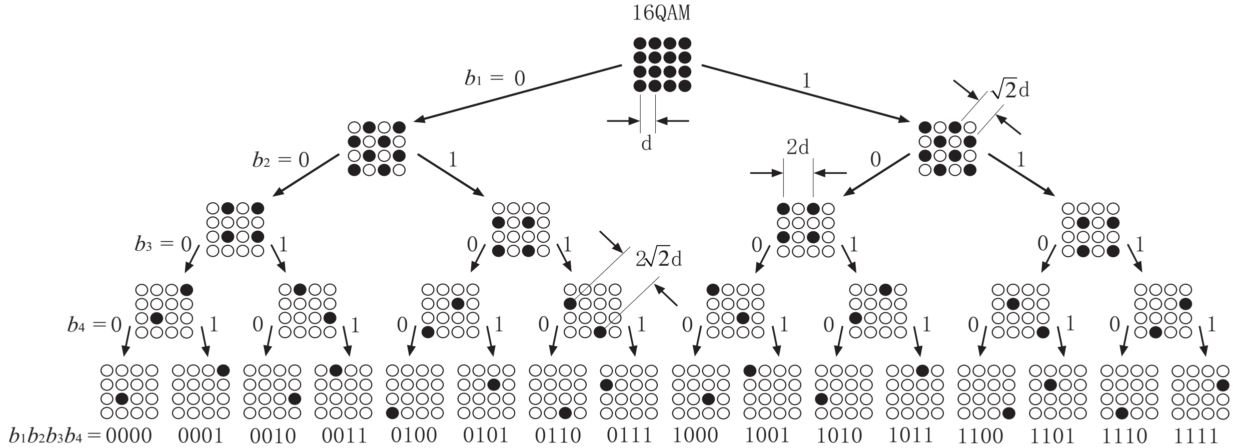

Figure 3. Ungerboeck's Set Partitioning of 16QAM

TABLE I.
THE SNR LOSS CAUSED BY THE UNGERBOECK'S SET PARTITIONING COMPARED WITHE THE CONVETIONAL MODULATION SCHEMES ($\mathbb{C}^i_{2^m \leftarrow 2^{n+m}}$ VS $\mathbb{C}_{2^m}$).

	$\mathbb{C}^i_{2\leftarrow 2^{n+m}}$ vs BPSK	$\mathbb{C}^i_{2^2\leftarrow 2^{n+m}}$ vs QPSK	$\mathbb{C}^i_{2^3\leftarrow 2^{n+m}}$ vs 8PSK	$\mathbb{C}^i_{2^4\leftarrow 2^{n+m}}$ vs 16QAM	$\mathbb{C}^i_{2^5\leftarrow 2^{n+m}}$ vs 32QAM	$\mathbb{C}^i_{2^6\leftarrow 2^{n+m}}$ vs 64QAM	$\mathbb{C}^i_{2^7\leftarrow 2^{n+m}}$ vs 128QAM
QPSK($n+m=2$)	0	0	/	/	/	/	/
8PSK($n+m=3$)	0	0	0	/	/	/	/
16QAM($n+m=4$)	0.97	0.97	-1.35	0	/	/	/
32QAM($n+m=5$)	1.18	1.18	-1.14	0.21	0	/	/
64QAM($n+m=6$)	1.18	1.18	-1.14	0.21	0	0	/
128QAM($n+m=7$)	1.23	1.23	-1.09	0.26	0.05	0.05	0
256QAM($n+m=8$)	1.23	1.23	-1.09	0.26	0.05	0.05	0

that for the partitioned subsets $\mathbb{C}^i_{2\leftarrow 2^{n+m}}$ and $\mathbb{C}^i_{2^2\leftarrow 2^{n+m}}$, there is no SNR loss when MPSK ($M=2^m, m>2$) is partitioned, while about 1dB SNR loss exists when MQAM ($M=2^m, m>3$) is partitioned. Moreover when $\mathbb{C}^i_{2^m\leftarrow 2^{n+m}},\ m>2$ is partitioned from MQAM, only a small SNR loss exists. It is also interesting that some SNR gain could be achieved for some cases, for example, $\mathbb{C}^i_{8\leftarrow 2^{n+m}}, n+m>3$ partitioned from MQAM ($M=2^m, m>3$) will have better performance than 8PSK.

B. Optimal detection for the novel spatial modulation

As similar with (11), the optimal detector based on the ML principle for the novel SM scheme is given by

$$\begin{aligned}[\hat{j}_{\mathrm{ML}}, \hat{q}_{\mathrm{ML}}] &= \arg\max_{j,q} p_{\mathbf{Y}}\left(\mathbf{y}|\mathbf{x}_{jq'}, \mathbf{H}\right) \\ &= \arg\min_{j,q} \|\mathbf{y}-\mathbf{H}\mathbf{x}_{jq'}\|_F^2, \end{aligned} \quad (19)$$

with

$$q' = (j-1)\times 2^m + q.$$

Here, $j \in \{1,2,\ldots,N_t\}$ and $q \in \{1,2,\ldots,M\}$ are determined by $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$, which are given by (7) and (8), respectively. And the transmit SM constellation is defined as

$$\mathbf{x}_{jq'} \triangleq \begin{bmatrix} 0 & \ldots & 0 & x_{q'} & 0 & \ldots & 0 \\ & & & \uparrow & & & \\ & & & j\text{th position} & & & \end{bmatrix}^T \quad (20)$$

with $x_{q'} \in \mathbb{C}_{2^{n+m}}$.

By comparing (19) with (11), it is clear that the novel SM scheme has the same decoding complexity as the traditional SM one. On the other hand, although the modulation block of the novel SM includes two modules as shown in Fig. 2, in terms of architectural or software design these two modules will be one looking-up-table, just as the conventional modulation scheme. Specially, with any given configuration of $n+m$, one bits-to-constellation mapping will be produced by adopting the two proposed operations of "Subset Selection" and "Point Selection". For instance, when $n+m=4$, the bits-to-constellation mapping is produced as shown in Fig. 3. For implementation we only need to store this mapping table. By this way, for each input of $n+m$ bits, one modulation signal point will be looked up. As a result, the complexity comparison of the two schemes is mainly determined by the size of the looking-up-table, which is equal to 2^{n+m} and 2^m for the proposed SM and traditional SM, respectively. In the practical wireless applications, such difference is quite trivial. Consequently, we can say that the proposed scheme has little effect on the system implementation.

V. PERFORMANCE ANALYSIS

A. Performance of the traditional SM scheme

Based on the channel model of (13), [9] has proposed a general method for the error analysis of SM systems over correlated/uncorrelated Rayleigh and Rician fading channels. The related works could also be found in [7], [8], [17]. In the following, we will extend them to the novel proposed SM scheme.

As shown in [9], [17], the upper-bounded average bit error probability (ABEP) is derived by using the well-known union upper bounding technique in [18] for the conventional SM scheme with the optimal detection [3], which is written as

$$\bar{P}_b \leq c \sum_{j=1}^{N_t} \sum_{\hat{j}=1}^{N_t} \sum_{q=1}^{M} \sum_{\hat{q}=1}^{M} \bar{P}_b(j,q;\hat{j},\hat{q}) \tag{21}$$

with

$$\bar{P}_b(j,q;\hat{j},\hat{q}) = \frac{N_b(j,q;\hat{j},\hat{q})}{\log_2(N_t M)} \bar{P}_s(j,q;\hat{j},\hat{q}). \tag{22}$$

In which $c = \frac{1}{N_t M}$, $\bar{P}_s(j,q;\hat{j},\hat{q})$ is the average pairwise symbol error probability (ASEP) of pair of SM symbols denoted by $\mathbf{x}_{jq}$ and $\mathbf{x}_{\hat{j}\hat{q}}$, respectively. $N_b(j,q;\hat{j},\hat{q})$ denotes the Hamming distance between two transmit bit sequences corresponding to $\mathbf{x}_{jq}$ and $\mathbf{x}_{\hat{j}\hat{q}}$, respectively. Notice that $\mathbf{x}_{jq}$ and $\mathbf{x}_{\hat{j}\hat{q}}$ have the definition of (9). According to [9], $\bar{P}_s(j,q;\hat{j},\hat{q})$ is calculated as

$$\bar{P}_s(j,q;\hat{j},\hat{q}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{e^{-\mathbf{m}_z^H \left[\mathbf{I}\sin^2\theta + \mathbf{R}_z\right]^{-1} \mathbf{m}_z}}{\det\left(\mathbf{I} + \frac{\mathbf{R}_z}{\sin^2\theta}\right)} d\theta \tag{23}$$

with

$$\mathbf{m}_z = \sqrt{\frac{\gamma K}{4(K+1)}} (x_q - x_{\hat{q}}) \mathbf{e}_{N_r}, \tag{24}$$

$$\mathbf{R}_z = \frac{\gamma}{4(K+1)} \left[|x_q|^2 + |x_{\hat{q}}|^2 - 2\Re\{\varphi_{j\hat{j}}^t x_q x_{\hat{q}}^*\} \right] \mathbf{R}_r, \tag{25}$$

where $\mathbf{e}_{N_r}$ is an N_r-dim column vector with all entries equal to one.

Similar with (21), we have the total ASEP of the conventional SM scheme, which is written

$$\bar{P}_s \leq c \sum_{j=1}^{N_t} \sum_{\hat{j}=1}^{N_t} \sum_{q=1}^{M} \sum_{\hat{q}=1}^{M} \bar{\delta}(j,q;\hat{j},\hat{q}) \bar{P}_s(j,q;\hat{j},\hat{q}) \tag{26}$$

with $\bar{\delta}(a_1,a_2,\ldots,a_k;b_1,b_2,\ldots,b_k)$ defined as

$$\begin{aligned} &\bar{\delta}(a_1,a_2,\ldots,a_k;b_1,b_2,\ldots,b_k) \\ &= \begin{cases} 0, & if\ a_j = b_j, j = 1,2,\ldots,k \\ 1, & else \end{cases}. \end{aligned} \tag{27}$$

B. Performance of the novel SM scheme

By comparing the two optimal detection formulas of the traditional and novel SM schemes, which are given by (11) and (19), respectively, we could know that the ABEP and ASEP of the novel scheme will have a similar form with the traditional SM scheme. According to (19), (21) and (26), the ABEP and ASEP of the novel SM scheme are given by

$$\bar{P}_{nb} \leq c \sum_{j=1}^{N_t} \sum_{\hat{j}=1}^{N_t} \sum_{q=1}^{M} \sum_{\hat{q}=1}^{M} \bar{P}_{nb}(j,q';\hat{j},\hat{q}') \tag{28}$$

and

$$\bar{P}_{ns} \leq c \sum_{j=1}^{N_t} \sum_{\hat{j}=1}^{N_t} \sum_{q=1}^{M} \sum_{\hat{q}=1}^{M} \bar{\delta}(q';\hat{q}') \bar{P}_{ns}(j,q';\hat{j},\hat{q}'), \tag{29}$$

respectively. Here,

$$q' = (j-1) \times 2^m + q = (j-1)M + q, \tag{30}$$

$$\hat{q}' = (\hat{j}-1) \times 2^m + \hat{q} = (\hat{j}-1)M + \hat{q}, \tag{31}$$

$$\bar{P}_{nb}(j,q';\hat{j},\hat{q}') = \frac{N_b(j,q';\hat{j},\hat{q}')}{\log_2(N_t M)} \bar{P}_{ns}(j,q';\hat{j},\hat{q}'), \tag{32}$$

in which $j \in \{1,2,\ldots,N_t\}$ and $q \in \{1,2,\ldots,M\}$ are the inter values determined by $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$, respectively. $N_b(j,q';\hat{j},\hat{q}')$ is the Hamming distance between two transmit bit sequences corresponding to $\mathbf{x}_{jq'}$ and $\mathbf{x}_{\hat{j}\hat{q}'}$, respectively. Notice that $\mathbf{x}_{jq'}$ and $\mathbf{x}_{\hat{j}\hat{q}'}$ have the definition of (20). $\bar{P}_{ns}(j,q';\hat{j},\hat{q}')$ denotes the ASEP of pair of SM symbols denoted by $\mathbf{x}_{jq'}$ and $\mathbf{x}_{\hat{j}\hat{q}'}$, which is calculated as

$$\bar{P}_{ns}(j,q';\hat{j},\hat{q}') = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{e^{-\tilde{\mathbf{m}}_z^H \left[\mathbf{I}\sin^2\theta + \tilde{\mathbf{R}}_z\right]^{-1} \tilde{\mathbf{m}}_z}}{\det\left(\mathbf{I} + \frac{\tilde{\mathbf{R}}_z}{\sin^2\theta}\right)} d\theta \tag{33}$$

with

$$\tilde{\mathbf{m}}_z = \sqrt{\frac{\gamma K}{4(K+1)}} (x_{q'} - x_{\hat{q}'}) \mathbf{e}_{N_r} \tag{34}$$

$$\tilde{\mathbf{R}}_z = \frac{\gamma}{4(K+1)} \left[|x_{q'}|^2 + |x_{\hat{q}'}|^2 - 2\Re\{\varphi_{j\hat{j}}^t x_{q'} x_{\hat{q}'}^*\} \right] \mathbf{R}_r. \tag{35}$$

C. Unequal error protection (UEP) performance analysis

According to (11) and (19), the traditional and novel SM schemes imply two components' estimation process, one is the estimation of $\mathbf{b}_{1:n}$, the other is the estimation of $\mathbf{b}_{n+1:n+m}$. For the the traditional SM scheme, the former means the transmit antenna index is estimated, and the latter means to estimate the transmitted modulation symbol. While for the novel SM scheme, the former means the transmit antenna index and the partitioned subset are estimated at the same time, and the latter means to estimate the transmitted modulation symbol in the partitioned subset.

Notice that $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$ are mapped to different signal space, which implies that they may have different error performance. In other words, the UEP performance is introduced. In the following, we will study the UEP performance of the two SM schemes, from which we could know the advantages of the proposed SM scheme. Without loss of generality, we mainly focus on the ASEP analysis for simply.

1) UEP performance analysis of the traditional SM scheme: According to (26), we have the two ASEPs for $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$ given by

$$\bar{P}_{s,1} \leq c \sum_{j=1}^{N_t} \sum_{\hat{j}=1}^{N_t} \sum_{q=1}^{M} \sum_{\hat{q}=1}^{M} \bar{\delta}(j;\hat{j}) \bar{P}_s(j,q;\hat{j},\hat{q}) \tag{36}$$

and

$$\bar{P}_{s,2} \leq c\sum_{j=1}^{N_t}\sum_{\hat{j}=1}^{N_t}\sum_{q=1}^{M}\sum_{\hat{q}=1}^{M}\bar{\delta}(q;\hat{q})\bar{P}_s(j,q;\hat{j},\hat{q}), \tag{37}$$

respectively. Here $\bar{\delta}(\cdot)$ has the same definition as (27). And the difference between $\bar{P}_{s,1}$ and $\bar{P}_{s,2}$ is given by

$$\bar{P}_{s,1-s,2} \triangleq \bar{P}_{s,1} - \bar{P}_{s,2} \tag{38}$$

Since $\bar{\delta}(j;\hat{j}) = \bar{\delta}(q;\hat{q}) = 1$ for $j \neq \hat{j}$ and $q \neq \hat{q}$, we have

$$\begin{aligned}\bar{P}_{s,1-s,2} \approx\, & \overbrace{c\sum_{j=1}^{N_t}\sum_{\hat{j}=1,\hat{j}\neq j}^{N_t}\sum_{q=\hat{q}=1}^{M}\bar{P}_s(j,q;\hat{j},\hat{q})}^{P_A} \\ & \underbrace{-\,c\sum_{j=\hat{j}=1}^{N_t}\sum_{q=1}^{M}\sum_{\hat{q}=1,\hat{q}\neq q}^{M}\bar{P}_s(j,q;\hat{j},\hat{q})}_{P_B},\end{aligned} \tag{39}$$

where P_A means to estimate two different transmit antennas with the same transmit modulation symbol, while P_B means to estimate two different modulation signals at the same transmit antenna. In other words, P_A is an equivalent SSK modulation scheme with a specified complex signal transmission [4], while P_B is corresponding to the conventional digital modulation schemes with a single transmit antenna.

For P_A, i.e., $j \neq \hat{j}, q = \hat{q}$, according to (23), (24) and (25) we have

$$\mathbf{m}_{Az} = 0\mathbf{e}_{N_r}, \tag{40}$$

$$\mathbf{R}_{Az} = \frac{\gamma|x_q|^2}{2(K+1)}\left[1 - \Re\{\varphi^t_{j\hat{j}}\}\right]\mathbf{R}_r, \tag{41}$$

$$\bar{P}_s(j,q;\hat{j},\hat{q}) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left[\det\left(\mathbf{I} + \frac{\mathbf{R}_{Az}}{\sin^2\theta}\right)\right]^{-1}d\theta. \tag{42}$$

Similarly, for P_B, i.e., $j = \hat{j}, q \neq \hat{q}$, we have

$$\mathbf{m}_{Bz} = \sqrt{\frac{\gamma K}{4(K+1)}}(x_q - x_{\hat{q}})\mathbf{e}_{N_r}, \tag{43}$$

$$\mathbf{R}_{Bz} = \frac{\gamma|x_q - x_{\hat{q}}|^2}{4(K+1)}\mathbf{R}_r, \tag{44}$$

$$\bar{P}_s(j,q;\hat{j},\hat{q}) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\frac{e^{-\mathbf{m}^H_{Bz}\left[\mathbf{I}\sin^2\theta + \mathbf{R}_{Bz}\right]^{-1}\mathbf{m}_{Bz}}}{\det\left(\mathbf{I} + \frac{\mathbf{R}_{Bz}}{\sin^2\theta}\right)}d\theta. \tag{45}$$

By analyzing (40), (41), (42), (43), (44) and (45), we could define the virtual SNR at each received antenna for P_A and P_B, respectively, which are given by

$$\tilde{\gamma}_A(j \neq \hat{j}, q = \hat{q}) = \frac{\gamma|x_q|^2}{2(K+1)}\left[1 - \Re\{\varphi^t_{j\hat{j}}\}\right], \tag{46}$$

$$\begin{aligned}\tilde{\gamma}_B(j = \hat{j}, q \neq \hat{q}) &= \frac{\gamma K|x_q - x_{\hat{q}}|^2}{4(K+1)} + \frac{\gamma|x_q - x_{\hat{q}}|^2}{4(K+1)} \\ &= \frac{\gamma|x_q - x_{\hat{q}}|^2}{4}.\end{aligned} \tag{47}$$

In fact, the total error performance is mainly determined by minimum values of $\tilde{\gamma}_A(j \neq \hat{j}, q = \hat{q})$ and $\tilde{\gamma}_B(j = \hat{j}, q \neq \hat{q})$, so we further define

$$\tilde{\gamma}_{A,min} = \min_{j,\hat{j},q}\left\{\frac{\gamma|x_q|^2}{2(K+1)}\left[1 - \Re\{\varphi^t_{j\hat{j}}\}\right]\right\}, \tag{48}$$

$$\tilde{\gamma}_{B,min} = \min_{q,\hat{q}}\left\{\frac{\gamma|x_q - x_{\hat{q}}|^2}{4}\right\}. \tag{49}$$

Clearly, P_A is mainly determined by the maximum correlation coefficient and the Rician factor, while P_B is mainly determined by the minimum Euclidean distance between pair of modulation signals. Define

$$g_{A/B} = \frac{\tilde{\gamma}_{A,min}}{\tilde{\gamma}_{B,min}} = \frac{2\min_{j,\hat{j},q}\left\{|x_q|^2\left[1 - \Re\{\varphi^t_{j\hat{j}}\}\right]\right\}}{(K+1)\min_{q,\hat{q}}\left\{|x_q - x_{\hat{q}}|^2\right\}}. \tag{50}$$

We could approximately determine that if $g_{A/B} \geq 1$, P_A is better than P_B.

As a special case for independent fading channels with $K = 0$, we have $g_{A/B} = \frac{\min_{j,\hat{j},q}\{2|x_q|^2\}}{\min_{q,\hat{q}}\{|x_q - x_{\hat{q}}|^2\}} \geq 1$ for the conventional MPSK and MQAM constellations, hence the antenna selection has better performance than the modulation symbols.

But unfortunately, with larger correlation coefficient or larger Rician factor, $g_{A/B}$ will become less, and hereby the performance of the antenna selection will become worse than the performance of the modulation symbol. Thus the total SM performance will be mainly determined by P_A. For the special case with $\Re\{\varphi^t_{j\hat{j}}\} \to 1$, $g_{A/B} \to 0$ and the antenna index will become undetectable, which could also be derived from (42). However, in this case the modulation symbol could be still correctly decoded according to (45) and (49). In order to tackle this issue, we proposed the novel SM scheme to balance the performance of P_A and P_B.

2) UEP performance analysis of the novel SM scheme: According to (29), we have two ASEPs for $\mathbf{b}_{1:n}$ and $\mathbf{b}_{n+1:n+m}$ given by

$$\bar{P}_{ns,1} \leq c\sum_{j=1}^{N_t}\sum_{\hat{j}=1}^{N_t}\sum_{q=1}^{M}\sum_{\hat{q}=1}^{M}\bar{\delta}(j;\hat{j})\bar{P}_{ns}(j,q';\hat{j},\hat{q}') \tag{51}$$

and

$$\bar{P}_{ns,2} \leq c\sum_{j=1}^{N_t}\sum_{\hat{j}=1}^{N_t}\sum_{q=1}^{M}\sum_{\hat{q}=1}^{M}\bar{\delta}(q;\hat{q})\bar{P}_{ns}(j,q';\hat{j},\hat{q}'), \tag{52}$$

respectively. Here $q' = (j-1)M + q$, $\hat{q}' = (\hat{j}-1)M + \hat{q}$ and $\bar{\delta}(\cdot)$ is defined as (27). Define

$$\bar{P}_{ns,1-ns,2} \triangleq \bar{P}_{ns,1} - \bar{P}_{ns,2}. \tag{53}$$

Since $\bar{\delta}(j;\hat{j}) = \bar{\delta}(q;\hat{q}) = 1$ for $j \neq \hat{j}$ and $q \neq \hat{q}$, we have

$$\bar{P}_{ns,1-ns,2} \approx \overbrace{c\sum_{j=1}^{N_t}\sum_{\hat{j}=1,\hat{j}\neq j}^{N_t}\sum_{q=\hat{q}=1}^{M}\bar{P}_{ns}(j,q';\hat{j},\hat{q}')}^{P_{nA}} \underbrace{- c\sum_{j=\hat{j}=1}^{N_t}\sum_{q=1}^{M}\sum_{\hat{q}=1,\hat{q}\neq q}^{M}\bar{P}_{ns}(j,q';\hat{j},\hat{q}')}_{P_{nB}}, \quad (54)$$

where $\bar{P}_{ns}(j,q';\hat{j},\hat{q}')$ has the form of (33). According to (34) and (35), we could define the virtual SNR at each received antenna for P_{nA} and P_{nB}, respectively,

$$\begin{aligned}&\tilde{\gamma}_{nA}(j \neq \hat{j}, q = \hat{q})\\ =&\frac{\gamma K|x_{q'} - x_{\hat{q}'}|^2}{4(K+1)} + \frac{\gamma\left[|x_{q'}|^2 + |x_{\hat{q}'}|^2 - 2\Re\{\varphi^t_{j\hat{j}}x_{q'}x^*_{\hat{q}'}\}\right]}{4(K+1)}\\ =&\frac{\gamma|x_{q'} - x_{\hat{q}'}|^2}{4} + \frac{\gamma\left[\Re\{(1-\varphi^t_{j\hat{j}})x_{q'}x^*_{\hat{q}'}\}\right]}{2(K+1)}, \quad (55)\end{aligned}$$

$$\begin{aligned}&\tilde{\gamma}_{nB}(j = \hat{j}, q \neq \hat{q}) = \frac{\gamma K|x_{q'} - x_{\hat{q}'}|^2}{4(K+1)} + \frac{\gamma|x_{q'} - x_{\hat{q}'}|^2}{4(K+1)}\\ =&\frac{\gamma|x_{q'} - x_{\hat{q}'}|^2}{4}. \quad (56)\end{aligned}$$

And the minimum values of $\tilde{\gamma}_{nA}(j \neq \hat{j}, q = \hat{q})$ and $\tilde{\gamma}_{nB}(j = \hat{j}, q \neq \hat{q})$ are denoted as

$$\tilde{\gamma}_{nA,min} = \min_{j,\hat{j},q}\left\{\tilde{\gamma}_{nA}(j \neq \hat{j}, q = \hat{q})\right\}, \quad (57)$$

$$\tilde{\gamma}_{nB,min} = \min_{j,q,\hat{q}}\left\{\tilde{\gamma}_{nB}(j = \hat{j}, q \neq \hat{q})\right\}. \quad (58)$$

By comparing {(57), (55)} with {(48), (46)}, we know that the minimum value of the component $\frac{\gamma\left[\Re\{(1-\varphi^t_{j\hat{j}})x_{q'}x^*_{\hat{q}'}\}\right]}{2(K+1)}$ in (55) may be less than $\tilde{\gamma}_{A,min}$, however another additional part of $\frac{\gamma|x_{q'}-x_{\hat{q}'}|^2}{4}$ was introduced in (55), which could make up the former SNR loss. Furthermore, when the correlation coefficient or the Rician factor becomes larger, the performance of $\mathbf{b}_{1:n}$ will be mainly determined by the additional part of $\frac{\gamma|x_{q'}-x_{\hat{q}'}|^2}{4}$ for the novel SM scheme, while for the traditional SM scheme, its virtual SNR will become very small, which is really terrible for the correctly decoding. In the following, let's further compare their performance in terms of virtual SNRs in detail.

D. Performance comparison

It is difficult to obtain the performance difference of the two SM schemes exactly since we can not get their ABEP or ASEP in closed form. But we still think this point is indeed very important to clarify our contributions. In this part, we will address this issue via comparing their minimum virtual SNRs for simply.

Firstly, by comparing {(47), (49)} and {(56), (58)}, the different performance behavior of $\mathbf{b}_{n+1:n+m}$ could be easily formulated. Which exactly reflects the performance loss caused by the set-partitioning process in the novel SM scheme, and some detailed values have been provided in Table I. Obviously, there is no performance loss for PSK modulation, whereas the performance loss exists for QAM modulation.

Next, let's further analyze the difference of {(55), (57)} and {(46), (48)}, which reflects the performance of $\mathbf{b}_{1:n}$.

When PSK modulation scheme is employed, according to (55) and (57), the minimum virtual SNR of $\tilde{\gamma}_{nA}$ for the novel SM could be calculated as

$$\tilde{\gamma}^{\text{PSK}}_{nA,min} = \frac{\gamma(1-\cos\frac{2\pi}{N_tM})}{2} + \frac{\gamma(1-\rho_t)\cos\frac{2\pi}{N_tM}}{2(K+1)}, \quad (59)$$

which occurs on the condition of selecting two adjacent symbols in the expanded modulation constellation. Correspondingly, from (48) we could get the minimum virtual SNR of $\tilde{\gamma}_A$ for the traditional SM given by

$$\tilde{\gamma}^{\text{PSK}}_{A,min} = \frac{\gamma(1-\rho_t)}{2(K+1)}, \quad (60)$$

and

$$\tilde{\gamma}^{\text{PSK}}_{nA,min} - \tilde{\gamma}^{\text{PSK}}_{A,min} = \frac{\gamma(K+\rho_t)(1-\cos\frac{2\pi}{N_tM})}{2(K+1)} > 0. \quad (61)$$

Moreover, we have known that there is no performance loss for $\mathbf{b}_{n+1:n+m}$ when PSK is employed, and hereby the SNR gain could be achieved for the novel SM scheme over the traditional one with PSK employed. Through comparing (59) and (60), we can find that larger correlated coefficients or larger Rician factor will lead to more SNR gain.

Similarly, when QAM modulation scheme is employed, according to (55) and (57), we have the minimum virtual SNR of $\tilde{\gamma}_{nA}$ for the novel SM scheme calculated as

$$\tilde{\gamma}^{\text{QAM}}_{nA,min} = \frac{\gamma d^2_{min}(\mathbb{C}_{2^{n+m}})}{2}, \quad (62)$$

which occurs on the condition of selecting two adjacent symbols among the candidates closest to the origin. Obviously, these two symbols satisfy $\Re\{x_q x^*_{q'}\} = 0$. (62) means that the novel SM could provide at least the same minimum Euclidean distance as the QAM constellation with the same spectrum efficiency for single antenna transmission.

On the other hand, the minimum virtual SNR of $\tilde{\gamma}_A$ for the traditional SM is given by

$$\tilde{\gamma}^{\text{QAM}}_{A,min} = \frac{\gamma(1-\rho_t)d^2_{min}(\mathbb{C}_{2^m})}{2(K+1)}. \quad (63)$$

Through comparing (62) and (63) we could know that the SNR gain only exists at the case of $\tilde{\gamma}^{\text{QAM}}_{nA,min} > \tilde{\gamma}^{\text{QAM}}_{A,min}$, i.e., over the correlated fading channels with larger correlated coefficients or larger Rician factor. Otherwise, the novel SM will lead to a performance loss. From this point of view, the novel SM scheme is more suitable for the PSK modulation schemes.

VI. SIMULATION RESULTS

In this section, we show the performance of the novel and traditional SM schemes over different channel conditions by using the Monte Carlo simulation and numerical analysis. In simulated figures NSM and TSM denote the novel and traditional SM schemes, respectively, and N_r accounts for the number of receive antennas. The upper bounds are calculated by numerical integration.

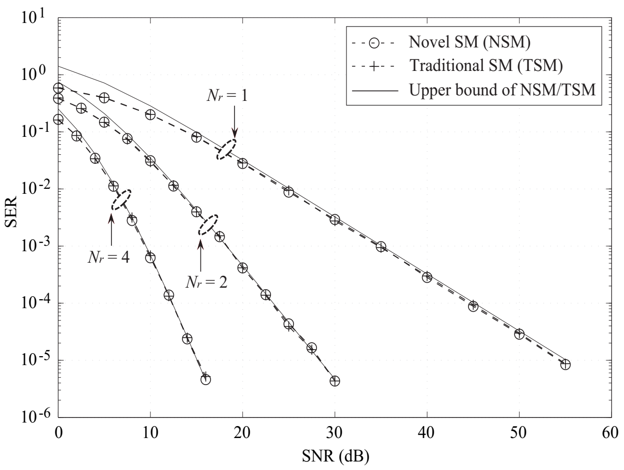

Figure 4. SER performance of two SM schemes over independent Rayleigh fading channels (3 bps/Hz: $N_t = 2$, QPSK for the traditional SM scheme and 8PSK for the novel SM scheme).

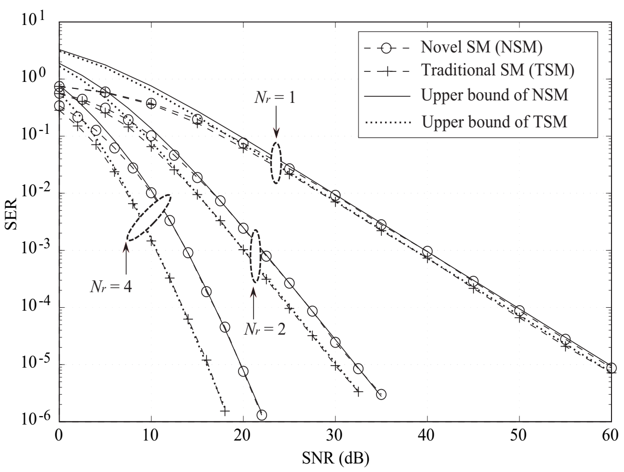

Figure 5. SER performance of two SM schemes over independent Rayleigh fading channels (4 bps/Hz: $N_t = 4$, QPSK for the traditional SM scheme and 16QAM for the novel SM scheme).

With $N_t = 2,\ 4$ as examples, Fig. 4 and Fig. 5 demonstrate the performance of the novel SM scheme compared with the traditional SM scheme over independent Rayleigh fading channels. The total spectrum efficiencies are set to be 3bps/Hz and 4bps/Hz for Fig. 4 and Fig. 5, respectively. From Fig. 4 we could see that almost the same symbol error rate (SER) performance is achieved for the novel and traditional SM schemes. The reason is straightforward. Because over independent Rayleigh fading channels the overall performance is mainly determined by the modulation signal domain, which is worse than the antenna domain. So the performance of the novel SM scheme will be determined by the minimum Euclidean distance between pair of modulation signals in the partitioned subset. From Table I we could see that there is no performance loss by partitioning 8PSK to QPSK, and hereby the same performance is achieved for the novel and traditional SM schemes in Fig. 4. With the same reason, the performance loss is observed in Fig. 5, because there is about 1 dB SNR loss for each receive antenna by partitioning 16QAM to QPSK, as shown in Table I. If there is more receive antennas, there will be more overall SNR loss, as demonstrated in Fig. 5. Hence we conclude that the novel scheme have no advantages over the independent Rayleigh fading channels.

With $N_t = 4$ as examples, Fig. 6 illustrates the SER performance of the two SM schemes over Rician fading channels with respect to different Rician factor. From them we could see the improved SER performance is achieved for the novel SM scheme compared with the traditional one. Although the later is better than the former over the independent Rayleigh fading channels. Via jointly comparing Fig. 5 and Fig. 6, another phenomena is that with the increasing of the Rician factor (K), the performance of the traditional SM becomes worse, while the performance of the novel SM scheme is improved. This is consistent with our theoretical analysis about the UEP performances.

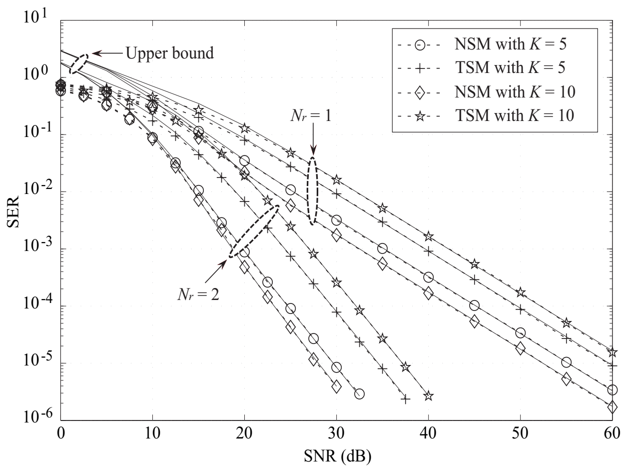

Figure 6. SER performance of two SM schemes over Rician fading channels with Rician factor (K) equal to 5 and 10, respectively (4 bps/Hz: $N_t = 4$, QPSK for the traditional SM scheme and 16QAM for the novel SM scheme).

Fig. 7 shows the SER performance of the two SM schemes over correlated fading channels. Here two different correlated coefficients are considered, i.e., $\rho = \rho_t = \rho_r = 0.5$ and 0.9, respectively. The correlated coefficients are generated by using the exponential correlation model of (17) with ρ denoting the correlation coefficient between adjacent transmit/receive antennas. The obvious result is when the the correlated coefficients become large enough, the performance of the novel SM will become better than the traditional one. And larger correlated coefficients, more performance improvement. This is also consistent with our UEP performance analysis.

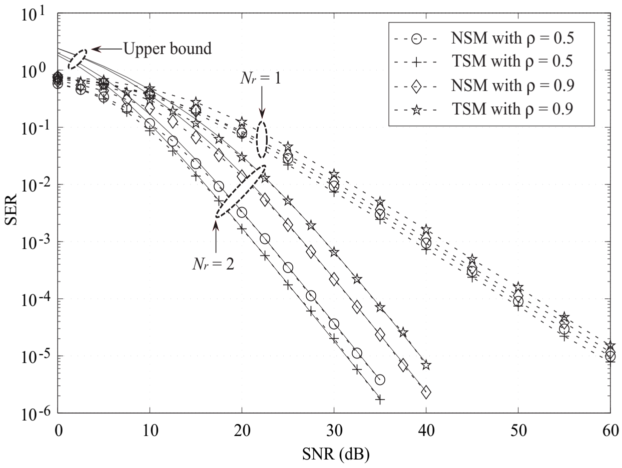

Figure 7. SER performance of two SM schemes over correlated fading channels with correlated factor $\rho = 0.5$ and 0.9, respectively (4 bps/Hz: $N_t = 4$, QPSK for the traditional SM scheme and 16QAM for the novel SM scheme).

Through comparing Fig. 7 and Fig. 6, another phenomena is observed that the channel correlation has a negative impact on the error performance, while the effect of the Rician factor is positive. That is to say, the system performance will be weakened with the increasing of the correlation coefficients, whereas the larger Rician factor will improve the system performance. The reason is also straightforward, because larger Rician factor means more close to the AWGN channel and reduced fading impact. On the contrary, larger correlation means more difficult to distinguish the different transmit antennas while the channel fading still doesn't weaken.

VII. CONCLUSION

In this paper, we have introduced a novel spatial modulation scheme based on the Ungerboeck's set partitioning rule. Different from the traditional SM scheme with only a part of input bits mapping to the modulation signal, the proposed scheme maps all the input bits to the modulation signal constellation. The antenna selection is the same with that of traditional SM scheme. Therefore, the relationship is established between antenna domain and signal domain, with which the balanced error performance between this two domains could be achieved, and thereby the total system performance is improved. The strong correlation or large Rician factor could significantly deteriorate the error performance of antenna domain, but it has only very limited effect on the signal domain. Both performance analysis and simulation results have reinforced this phenomena and the advantages of the novel SM scheme. Note that the decoding complexity of the novel scheme is kept unchanged compared with the conventional SM scheme. Moreover, the set partitioning ideas and the balanced error performance will facilitate the joint design of channel coding and spatial modulation. Detailed study will be included in our further research works.

REFERENCES

[1] S. Ganesan, R. Mesleh, H. Haas, C. W. Ahn, and S. Yun, "On the performance of spatial modulation OFDM," in *Proc. Fortieth Asilomar Conference on Signals, Systems and Computers*, Pacific Grove, CA, USA, October 2006, pp. 1825–1829.

[2] R. Y. Mesleh, H. Haas, S. Sinanovic, C. W. Ahn, and S. Yun, "Spatial modulation," *IEEE Transactions on Vehicular Technology*, vol. 57, no. 4, pp. 2228–2241, July 2008.

[3] J. Jeganathan, A. Ghrayeb, and L. Szczecinski, "Spatial modulation: optimal detection and performance analysis," *IEEE Communications Letters*, vol. 12, no. 8, pp. 545–547, August 2008.

[4] J. Jeganathan, A. Ghrayeb, L. Szczecinski, and A. Ceron, "Space shift keying modulation for MIMO channels," *IEEE Transactions on Wireless Communications*, vol. 8, no. 7, pp. 3692–3703, July 2009.

[5] M. D. Renzo and H. Haas, "A general framework for performance analysis of space shift keying (SSK) modulation for MISO correlated Nakagami-m fading channels," *IEEE Transactions on Communications*, vol. 58, no. 9, pp. 2590–2603, September 2010.

[6] M. D. Renzo and H. Haas, "Space shift keying (SSK) MIMO over correlated Rician fading channels: Performance analysis and a new method for transmit diversity," *IEEE Transactions on Communications*, vol. 59, no. 1, pp. 116–129, January 2011.

[7] M. D. Renzo and H. Haas, "Performance analysis of spatial modulation," in *Proc. International ICST Conference on Communications and Networking in China (CHINACOM)*, August 2010, pp. 1–7.

[8] M. D. Renzo and H. Haas, "Performance comparison of different spatial modulation schemes in correlated fading channels," in *Proc. IEEE International Conference on Communications (ICC)*, May 2010, pp. 1–6.

[9] M. Koca and H. Sari, "A general framework for performance analysis of spatial modulation over correlated fading channels," in *Submitted to IEEE International Conference on Communications (ICC) 2012*, Ottawa, Canada, 2012, pp. 1–6.

[10] R. Mesleh, M. D. Renzo, H. Haas, and P. M. Grant, "Trellis coded spatial modulation," *IEEE Transactions on Wireless Communications*, vol. 9, no. 7, pp. 2349–2361, July 2010.

[11] E. Basar, U. Aygolu, E. Panayirci, and H. V. Poor, "New trellis code design for spatial modulation," *IEEE Transactions on Wireless Communications*, vol. 10, no. 8, pp. 2670–2680, August 2011.

[12] G. Ungerboeck, "Channel coding with multilevel/phase signals," *IEEE Transactions on Information Theory*, vol. 28, no. 1, pp. 55–67, January 1982.

[13] A. Forenza and D. Love and R. Heath Jr., "A low complexity algorithm to simulate the spatial covariance matrix for clustered MIMO channel Models," in *Proc. IEEE Vehicular Technology Conference*, vol. 2, Los Angeles, CA, USA, May 2004, pp. 889–893.

[14] D.-S. Shiu, G. J. Foschini, M. J. Gans, and J. M. Kahn, "Fading correlation and its effect on the capacity of multielement antenna systems," *IEEE Transactions on Communications*, vol. 48, no. 3, pp. 502–513, March 2000.

[15] A. Forenza, D. J. Love, and R. W. H. Jr., "Simplified spatial correlation models for clustered MIMO channels with different array configurations," *IEEE Transactions on Vehicular Technology*, vol. 56, no. 4, pp. 1924–1934, July 2007.

[16] S. L. Loyka, "Channel capacity of MIMO architecture using the exponential correlation matrix," *IEEE Communications Letters*, vol. 5, pp. 369–371, September 2001.

[17] M. D. Renzo and H. Haas, “Bit error probability of SM-MIMO over generalized fading channels,” *IEEE Transactions on Vehicular Technology*, vol. 61, no. 3, pp. 1124–1144, March 2012.
[18] J. G. Proakis, *Digital communications*, 4th ed. McGraw-Hill Higher Education, Dec. 2000.

Peng Zhang is with School of Information Science and Engineering, Shandong University, Jinan, China. He received the M.S. degree from Shandong University in 2002 in electronic engineering, and currently he is a Ph.D. candidate in communication and information systems in Shandong University. His current research interests focus on advanced coded modulation (TCM, MLC, BICM), iterative decoding mechanism, MIMO techniques, cross layer design and wireless network coding. By now he has published over ten papers in some technical journals and some important international conferences.

Dongfeng Yuan (SM’01) received the M.S. degree in Department of Electronic Engineering, Shandong University, China in 1988, and got the Ph.D. degree in Department of Electronic Engineering, Tsinghua University, China in January 2000. Currently he is a full professor and dean in School of Information Science and Engineering, Shandong University, China. From 1993-1994, he was with Electrical and Computer Department at the University of Calgary, Alberta, Canada. He was with Department of Electrical Engineering in the University of Erlangen, Germany from 1998 to 1999; with Department of Electrical Engineering and Computer Science in the University of Michigan, Ann Arbor, USA, from 2001 to 2002; with Department of Electrical Engineering in Munich University of Technology, Germany in 2005, and with Department of Electrical Engineering Heriot-Watt University, UK, 2006. Currently he is the Chairman of IEEE Shandong Section. His current research interests include cognitive radio systems, cooperative (relay) communications, and 4G wireless communications.

Haixia Zhang (S’06, M’08) received the B.E. degree in Department of Communication and Information Engineering, Guilin University of Electronic Technology, China in 2001, and received the M. Eng. degree and Ph.D. degree in communication and information systems in the School of Information Science and Engineering, Shandong University, China, in 2004 and 2008. From 2006 to 2008, she was with the Institute for Circuit and Signal Processing, Munich University of Technology as an Academic Assistant. Currently, she works as a full professor at Shandong University. She has been actively participating many academic events, serving as TPC members, session chairs and giving invited talks for conferences and serving as reviewers for numerous journals. She is the associate editor for International Journal of Communication Systems. Her current research interests include cognitive radio systems, cooperative (relay) communications, cross-layer design of wireless communication networks, Space-time process techniques, precoding/beamforming, and 4G wireless communications and beyond.