Spatially Modulated Orthogonal Space-Time Block Codes with Non-Vanishing Determinants

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***Abstract*—This paper proposes a multiple-input multiple-output (MIMO) transmission scheme for M-ary modulations, called *Spatially Modulated Orthogonal Space-Time Block Coding (SM–OSTBC)*, based on the concept of Spatial Constellation (SC) codewords introduced by *Le at el.* [1]. In the proposed scheme, transmit codeword matrices are generated by multiplying SC matrices with codewords constructed from Orthogonal Space-time Block Codes (O–STBC). The maximum spectral efficiency of the proposed scheme is equal to $(n_T - 2 + \log_2 M)$ bpcu, where n_T is the number of transmit antennas and M is the modulation order. The SC matrices provide a means of carrying information bits together with the O–STBC codewords and allow the SM–OSTBC scheme to achieve second-order transmit diversity by satisfying the non-vanishing determinant property. A systematic method to design the SC codewords for an even number of transmit antennas greater than 3 is presented. A single-stream maximum-likelihood (ML) decoder, which requires a low computational complexity thanks to the structure of the SM–OSTBC codewords and to the orthogonality of the O–STBCs, and a sphere decoder with further reduced signal processing complexity are developed. The bit error rate (BER) performance of the proposed scheme is studied by using both theoretical union bound analysis and computer simulations. Finally, simulation results are presented in order to compare BER performance, energy efficiency and decoding complexity of the proposed scheme with those of several existing MIMO transmission schemes.**

***Index Terms*—MIMO systems, spatial modulation, space-time block coding, space-time block coded spatial modulation, spatial constellation matrices.**

I. INTRODUCTION

MULTIPLE-INPUT Multiple-Output (MIMO) wireless communication systems have been theoretically and practically shown to remarkably improve capacity and reliability compared to conventional single antenna wireless systems [2]-[6]. For instance, the Vertical Bell Laboratories Layered Space-Time (V–BLAST) scheme [4] is capable of achieving a high multiplexing gain by transmitting data streams in parallel from different transmit antennas, yet at the cost of a significant decoding complexity that is required for reducing the impact of Inter-Channel Interference (ICI). Orthogonal Space-Time Block Codes (O–STBCs),[1] on the contrary, were devised to attain full-diversity, i.e., the maximum achievable diversity order [5]–[9]. Unfortunately, rate-one full-diversity O–STBCs exist only for 2 transmit antennas. For more than 2 transmit antennas, the maximum achievable symbol rate of O–STBCs is limited to $\frac{3}{4}$ symbols per channel use, thus resulting in a loss of MIMO channel capacity [6]–[11]. In [11], Hassibi *et al.* proposed the so-called linear dispersion codes (LDCs), which subsume both V–BLAST and O–STBCs, in order to obtain a flexible trade-off between the achievable diversity order and the multiplexing gain. Furthermore, Quasi-Orthogonal STBCs (QO–STBCs) [12]–[14] and various high-rate full-diversity STBCs were proposed in [15]–[22].

Spatial Modulation (SM) was recently proposed by Mesleh *et al.* [23], [24]. In SM, the transmitter activates one out of n_T transmit antenna elements at every symbol time, and a signal modulated using conventional M-point constellations, such as Quadrature Amplitude Modulation (QAM) or Phase-Shift Keying (PSK), is transmitted from the activated antenna [25]. In SM, information bits are conveyed not only by the modulated symbol but also by the index of the activated antenna element. Since just one symbol is transmitted in every symbol time, the problem of ICI, as faced by the V–BLAST, is completely avoided by SM-MIMO systems. Hence, low-complexity Maximum-Likelihood (ML) detection can be implemented at the receiver [26]. A special case of SM, known as Space Shift Keying (SSK) [27], was devised by simply assigning or not energy to a specific transmit antenna according to the information bits, without activating the QAM/PSK-modulated symbol. Although SM and SSK potentially outperform several other MIMO systems [23], [24], [27], they do not achieve any transmit diversity gains, as shown in [28]–[29]. Thus, they have to rely on the deployment of multiple receive antenna elements, in order to combat the effects of fading channels. Furthermore, SM/SSK schemes can offer only a logarithmic increase of the data rate with the number of transmit antennas. This possibly prevents SM to attain a very high spectral efficiency for a practical number of antennas at the transmitter.

In order to overcome the lack of transmit diversity inherent in SM/SSK, a number of MIMO transmission schemes were recently proposed. In [30] and [31], Di Renzo *et al.* propose

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[1]Depending on the context, "STBC(s)" can be interpreted as space-time block coding/codes

a second-order transmit-diversity scheme, which uses time-orthogonal shaping filters at the transmitter to enjoy, without rate reduction, transmit-diversity gains with a single active RF chain. In [32], [33] and [34], the authors extend the scheme by amalgamating SM with OSTBCs in a MIMO architecture that provides a flexible trade-off between achievable diversity and throughput. Low-complexity demodulation schemes are also provided. The proposed SM-aided MIMO architecture subsumes many state-of-the-art MIMO schemes based on space-time coded transmission. As such, it can be readily integrated into current wireless communications standards, such as the Long Term Evolution Advanced (LTE-A) [35]. In [36], Başar *et al.* propose Space-Time Block Coded Spatial Modulation (STBC–SM), which combines SM with the Alamouti STBC (as the core STBC). In STBC–SM, both the Alamouti STBC codewords and the indices of the transmit antennas, from which these codewords are transmitted, are used to transmit information bits, whereby simultaneously achieving second-order transmit diversity and a spectral efficiency of $m = \frac{1}{2}\log_2 c + \log_2 M$ bpcu, where M is the constellation size and c is the largest integer power of 2 no greater than the total number of antenna-index combinations [36]. Moreover, thanks to the orthogonality of the Alamouti STBC, a low-complexity ML decoder[2] is proposed. Nevertheless, optimized rotation phases need to be computed as a function of the number of transmit antennas and of the adopted modulation scheme. In [37], Sugiura *et al.* extend the concept of SM by including both the space and time dimensions, so as to provide a general shift-keying framework known as Space-Time Shift Keying (STSK). The STSK scheme is based on the activation of one out of Q dispersion matrices during each transmission block, rather than than using the transmit-antenna index for encoding the information bits, as SM and SSK do. By optimizing the number, the size of the dispersion matrices and the number of transmit and receive antennas, the STSK scheme is able to provide a flexible diversity and multiplexing tradeoff. Unlike the conventional SM and SSK schemes, which can only attain receive diversity gain, the STSK scheme is capable of obtaining both transmit and receive diversity gains. Besides, since no ICI is imposed by the resultant equivalent system model of the STSK scheme, it is possible to utilize a single-antenna based ML detector at the receiver. In [38], Sugiura *et al.* propose the so-called Generalized Space-Time Shift Keying (G–STSK) architecture, which acts as a unified MIMO framework. Similar to STSK, the G–STSK scheme is also capable of offering a flexible diversity and multiplexing tradeoff. However, unlike the STSK scheme, which activates only one out of Q dispersion matrices during each transmission block, the G–STSK scheme selects P out of Q dispersion matrices from which P PSK/QAM-modulated symbols are transmitted. Due to its high flexibility, the G–STSK framework subsumes many of the existing MIMO arrangements, such as SM/SSK, LDC, STBC and V–BLAST [38]. A comprehensive state-of-the-art survey of transmit-diversity assisted SM/SSK MIMO schemes is available in [35], while a comparison with many state-of-the-art MIMO schemes is available in [34]. Recent experimental activities on the implementation of SM-MIMO schemes are available in [39] and [40].

[2] In this paper the terms "detector" and "decoder" are used interchangeably.

In this paper, motivated by the works of Başar *et al.* [36] and Sugiura *et al.* [38], we propose a MIMO scheme, called *Spatially Modulated Orthogonal Space-Time Block Coding (SM–OSTBC)*, based on the concept of *Spatial Constellation* (SC) codewords originally introduced in [1]. In [1], the method adopted for the construction of the SC codewords is not applicable to an arbitrary number of transmit antennas and the codewords are obtained by using a computer search. Furthermore, no proof of the achievable diversity order is provided. These limitations are removed in the present paper. In particular, the specific contributions of this paper are as follows: i) a systematic way to construct the SC codewords by taking into account the activation of either all or of a subset of transmit antennas is provided; and ii) a theoretical proof of the achievable transmit diversity is given. Compared to the STBC–SM scheme introduced by Başar *et al.* [36], the proposed SM–OSTBC scheme offers the following advantages: 1) it achieves second-order transmit diversity for any modulation scheme and for any even n_T, without the need of optimizing the rotation phases and 2) it provides a higher spectral efficiency. In particular, the achievable spectral efficiency is $(n_T - 2 + \log_2 M)$ bpcu, for the same number of transmit antennas. Likewise, compared to the G–STSK scheme introduced by Sugiura *et al.* in [38], the proposed SM–OSTBC scheme provides the following advantages: 1) the construction of the SC codewords is systematic, without the need of using a computer search and 2) it can be decoded with ML-optimum single-stream decoding complexity, hence reducing the signal processing complexity at the receiver.

In summary, the contributions of the present paper can be summarized as follows:

- A systematic approach to the design of the SC matrices is introduced. It is applicable when the number of transmit-antennas n_T is even and greater than or equal to 4. Furthermore, it can applied when either all or a sub-set of the transmit-antennas are activated. It is shown that in all cases the resulting SM–OSTBC scheme attains a second-order transmit diversity with the Non-Vanishing Determinant (NVD) property. This property can be exploited to realize adaptive modulation schemes, by changing the transmission rate according to the wireless channel quality.
- A single-stream low-complexity ML decoder is proposed, which exploits the orthogonality of the Alamouti STBC. Furthermore, a sphere decoder (SD) with further reduced signal processing complexity is designed.
- Computer simulation results, supported by a mathematical framework based on the union-bound [1], are provided to study the bit error rate (BER) performance of the proposed SM–OSTBC scheme in comparison with those of existing MIMO transmission schemes, such as the Alamouti STBC, the STBC–SM, the STSK, the G–STSK, the Srinath STBC [21], the perfect space-time code (Perfect Code) [20] and the block orthogonal space-time code (BOSTC) [22].

The rest of the paper is organized as follows. Section II introduces the system model. In Section III, the design of the SC codewords is introduced. In Section IV, the single-

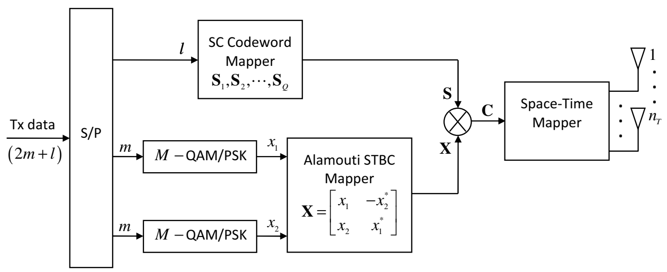

Fig. 1. Block diagram of the SM–OSTBC transmitter.

stream ML-based and SD-based decoders are derived. Section V provides simulation results and a performance comparison with state-of-the-art MIMO schemes. Section VI concludes the paper.

Notation: Column vectors and matrices are denoted by bold lowercase and capital letters, respectively. $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ denote complex conjugation, transposition, and Hermitian transposition operators, respectively. $\Re\{a\}$ and $\Im\{a\}$ denote the real and imaginary parts of the complex number a, respectively. $\binom{n}{k}$ and $\lfloor a \rfloor$ are used for denoting the binomial coefficient and the largest integer less than or equal to a, respectively. $\|\cdot\|$, $\mathrm{tr}(\cdot)$ and $\det(\cdot)$ denote the Frobenius norm, the trace and the determinant of a matrix, respectively. $E\{x\}$ denotes the ensemble average of the random variable x.

II. System Model

A. System Model

Figure 1 shows the transmitter architecture of the proposed SM–OSTBC scheme. The encoding can be summarized as follows. Every two consecutive symbol periods, $(2m+l)$ bits enter the SM–OSTBC transmitter. The first l bits are fed into the SC Codeword Mapper in order to select 1 out of Q SC codewords, ($\mathbf{S}_1, \ldots, \mathbf{S}_Q$, $Q = 2^l$), as the transmitted SC codeword $\mathbf{S}$. The remaining $2m$ bits ($m = \log_2 M$) are modulated by the M–QAM/PSK modulators, which provide the pair of symbols (x_1, x_2), which are then fed into the Alamouti STBC Mapper in order to generate the Alamouti STBC codeword $\mathbf{X}$. The SM–OSTBC codeword $\mathbf{C}$ is obtained by multiplying $\mathbf{S}$ with $\mathbf{X}$, i.e., $\mathbf{C} = \mathbf{SX}$, and subsequently transmitted via the n_T or a subset of the n_T antennas available at the transmitter.

By assuming n_R antennas at the receiver and a quasi-static Rayleigh fading channel, the received $n_R \times 2$ signal matrix $\mathbf{Y}$ can be formulated as:

$$\begin{aligned} \mathbf{Y} &= \sqrt{\gamma}\mathbf{HC} + \mathbf{N} \\ &= \sqrt{\gamma}\mathbf{HSX} + \mathbf{N} \end{aligned} \tag{1}$$

where $\mathbf{H}$ and $\mathbf{N}$ denote the $n_R \times n_T$ channel matrix and the $n_R \times 2$ noise matrix, respectively. The entries of $\mathbf{H}$ and $\mathbf{N}$ are assumed to be independent and identically distributed (i.i.d.) complex Gaussian random variables with zero mean and unit variance. Besides, $\mathbf{H}$ is assumed to remain constant within a codeword of 2 symbol periods and to change independently from one codeword to another. $\mathbf{C}$, $\mathbf{X}$ and $\mathbf{S}$ denote a $n_T \times 2$ SM–OSTBC codeword, a 2×2 Alamouti STBC codeword and a $n_T \times 2$ SC codeword, respetively. The transmit codeword $\mathbf{C}$ is normalized such that the ensemble average of the trace of $\mathbf{C}^H\mathbf{C}$ is equal to 2, i.e., $E\{\mathrm{tr}(\mathbf{C}^H\mathbf{C})\} = 2$. γ is the average SNR at each receive antenna.

It is worth mentioning that in the proposed scheme the Alamouti STBC can be replaced by other STBCs, such as the rate–$\frac{3}{4}$ OSTBCs for 3 and 4 transmit antennas presented in [6] or in [8] and [9], in order to achieve a higher diversity order. For example, the use of the rate-$\frac{3}{4}$ OSTBC for 3 transmit antennas introduced in [8] and [9] was presented in [41]. The SM-STBC scheme in [41] attains third-order transmit diversity with an increase of 1 bit per channel use (bpcu) in spectral efficiency compared to the original rate–$\frac{3}{4}$ OSTBC. Nevertheless, the design of SM–OSTBC schemes based on high-diversity order OSTBCs may lead to the following design challenges:

- Computer search is probably the only way to construct SC codewords for an arbitrary number of transmit antennas. In fact, it is difficult to systematically design SC codewords when the number of columns in an SC code matrix increases.
- As for the the example in [41], increasing the spectral efficiency of 1 bpcu requires the number of SC codewords to be increased by a factor of 2^T if an OSTBC with code length T symbol periods is used. This leads to a non-negligible increase of the detection complexity, a high computational load for searching the SC codewords, as well as a reduction of the coding gain, especially for $T \geq 4$.

The above disadvantages probably make the combination of the proposed scheme with OSTBCs having a code length greater than or equal to 4 less attractive.

B. Differences Between the Proposed SM-OSTBC and the STBC-SM Schemes

The STBC-SM scheme with 4 transmit antennas introduced in [36] can be formulated by using the Alamouti STBC codeword and the following four 4×2 SC codewords:

$$\mathbf{S}_1 = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}^T, \quad \mathbf{S}_2 = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}^T$$
$$\mathbf{S}_3 = \begin{bmatrix} 0 & e^{j\theta} & 0 & 0 \\ 0 & 0 & e^{j\theta} & 0 \end{bmatrix}^T, \quad \mathbf{S}_4 = \begin{bmatrix} e^{j\theta} & 0 & 0 & 0 \\ 0 & 0 & 0 & e^{j\theta} \end{bmatrix}^T$$

where $j = \sqrt{-1}$ is the imaginary unit. Similar SC codewords can be written for an arbitrary number of transmit antennas, but the details are omitted due to space limitations.

In order to achieve second-order diversity, the rotation phase θ of the STBC–SM scheme needs to be optimized as a function of n_T and of the constellation size M. Furthermore, the number of STBC–SM codewords constructed for a given n_T is small. More specifically, the spectral efficiency provided by the STBC–SM scheme is limited to $\frac{1}{2}\log_2 c + \log_2 M$ bpcu, where c is the number of information–bearing STBC–SM codewords. If $n_T = 8$, for example, the additional spectral efficiency (i.e., $\frac{1}{2}\log_2 c$) compared to single-antenna transmission is only 2 bpcu. The spectral efficiency of STBC-SM can be increased by increasing the number of SC codewords. This, however, makes the optimization of the phases θ more complicated and reduces the coding gain of the STBC–SM scheme.

C. Differences Between The Proposed SM-OSTBC and the G-STSK Schemes

Using a linear dispersion representation, the Alamouti STBC codeword $\mathbf{X}$ can be formulated as [11], [42]:

$$\mathbf{X} = \sum_{n=1}^{2} \left(\mathbf{A}_n \tilde{x}_n + j \mathbf{B}_n \breve{x}_n \right) \tag{2}$$

where $\mathbf{A}_n$ and $\mathbf{B}_n$ are 2×2 real dispersion matrices, as well as $\tilde{x}_n = \Re\{x_n\}$ and $\breve{x}_n = \Im\{x_n\}$.

If x_n is drawn from the M-QAM constellation, $\tilde{x}_n$ and $\breve{x}_n$ would correspond to the associated pulse amplitude modulation (PAM) constellation. For example, if a square M-QAM is utilized, $\tilde{x}_n$ and $\breve{x}_n$ belong to a $\sqrt{M}$-PAM constellation. On the other hand, if a rectangular M-QAM is utilized, $\tilde{x}_n$ and $\breve{x}_n$ are from a $\sqrt{2M}$-PAM and a $\sqrt{\frac{M}{2}}$-PAM constellation, respectively. In this case, (2) can be rewritten as follows:

$$\mathbf{X} = \sum_{i=1}^{4} \bar{\mathbf{A}}_i \bar{x}_i \tag{3}$$

where $(\bar{\mathbf{A}}_i, \bar{x}_i) = (\mathbf{A}_n, \tilde{x}_n)$ for $i = 1, 3$ and $n = 1, 2$, while $(\bar{\mathbf{A}}_i, \bar{x}_i) = (j\mathbf{B}_n, \breve{x}_n)$ for $i = 2, 4$ and $n = 1, 2$.

For a given SC codeword $\mathbf{S}_q (q = 1, 2, \ldots, Q)$, the SM–OSTBC codeword can be written as follows:

$$\mathbf{C} = \mathbf{S}_q \mathbf{X} = \sum_{i=1}^{4} \breve{\mathbf{A}}_i^q \bar{x}_i \tag{4}$$

where $\breve{\mathbf{A}}_i^q = \mathbf{S}_q \bar{\mathbf{A}}_i (i = 1, \ldots, 4)$, are equivalent dispersion matrices.

Since the dispersion matrices of a G–STSK code matrix are, in general, complex-valued, equation (4) implies that a SM–OSTBC codeword could be represented in the form of a G–STSK codeword[3] [38] with PAM-modulated symbols $\bar{x}_i (i = 1, \ldots, 4)$. As a consequence, the hard-decision ML decoder of [38] can be used for decoding the proposed SM–OSTBC scheme. However, the decoder of [38] is a multi-stream ML decoder, which jointly decodes P QAM/PSK symbols encoded in a G–STSK codeword. Therefore, its computational complexity increases exponentially with M. Furthermore, the dispersion matrices of the G–STSK scheme are not constructed systematically. Rather, they are obtained with the aid of a computer search, usually requiring millions of dispersion matrices to be tested in order to identify the best set of dispersion matrices based on some design criteria, such as either the rank and determinant criterion or the Discrete-input Continuous-output Memoryless Channel (DCMC) capacity maximization criterion [38]. As a result, the computation complexity for performing the optimization is non-negligible, especially for large sets of dispersion matrices and for large constellation sizes. Therefore, to the best of authors' knowledge, currently available G–STSK schemes are for MIMO setups with up to 4 transmit antennas and for up to 16-QAM modulation.

[3]SM–OSTBC codewords with $x_n (n = 1, 2)$ drawn from a M–PSK constellation with $M \geq 8$ can not be expressed in the form of G–STSK codewords. Moreover, although the total number of equivalent dispersion matrices of the SM–OSTBC scheme is equal to $4Q$, only 4 equivalent dispersion matrices (out of $4Q$) are activated at a time. As a consequence, the number of available combinations of 4 equivalent dispersion matrices of the proposed SM–OSTBC scheme is equal to Q, rather than equal to $f(4Q, 4)$, as for the G–STSK framework, where $f(n,k) = 2^{\kappa} \leq \binom{n}{k} < 2^{\kappa+1}$ and κ is a positive integer.

III. SC Codeword Design

A. Construction of SC Codewords for n_T Simultaneously Activated Transmit Antennas

Let us consider the following $n_T \times 2$ generator matrix:

$$\mathbf{G}(\mathbf{s}) = \frac{1}{\Gamma} \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \\ \vdots & \vdots \\ s_{n_T-1} & s_{n_T} \\ -s_{n_T}^* & s_{n_T-1}^* \end{bmatrix} \tag{5}$$

where $\mathbf{s} = [s_1 \; s_2 \; \ldots s_{n_T}]$ is a $1 \times n_T$ vector with complex-valued entries s_n , $n_T = 2\nu$ for some positive integer $\nu \geq 2$ and $\Gamma = \|\mathbf{s}\| = \sqrt{\sum_{n=1}^{n_T} |s_n|^2}$ is the magnitude of $\mathbf{s}$, which is used to normalize the transmit power.

The proposed general and systematic procedure for designing the SC codewords for n_T activated transmit antennas can be summarized as follows:

1) Set the first two entries of the $1 \times n_T$ vector $\mathbf{s}_q$ equal to 1, i.e., $s_{q,1} = s_{q,2} = 1$.
2) Randomly select the remaining $(n_T - 2)$ entries of $\mathbf{s}_q$ from the set $\{\pm 1, \pm j\}$.
3) Generate the corresponding SC codewords as $\mathbf{S}_q = \mathbf{G}(\mathbf{s}_q)$ for $q = 1, 2, \ldots, Q$.

As an example, let us consider the MIMO setup with $n_T = 4$. In this case, the following $Q = 4^{n_T - 2} = 16$ SC codewords, or equivalently, 16 vectors $\mathbf{s}_q$, can be constructed:

$$\begin{array}{ll}
\mathbf{s}_1 = \begin{bmatrix} 1 & 1 & 1 & 1 \end{bmatrix} & \mathbf{s}_2 = \begin{bmatrix} 1 & 1 & 1 & j \end{bmatrix} \\
\mathbf{s}_3 = \begin{bmatrix} 1 & 1 & 1 & -1 \end{bmatrix} & \mathbf{s}_4 = \begin{bmatrix} 1 & 1 & 1 & -j \end{bmatrix} \\
\mathbf{s}_5 = \begin{bmatrix} 1 & 1 & j & 1 \end{bmatrix} & \mathbf{s}_6 = \begin{bmatrix} 1 & 1 & j & j \end{bmatrix} \\
\mathbf{s}_7 = \begin{bmatrix} 1 & 1 & j & -1 \end{bmatrix} & \mathbf{s}_8 = \begin{bmatrix} 1 & 1 & j & -j \end{bmatrix} \\
\mathbf{s}_9 = \begin{bmatrix} 1 & 1 & -1 & 1 \end{bmatrix} & \mathbf{s}_{10} = \begin{bmatrix} 1 & 1 & -1 & j \end{bmatrix} \\
\mathbf{s}_{11} = \begin{bmatrix} 1 & 1 & -1 & -1 \end{bmatrix} & \mathbf{s}_{12} = \begin{bmatrix} 1 & 1 & -1 & -j \end{bmatrix} \\
\mathbf{s}_{13} = \begin{bmatrix} 1 & 1 & -j & 1 \end{bmatrix} & \mathbf{s}_{14} = \begin{bmatrix} 1 & 1 & -j & j \end{bmatrix} \\
\mathbf{s}_{15} = \begin{bmatrix} 1 & 1 & -j & -1 \end{bmatrix} & \mathbf{s}_{16} = \begin{bmatrix} 1 & 1 & -j & -j \end{bmatrix}
\end{array}$$

The rationale of assigning "the first two" elements of $\mathbf{s}_q$ to 1 is for guaranteeing that the proposed SM–OSTBC scheme satisfies the NVD property, hence providing second-order transmit diversity. As shown in the Appendix, thanks to this choice, (58) can be obtained if $\mathbf{S} \neq \hat{\mathbf{S}}$ and $\mathbf{X} \neq \hat{\mathbf{X}}$, which shows that the codeword distance matrix $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ has rank equal to 2. Without this assignment, some matrices $\mathbf{S} \neq \hat{\mathbf{S}}$ and $\mathbf{X} \neq \hat{\mathbf{X}}$ may exist such that $\mathbf{C} = \hat{\mathbf{C}}$, which would lead to a rank-one codeword distance matrix $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$. As a result, the second-order transmit diversity of the SM–OSTBC scheme will not be guaranteed in this case.

It is worth mentioning that a generator matrix different from (5) may be chosen. However, it should satisfy two main conditions: 1) to be consistent with the number of transmit antennas and the dimensions of the core STBC matrix (i.e., the code matrix $\mathbf{X}$) and 2) to guarantee the desired diversity order and/or data rate of the resulting SM–OSTBC.

According to the design procedure, a total of $Q = 4^{n_T-2}$ SC codewords can be obtained for a given n_T. Therefore, one SC codeword is able to carry $l = \log_2 4^{n_T-2} = 2(n_T - 2)$ information bits. In addition, an Alamouti STBC codeword corresponds to $2m = 2\log_2 M$ information bits. Both the Alamouti STBC codeword and SC codeword are transmitted within 2 symbol periods. Consequently, the spectral efficiency of the SM–OSTBC scheme is equal to $\frac{1}{2}(l + 2m) = (n_T - 2 + \log_2 M)$ bpcu. Compared to SM/SSK, the rate of SM–OSTBC increases linearly instead of logaritmically with n_T. Furthermore, the additional spectral efficiency offered by the SM–OSTBC scheme is substantially higher than that of the STBC–SM scheme, especially when n_T is large. If $n_T = 8$, for example, the additional spectral efficiency of the SM–OSTBC scheme is equal to 6 bpcu, while that of the STBC–SM scheme is equal to 2 bpcu.

For notational convenience, a SM–OSTBC scheme with n_T transmit antennas, n_R receive antennas and n_A simultaneously activated transmit antennas is denoted by $C(n_T, n_R, n_A)$, in what follows. The achievable diversity of a $C(n_T, n_R, n_T)$ SM–OSTBC scheme is summarized in Proposition 1.

Proposition 1: Let a $C(n_T, n_R, n_T)$ SM–OSTBC scheme. It achieves a transmit diversity order equal to 2 and a minimum determinant equal to:

$$\delta_{min} = \begin{cases} \frac{64}{\Gamma^4} & \text{for } M\text{-QAM} \\ \frac{16}{\Gamma^4} & \text{for BPSK and QPSK} \\ 16\sin^4\frac{\pi}{M} & \text{for } M\text{-PSK with } M \geq 8 \end{cases} \quad (6)$$

Proof: See the Appendix.

Equation (6) highlights that the proposed SM–OSTBC scheme satisfies the NVD property.

B. Construction of SC Codewords for n_A Simultaneously Activated Transmit Antennas

Let us assume that a subset of n_A out of n_T transmit antennas are allowed to be activated simultaneously. This is useful in order to reduce the complexity of the transmitter as well as its total power dissipation, since they are both related to the number of RF chains (power amplifiers) and, thus, to the number of active transmit antennas [35], [43], [44].

Based on the structure of the generator matrix in (5), n_A is an even number. The SC codewords can be obtained by setting either $(n_A - 1)$ or n_A entries of $\mathbf{s}$ in (5) equal to non–zero values and the remaining entries equal to zero. In what follows, the SM–OSTBC schemes obtained by setting $(n_A - 1)$ and n_A non-zero entries are denoted by $C(n_T, n_R, n_A)$ and $C_1(n_T, n_R, n_A)$, respectively.

Let n_K be the number of groups of $\mathbf{s}$, each of which is generated by fixing two consecutive entries of $\mathbf{s}$ equal to 1, i.e., $(s_i, s_{i+1}) = (1, 1)$, for $i = 1, 3, 5, \ldots, 2^{n_K} - 1$. The total number of available antenna combinations is:

$$n_C = n_K 2^{\lfloor \log_2 n_L \rfloor} \quad (7)$$

where $n_L = \binom{n_T - 2^{n_K}}{\tilde{n}_A - 2}$, $\tilde{n}_A = n_A$ for $C(n_T, n_R, n_A)$, and $\tilde{n}_A = (n_A - 1)$ for $C_1(n_T, n_R, n_A)$.

The procedure for constructing the SC matrices is as follows. The design objective is to find n_K for $1 \leq n_K \leq \lfloor \log_2(n_T - \tilde{n}_A + 2) \rfloor$, such that n_C is maximized. Once n_K and n_C are determined, $(\tilde{n}_A - 2)$ entries out of the last $(n_T - 2^{n_K})$ entries of $\mathbf{s}$ are randomly assigned equal to $\{\pm 1, \pm j\}$. Therefore, the total number of SC codewords is equal to $4^{(\tilde{n}_A - 2)} n_C$. Among these available $4^{(\tilde{n}_A - 2)} n_C$ SC codewords, only $Q = 2^{\lfloor \log_2(4^{(\tilde{n}_A - 2)} n_C) \rfloor}$ are kept as useful SC codewords for data transmission.

In summary, the step-by-step construction procedure of $C(n_T, n_R, n_A)$ and $C_1(n_T, n_R, n_A)$ SM–OSTBC schemes can be formulated as follows:

1. Given n_T and n_A, find $\tilde{n}_A$ and n_K that maximize n_C in (7)
2. Compute $Q = 2^{\lfloor \log_2(4^{(\tilde{n}_A - 2)} n_C) \rfloor}$ and set $k = 1$.
3. Let $(s_{q,k}, s_{q,k+1})$ of $\mathbf{s}_q$ be fixed to 1 for $q = (k-1)\frac{Q}{n_K} + 1, (k-1)\frac{Q}{n_K} + 2, \ldots, k\frac{Q}{n_K}$.
4. Let $(\tilde{n}_A - 2)$ out of the last $(n_T - 2^{n_K})$ entries of $\mathbf{s}_q$ be randomly selected from the set $\{\pm 1, \pm j\}$, for $q = (k-1)\frac{Q}{n_K} + 1, (k-1)\frac{Q}{n_K} + 2, \ldots, k\frac{Q}{n_K}$.
5. Let the remaining $(n_T - \tilde{n}_A)$ entries of $\mathbf{s}_q$, for $q = (k-1)\frac{Q}{n_K} + 1, (k-1)\frac{Q}{n_K} + 2, \ldots, k\frac{Q}{n_K}$, be set equal to zero. Set $k = k + 2$.
6. If $k \leq 2^{n_K}$, construct other matrices $\mathbf{s}_q$ by applying again steps 3, 4 and 5.
7. Generate the corresponding SC codewords as $\mathbf{S}_q = \mathbf{G}(\mathbf{s}_q)$ for $q = 1, 2, \ldots, Q$.

As an example, let us consider $n_T = 8$ and $n_A = 4$. Also, let us assume that we are interested in constructing an $C_1(n_T, n_R, n_A)$ SM–OSTBC scheme. We have $\tilde{n}_A = n_A - 1 = 3$. By inserting $\tilde{n}_A = 3$ in (7), we have $n_K = 2$, which provides $n_C = 8$. Therefore, a $C_1(n_T, n_R, n_A)$ SM–OSTBC scheme having $Q = 2^{\lfloor \log_2(4n_C) \rfloor} = 32$ SC codewords can be constructed. The first 4 vectors are as follows:

$$\begin{aligned} \mathbf{s}_1 &= \begin{bmatrix} 1 & 1 & 0 & 0 & 1 & 0 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_2 &= \begin{bmatrix} 1 & 1 & 0 & 0 & j & 0 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_3 &= \begin{bmatrix} 1 & 1 & 0 & 0 & -1 & 0 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_4 &= \begin{bmatrix} 1 & 1 & 0 & 0 & -j & 0 & 0 & 0 \end{bmatrix}. \end{aligned}$$

For ease of notation, the 4 vectors can be re-written as:

$$\mathbf{s}_{1:4} = \begin{bmatrix} 1 & 1 & 0 & 0 & s_5 & 0 & 0 & 0 \end{bmatrix}$$

where $s_5 = \{1, j, -1, -j\}$. Using the same notation, the remaining 28 vectors $\mathbf{s}_q$ for $q = 5, 6, \ldots, 32$, can be formulated as follows:

$$\begin{aligned} \mathbf{s}_{5:8} &= \begin{bmatrix} 1 & 1 & 0 & 0 & 0 & s_6 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_{9:12} &= \begin{bmatrix} 1 & 1 & 0 & 0 & 0 & 0 & s_7 & 0 \end{bmatrix} \\ \mathbf{s}_{13:16} &= \begin{bmatrix} 1 & 1 & 0 & 0 & 0 & 0 & 0 & s_8 \end{bmatrix} \\ \mathbf{s}_{17:20} &= \begin{bmatrix} 0 & 0 & 1 & 1 & s_5 & 0 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_{21:24} &= \begin{bmatrix} 0 & 0 & 1 & 1 & 0 & s_6 & 0 & 0 \end{bmatrix} \\ \mathbf{s}_{25:28} &= \begin{bmatrix} 0 & 0 & 1 & 1 & 0 & 0 & s_7 & 0 \end{bmatrix} \\ \mathbf{s}_{29:32} &= \begin{bmatrix} 0 & 0 & 1 & 1 & 0 & 0 & 0 & s_8 \end{bmatrix} \end{aligned}$$

where $s_6, s_7, s_8 = \{1, j, -1, -j\}$.

A similar procedure can be used to construct a $C(n_T, n_R, n_A)$ SM–OSTBC scheme with $\tilde{n}_A = n_A = 4$.

In this case, we would have $(n_K, n_C, Q) = (1, 15, 128)$. However, due to space limitations, $\mathbf{s}_q$ are not listed here.

The achievable diversity order of $C(n_T, n_R, n_A)$ and $C_1(n_T, n_R, n_A)$ SM–OSTBC schemes are summarized in Proposition 2.

Proposition 2: Let either a $C(n_T, n_R, n_A)$ or a $C_1(n_T, n_R, n_A)$ SM–OSTBC scheme. They both achieve a transmit diversity order equal to 2 and a minimum determinant equal to:

$$\delta_{min,\tilde{n}_A} = \begin{cases} \frac{64}{\tilde{n}_A^2} & \text{for } M\text{–QAM} \\ \frac{16}{\tilde{n}_A^2} & \text{for BPSK and QPSK} \\ 16\sin^4\frac{\pi}{M} & \text{for } M\text{–PSK with } M \geq 8 \end{cases} \tag{8}$$

Proof: The proof follows the same line of thought as the proof of Proposition 1, hence it is omitted. The main difference consists in the normalization factor $\Gamma = \|\mathbf{s}\|$. When only n_A transmit antennas are activated simultaneously, $\Gamma = \sqrt{\sum_{n=1}^{\tilde{n}_A} |s_n|^2} = \sqrt{\tilde{n}_A}$. Therefore, the minimum determinant is equal to $\delta_{min,\tilde{n}_A}$ in (8).

Compared to the STBC–SM scheme [36], the disadvantage of the proposed scheme is that it requires a higher number of RF chains. For example, the SM–OSTBC scheme requires $n_A \geq 4$, while the STBC–SM scheme [36] works with $n_A = 2$. However, this increase of the number of RF chains is rewarded by an increase of the spectral efficiency. If $n_T = 8$ and $n_A = 4$, for example, the proposed SM–OSTBC scheme provides $Q = 128$ SC codewords, thus offering an additional spectral efficiency of $\frac{\log_2 Q}{2} = 3.5$ bpcu. On the other hand, the additional spectral efficiency of the STBC–SM scheme is only 2 bpcu. Likewise, the proposed SM–OSTBC scheme provides $Q = 2048$ SC codewords if $n_T = 16$ and $n_A = 4$. Thus, the additional spectral efficiency is equal to $\frac{\log_2 Q}{2} = 5.5$ bpcu, while the STBC–SM scheme is capable of improving the spectral efficiency only by 3 bpcu.

IV. Detection of the SM–OSTBC Scheme

If QAM modulation is used, the proposed SM–OSTBC scheme can be formulated in terms of a G–STSK scheme. Thus, demodulation can be performed with the aid of the multi-stream ML decoder of [38]. As a result, it requires an exhaustive search over all the combinations of dispersion matrices and modulation symbols, which result in a high demodulation complexity. To avoid this issue, in this section two low-complexity single-stream demodulators are proposed for the SM–OSTBC scheme. The first one is ML-optimum and it is referred to as SO-ML. The second one is based on the SD principle [47], [48] and it is referred to as SO-SD. The low-complexity of both demodulators originate from the structure of the SM–OSTBC codewords as well as from the orthogonality of the Alamouti STBC.

A. Proposed SO–ML Decoder

Let us assume that perfect channel state information is available at the receiver. Let Ω_S be the search space associated with the SC codewords $\mathbf{S}$. Let Ω_X be the search space associated with the Alamouti STBC codewords $\mathbf{X}$. The, ML-optimum demodulator of the proposed SM–OSTBC scheme can be formulated as follows:

$$(\hat{\mathbf{S}}, \hat{\mathbf{X}}) = \arg \min_{\mathbf{S}\in\Omega_S, \mathbf{X}\in\Omega_X} \|\mathbf{Y} - \sqrt{\gamma}\mathbf{HSX}\|^2 \tag{9}$$

Let a matrix $\mathbf{S}_q \in \Omega_S, q = 1, \ldots, Q$. Then, the corresponding $n_R \times 2$ equivalent matrix $\tilde{\mathbf{H}}_q = \mathbf{HS}_q$ can be obtained. Thus, (1) reduces to the well-known Alamouti scheme:

$$\mathbf{Y} = \sqrt{\gamma}\tilde{\mathbf{H}}_q\mathbf{X} + \mathbf{N} \tag{10}$$

and the ML demodulation rule in (9) reduces to that for $\mathbf{X}$ conditioned upon $\mathbf{S}_q$, as follows:

$$(\hat{\mathbf{X}})_q = \arg \min_{\mathbf{X}\in\Omega_X} \left\|\mathbf{Y} - \sqrt{\gamma}\tilde{\mathbf{H}}_q\mathbf{X}\right\|^2 \tag{11}$$

From (11), the following equality holds:

$$\begin{aligned}\left\|\mathbf{Y} - \sqrt{\gamma}\tilde{\mathbf{H}}_q\mathbf{X}\right\|^2 &= \|\mathbf{Y}\|^2 - 2\sqrt{\gamma}\Re\{\mathrm{tr}(\mathbf{Y}^H\tilde{\mathbf{H}}_q\mathbf{X})\} \\ &\quad + \gamma\left\|\tilde{\mathbf{H}}_q\mathbf{X}\right\|^2\end{aligned} \tag{12}$$

By using the linear dispersion representation of the Alamouti STBC codeword $\mathbf{X}$ in (2), the orthogonality of $\mathbf{X}$, and by neglecting some irrelevant constants, (12) can be re-written as follows:

$$\begin{aligned}\left\|\mathbf{Y} - \sqrt{\gamma}\tilde{\mathbf{H}}_q\mathbf{X}\right\|^2 &= \sum_{n=1}^{2}\Big[\gamma\big\|\tilde{\mathbf{H}}_q\big\|^2(\tilde{x}_n^2 + \breve{x}_n^2) \\ &\quad -2\sqrt{\gamma}\Re\left\{\mathrm{tr}\left(\mathbf{Y}^H\tilde{\mathbf{H}}_q\left(\mathbf{A}_n\tilde{x}_n + j\mathbf{B}_n\breve{x}_n\right)\right)\right\}\Big] \\ &= \gamma\big\|\tilde{\mathbf{H}}_q\big\|^2\sum_{n=1}^{2}\Bigg[(\tilde{x}_n^2 + \breve{x}_n^2) \\ &\quad - 2\frac{\Re\left\{\mathrm{tr}\left(\mathbf{Y}^H\tilde{\mathbf{H}}_q\mathbf{A}_n\right)\right\}}{\sqrt{\gamma}\big\|\tilde{\mathbf{H}}_q\big\|^2}\tilde{x}_n \\ &\quad +2\frac{\Im\left\{\mathrm{tr}\left(\mathbf{Y}^H\tilde{\mathbf{H}}_q\mathbf{B}_n\right)\right\}}{\sqrt{\gamma}\big\|\tilde{\mathbf{H}}_q\big\|^2}\breve{x}_n\Bigg]\end{aligned} \tag{13}$$

Let us define:

$$\bar{x}_{I,n}^q = \frac{\mathrm{tr}\left(\Re\left(\mathbf{Y}^H\mathbf{H}_q\mathbf{A}_n\right)\right)}{\sqrt{\gamma}\|\mathbf{H}_q\|^2} \tag{14}$$

$$\bar{x}_{Q,n}^q = -\frac{\mathrm{tr}\left(\Im\left(\mathbf{Y}^H\mathbf{H}_q\mathbf{B}_n\right)\right)}{\sqrt{\gamma}\|\mathbf{H}_q\|^2} \tag{15}$$

With the aid of some algebraic manipulations, (13) can be simplified as follows:

$$\begin{aligned}\left\|\mathbf{Y}_q - \tilde{\mathbf{H}}_q\mathbf{X}\right\|^2 = \gamma\left\|\tilde{\mathbf{H}}_q\right\|^2\Bigg[\Bigg(&\sum_{n=1}^{2} d_q\left(\tilde{x}_n\right) \\ &+ \sum_{n=1}^{2} d_q\left(\breve{x}_n\right)\Bigg) - R_q\Bigg]\end{aligned} \tag{16}$$

where $d_q(\tilde{x}_n)$ and $d_q(\breve{x}_n)$ are the Euclidean distances of the real and imaginary part of the transmitted symbol x_n:

$$d_q\left(\tilde{x}_n\right) = \left(\tilde{x}_n - \bar{x}_{I,n}^q\right)^2 \tag{17}$$

$$d_q\left(\breve{x}_n\right) = \left(\breve{x}_n - \bar{x}_{Q,n}^q\right)^2 \tag{18}$$

and

$$R_q = \sum_{n=1}^{2}\left[\left(\bar{x}_{I,n}^{q}\right)^2 + \left(\bar{x}_{Q,n}^{q}\right)^2\right] \quad (19)$$

Thus, the ML decoding rule conditioned upon $\mathbf{S}_q$ in (11) can be re-written as follows:

$$(\hat{x}_1^q, \hat{x}_2^q) = \arg\min_{(x_1,x_2)\in\Omega_x}\left(\sum_{n=1}^{2} d_q(\tilde{x}_n) + \sum_{n=1}^{2} d_q(\breve{x}_n)\right) \quad (20)$$

or, equivalently, as follows:

$$\hat{x}_n^q = \arg\min_{x_n\in\Omega_x} d_q(x_n) \quad (21)$$

where:

$$d_q(x_n) = d_q(\tilde{x}_n) + d_q(\breve{x}_n) \quad (22)$$

for $n = 1, 2$ and Ω_x is a M–QAM/PSK constellation.

From $(\hat{x}_1^q, \hat{x}_2^q)$, $q = 1, 2, \ldots, Q$, obtained from (21), the index $\hat{q}$ corresponding to the codeword $\mathbf{S} = \mathbf{S}_q$ is estimated as follows:

$$\hat{q} = \arg\min_{q} D_q \quad (23)$$

where:

$$D_q = \gamma\left\|\tilde{\mathbf{H}}_q\right\|^2\left[\sum_{n=1}^{2} d_q(\hat{x}_n^q) - R_q\right] \quad (24)$$

Finally, the transmitted codeword is estimated as $\hat{\mathbf{S}} = \mathbf{S}_{\hat{q}}, (\hat{x}_1, \hat{x}_2) = (\hat{x}_1^{\hat{q}}, \hat{x}_2^{\hat{q}})$.

As a consequence, the proposed SO–ML decoder can be formulated as follows:

1. For each matrix $\tilde{\mathbf{H}}_q$ and for each signal pair $(x_{1,m}, x_{2,m})$ in the transmit constellation Ω_x, compute the two Euclidean distances from (22) as follows:
   - $d_{1,q}^m = d_q(x_{1,m})$ for $m = 1, \ldots, M$.
   - $d_{2,q}^m = d_q(x_{2,m})$ for $m = 1, \ldots, M$.
2. Find the minimum among the M values of $d_{1,q}^m$ and $\hat{x}_1^q$ corresponding to $d_{1,q}^{\min}$.
3. Find the minimum among the M values of $d_{2,q}^m$ and $\hat{x}_2^q$ corresponding to $d_{2,q}^{\min}$.
4. Calculate $d_q = d_{1,q}^{\min} + d_{2,q}^{\min}$ for $q = 1, \ldots, Q$.
5. Find the index $\hat{q}$ corresponding to the minimum distance $d_q^{\min}$ among the Q values of d_q.
6. Compute the estimated SC matrix and the transmitted symbols as: $\hat{\mathbf{S}} = \mathbf{S}_{\hat{q}}$ and $(\hat{x}_1, \hat{x}_2) = (\hat{x}_1^{\hat{q}}, \hat{x}_2^{\hat{q}})$, respectively.
7. From $\hat{\mathbf{S}}$ and $(\hat{x}_1, \hat{x}_2)$, estimate the $(2m + l)$ information bits.

Similar low-complexity demodulation schemes have been proposed in [34] and [36], but for different SM-MIMO schemes.

B. Proposed SO–SD Decoder

The signal processing complexity of the ML demodulator of Section IV-A can be reduced with the aid of SD [48]. Let us start by re-writing (10) as follows:

$$\mathbf{y} = \bar{\mathbf{H}}_q\mathbf{x} + \mathbf{n} \quad (25)$$

where $\mathbf{x} = \begin{bmatrix} x_1 & x_2 \end{bmatrix}^T$ and $\bar{\mathbf{H}}_q$ is of the form:

$$\bar{\mathbf{H}}_q = \begin{bmatrix} \tilde{h}_{q,11} & \tilde{h}_{q,12} \\ \tilde{h}_{q,12}^* & -\tilde{h}_{q,11}^* \\ \vdots & \vdots \\ \tilde{h}_{q,n_R1} & \tilde{h}_{q,n_R2} \\ \tilde{h}_{q,n_R2}^* & -\tilde{h}_{q,n_R1}^* \end{bmatrix} = \begin{bmatrix} \bar{\mathbf{h}}_{q,1} & \bar{\mathbf{h}}_{q,2} \end{bmatrix} \quad (26)$$

From (25), the ML decoding rule for $\mathbf{x}$ conditioned upon $\mathbf{S}_q$ is:

$$(\hat{\mathbf{x}})_q = \arg\min_{\mathbf{x}\in\Omega_x}\left\|\mathbf{y} - \bar{\mathbf{H}}_q\mathbf{x}\right\|^2 \quad (27)$$

With the aid of the Modified Gram-Schmidt (MGS) algorithm, i.e., the QR decomposition, $\bar{\mathbf{H}}_q$ can be re-written as follows:

$$\bar{\mathbf{H}}_q = \mathbf{Q}_q\mathbf{R}_q \quad (28)$$

where $\mathbf{Q}_q$ is a $2n_R \times 2$ unit-norm orthogonal-column matrix, i.e., $\mathbf{Q}_q^H\mathbf{Q}_q = \mathbf{I}_2$, and $\mathbf{R}_q$ is a 2×2 upper triangular matrix. Since $\bar{\mathbf{h}}_{q,1}^H\bar{\mathbf{h}}_{q,2} = \bar{\mathbf{h}}_{q,2}^H\bar{\mathbf{h}}_{q,1} = 0$ and $\|\bar{\mathbf{h}}_1\| = \|\bar{\mathbf{h}}_2\|$, it follows that $\mathbf{R}_q$ can be written as follows:

$$\mathbf{R}_q = \begin{bmatrix} R_q & 0 \\ 0 & R_q \end{bmatrix} \quad (29)$$

Let pre-multiply both sides of (25) by $\mathbf{Q}_q$. We obtain:

$$\mathbf{v}_q = \mathbf{R}_q\mathbf{x} + \mathbf{w}_q \quad (30)$$

where $\mathbf{v}_q$ and $\mathbf{w}_q$ are the 2×1 received signal vector and the 2×1 noise vector after QR decomposition, respectively.

By using the equality:

$$\|\mathbf{v}_q - \mathbf{R}_q\mathbf{x}\|^2 = \|\mathbf{y} - \mathbf{H}_q\mathbf{x}\|^2 - \mathbf{y}^H\mathbf{y} + \mathbf{v}_q^H\mathbf{v}_q \quad (31)$$

the ML decoding rule in (27) can be re-written as follows:

$$(\hat{\mathbf{x}})_q = \arg\min_{\mathbf{x}\in\Omega_x}\|\mathbf{v}_q - \mathbf{R}_q\mathbf{x}\|^2 + \mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q \quad (32)$$

For a given q, the term $\left(\mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q\right)$ is not a function of $\mathbf{x}$, hence (32) simplifies to:

$$(\hat{\mathbf{x}})_q = \arg\min_{\mathbf{x}\in\Omega_x}\|\mathbf{v}_q - \mathbf{R}_q\mathbf{x}\|^2 \quad (33)$$

SD can be applied to (33) in order to search for $(\hat{\mathbf{x}})_q$. After finding $(\hat{\mathbf{x}})_q$ for $q = 1, 2, \ldots, Q$, the optimal solutions of $\hat{q}$ and $\hat{\mathbf{x}} = (\hat{\mathbf{x}})_{\hat{q}}$ have to be determined. This is possible as follows:

$$\hat{q} = \arg\min_{q\in 1,\ldots,Q}\left\|\mathbf{v}_q - \mathbf{R}_q(\hat{\mathbf{x}})_q\right\|^2 + \mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q \quad (34)$$

In order to further reduce the computational complexity of the SD, the term $\left(\mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q\right)$ in (34) can be taken into account from the beginning. More specifically, let us denote the initial sphere radius by C_0. The SD can search for the vectors $(\hat{\mathbf{x}})_q$ among the vectors $\mathbf{x} \in \Omega_x$ which satisfy the following condition:

$$\|\mathbf{v}_q - \mathbf{R}_q\mathbf{x}\|^2 \le C_0 - \left(\mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q\right) \quad (35)$$

If $C_0 - \left(\mathbf{y}^H\mathbf{y} - \mathbf{v}_q^H\mathbf{v}_q\right) \le 0$, there is no need to apply the SD to (33), since the signal points lie outside the sphere of radius C_0.

Based on the PMLD and QMLD decoders devised for the detection of PSK- and QAM-modulated signals [45], respectively, the SO-SD decoder for the proposed SM–OSTBC scheme can be summarized as follows:

1. Set $q = 1$, $D_{\text{opt}} = C_0$, $\hat{q} = \emptyset$, $\hat{\mathbf{x}} = \emptyset$.
2. Compute $\tilde{\mathbf{H}}_q = \mathbf{H}\mathbf{S}_q$, $\bar{\mathbf{H}}_q$, $\bar{\mathbf{H}}_q = \mathbf{QR}$ and $\mathbf{v} = \mathbf{Q}^H\mathbf{y}$.
3. Compute the sphere radius $R = D_{\text{opt}} - \left(\mathbf{y}^H\mathbf{y} - \mathbf{v}^H\mathbf{v}\right)$ and set $D_{q,\min} = \infty$.
4. If $R > 0$, apply either the QMLD decoder or the PMLD decoder to (33) and search for $(\hat{\mathbf{x}})_q$ by using R as the initial sphere radius. Then, compute $D_{q,\min} = \left\|\mathbf{v}_q - \mathbf{R}_q(\hat{\mathbf{x}})_q\right\|^2 + \left(\mathbf{y}^H\mathbf{y} - \mathbf{v}^H\mathbf{v}\right)$.
5. If $D_{q,\min} < D_{\text{opt}}$, save the new solution $\hat{\mathbf{x}} = (\hat{\mathbf{x}})_q$, $\hat{q} = q$ and assign $D_{\text{opt}} = D_{q,\min}$.
6. Set $q = q+1$. If $q \le Q$, then go to Step 2. If $q > Q$ and $\hat{q} = \emptyset$ (i.e., no solution found), increase the initial sphere radius C_0 and go to Step 1. Otherwise, stop and output the solution $\hat{\mathbf{x}}$ and $\hat{q}$.

C. Computational Complexity Analysis

In this section, we evaluate the signal processing complexity of the proposed SO-ML, SO-SD and exhaustive-search ML decoders (Ex-ML decoder) by computing the number of floating–point operations (flops). A floating-point operation (flop) is assumed to be either a real multiplication, a real addition, a real square root or a real division. The channel is assumed to be static for T symbol periods. Also, the following assumptions are used:

- A complex multiplication requires 4 real multiplications and 2 real additions.
- A complex addition requires 2 real additions.
- Let $\mathbf{A} \in \mathbb{C}^{M\times N}$ and $\mathbf{B} \in \mathbb{C}^{N\times L}$, a matrix multiplication $\mathbf{AB}$ requires MNL complex multiplications and $ML(N-1)$ complex additions.
- A squared Frobenius norm $\|\mathbf{A}\|^2$ requires MN complex multiplications and $(MN-1)$ complex additions.

By taking into account the structure of the generator matrix $\mathbf{G}(\mathbf{s})$ in (5), the number of flops required for computing $\mathbf{H}_q = \mathbf{HS}_q, q = 1, 2, \ldots, Q$, can be shown to be equal to:

$$F_1 = \left(\frac{4}{15}4^{n_T} + 224n_T - \frac{6604}{15}\right)n_R \quad (36)$$

The number of flops per bit of the proposed SO-ML decoder is as follows:

$$C_{\text{SO-ML}} = \frac{\frac{2}{T}[F_1 + Q(16n_R - 2)] + Q(64n_R + 10M + 50)}{l + 2m} \quad (37)$$

On the other hand, the number of flops per bit of the proposed SO-SD decoder, including pre-processing and searching stages, is equal to:

$$C_{\text{SO-SD}} = \frac{\frac{2}{T}[F_1 + (16Q + 6)n_R - 1] + Q(24n_R - 1) + \bar{F}_S}{l + 2m} \quad (38)$$

where $\bar{F}_S$ is the average number of flops of the searching stage.

By contrast, if the ML decoder of the G–STSK scheme in [38] is used, which corresponds to the Ex-ML decoder, the number of flops per bit is equal to:

$$C_{\text{Ex-ML}} = \frac{\frac{2}{T}(F_1 + 28n_RQM^2) + (24n_R - 2)QM^2}{l + 2m} \quad (39)$$

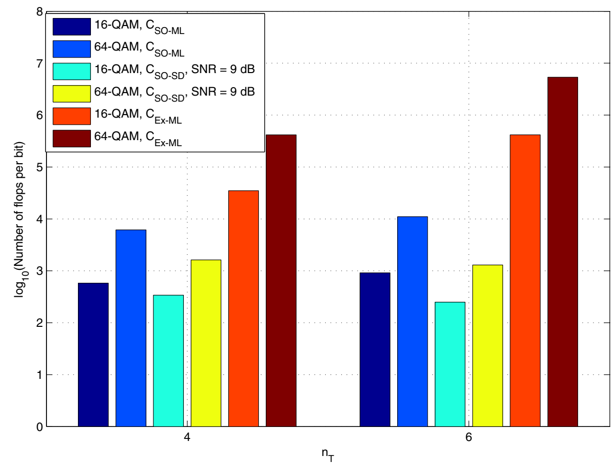

Fig. 2. Computational complexity of SO-ML, SO-SD and Ex-ML decoders for a different number of transmit antennas and constellation sizes. Setup: $C(n_T, 2, n_T)$ SM–OSTBC scheme with $T = 2$ symbol periods.

By comparing (37) and (39), it follows that the complexity of the proposed ML decoder grows linearly with the constellation size M, while the complexity of the Ex-ML decoder increases quadratically with M. Consequently, the proposed SO-ML decoder requires lower complexity compared to its Ex–ML counterpart.

Figure 2 shows the computational complexity of SO-ML, Ex-ML and SO-SD decoders for a Signal-to-Noise-Ratio (SNR) equal to 9 dB. As expected, the SO-SD decoder attains the lowest complexity, especially for large values of M. For example, assuming $n_T = 4$, $n_R = 2$, $Q = 4^2 = 16$ and 16-QAM, the complexities are equal to $C_{\text{SO-ML}} = 578$ (flops/bit), $C_{\text{SO-SD}} = 340$ (flops/bit) and $C_{\text{Ex-ML}} = 34903$ (flops/bit).

It is worth mentioning that the proposed SM–OSTBC scheme can be decoded without the need of using the linearized transceiver signal model of [46], provided that the SO-ML decoder is used. On the other hand, the signal linearization in (25) is necessary if the SO–SD decoder is used.

V. Simulation Results

In this section, Monte Carlo simulations are used to study the BER performance of the proposed $C(n_T, n_R, n_T)$ and $C(n_T, n_R, n_A)$ SM–OSTBC schemes for different antenna arrangements, as well as to compare them against state–of–the–art (SOTA) MIMO systems, such as V–BLAST [4], Alamouti STBC [5], QO–STBC [14], Sezginer STBC [16], SM [24], STBC–SM [36], G–STSK [38], Srinath STBC [21], perfect space–time code (Perfect Code) [20] and block orthogonal space time code (BOSTC) [22] schemes. We assume that the channel state information is perfectly known at the receiver. In addition, ML detection is applied to all systems under consideration.

Figure 3 shows the BER of the proposed $C(4, 2, 4)$, $C_1(4, 4, 4)$ and $C_1(6, 2, 4)$ SM–OSTBCs schemes with 4-PSK, 8-PSK and 8-QAM modulation. Both Monte Carlo simulations and the upper bound derived in [1] are shown. Monte Carlo simulations show that all the proposed demodulators (SO-ML, SO-SD and Ex-ML) provide the same BER. Furthermore, Fig. 3 shows that the upper bound overlaps with the

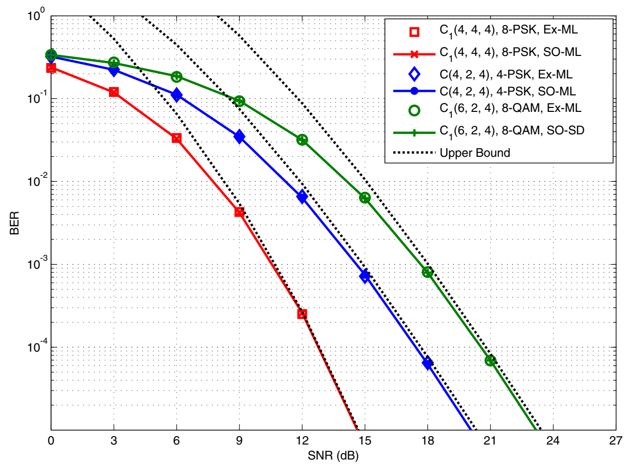

Fig. 3. Theoretical upper bounds and simulation results for the BER of the proposed $C_1(4,4,4)$, $C(4,2,4)$ and $C_1(6,2,4)$ SM–OSTBC schemes.

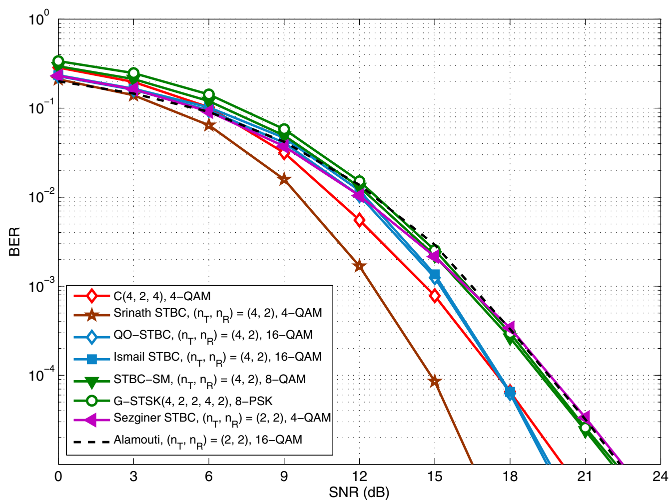

Fig. 4. BER curves of $C(4,2,4)$, G–STSK $(4,2,2,4,2)$, STBC–SM, QO–STBC, Ismail STBC and Srinath STBC for $(n_T, n_R) = (4,2)$ and of the Alamouti STBC and Sezginer STBC for $(n_T, n_R) = (2,2)$. The spectral efficiency is 4 bpcu.

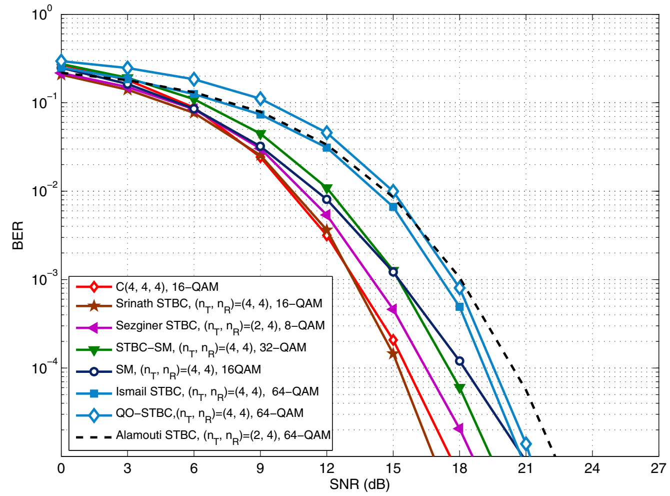

Fig. 5. BER curves of $C(4,4,4)$, STBC–SM, QO–STBC, Ismail STBC, SM and Srinath STBC for $(n_T, n_R) = (4,4)$ and of Alamouti STBC and Sezginer STBC for $(n_T, n_R) = (2,4)$. The spectral efficiency is 6 bpcu.

simulations for high SNR. In all cases, the difference between bounds and simulations is negligible if the BER is less than 10^{-3}. Therefore, the bound could be employed as a useful tool for estimating the BER of the SM–OSTBC scheme for different MIMO configurations at a sufficiently high SNR.

In Fig. 4, the BER of the $C(4,2,4)$ SM–OSTBC, G–STSK $(4,2,2,4,2)$, STBC–SM, QO–STBC, Ismail STBC, rate-2 Srinath STBC, Alamouti STBC and Sezginer STBC schemes are shown, by assuming a spectral efficiency equal to 4 bpcu. The SM–OSTBC, STBC–SM, QO–STBC, G–STSK $(4,2,2,4,2)$, Ismail STBC and Srinath STBC schemes are implemented by assuming 4 transmit antennas, whereas the Alamouti STBC and Sezginer STBC schemes need 2 transmit antennas. All the schemes use 2 receive antennas. An appropriate modulation technique is applied to each MIMO scheme in order to achieve the same spectral efficiency of 4 bpcu, as indicated in Fig. 4. For high SNR, the Alamouti STBC, Sezginer STBC, STBC–SM and G–STSK $(4,2,2,4,2)$ schemes provide a similar BER, since they all provide a diversity order of 4. The QO–STBC and Ismail STBC schemes, on the other hand, provide a diversity order of 8 and have almost identical BER. Compared to these systems, the proposed $C(4,2,4)$ SM–OSTBC scheme is capable of outperforming many MIMO systems, including the Alamouti STBC, Sezginer STBC, STBC–SM, and G–STSK $(4,2,2,4,2)$ schemes. For example, the $C(4,2,4)$ setup provides 2 dB gain over the STBC–SM scheme at $\mathrm{BER} = 10^{-5}$. The reason of this gain is follows. When using 8-QAM, the STBC–SM scheme has $\delta_{\min} = 11.4^4$, which is higher than $\delta_{\min} = 4$ of the $C(4,2,4)$ scheme. The $C(4,2,4)$ scheme, on the other hand, has to use 4-QAM to achieve the same spectral efficiency, hence providing better BER performance. Compared to the QO–STBC and Ismail STBC schemes, the proposed $C(4,2,4)$ scheme has better BER performance for low SNR. However a crossing point exists for $\mathrm{SNR} > 18$ dB. Compared to the Srinath STBC scheme, the proposed $C(4,2,4)$ provides worse performance. This degradation is due to the higher diversity order achieved by the Srinath STBC MIMO system (8 instead of 4). As better discussed in what follows, this loss in performance is counterbalanced by a significant reduction of the detection complexity offered by the proposed SM–OSTBC scheme.

Figure 5 compares the BER of the proposed $C(4,4,4)$ SM–OSTBC scheme with that of the STBC–SM, QO–STBC, Ismail STBC, SM, rate-1.5 Srinath STBC, Alamouti STBC and Sezginer STBC schemes, when the spectral efficiency is 6 bpcu.[5] All the schemes are studied by assuming the setup $(n_T, n_R) = (4,4)$, except for the Alamouti STBC and Sezginer STBC scheme where the setup $(n_T, n_R) = (2,4)$ is used. The rate-1.5 Srinath STBC scheme is obtained from

[4]Using computer search, the optimum rotation angle for the STBC–SM using 8-QAM is equal to $\theta_{opt} = 0.96$ and the corresponding minimum determinant is equal to $\delta_{\min} = 11.4$.

[5]No comparison with the G–STSK scheme is provided because the dispersion matrices of the G–STSK $(4,2,2,4,2)$ scheme is designed for 8–PSK modulation. To the best of the authors' knowledge, there are no available designs of the G–STSK scheme that are capable of providing a spectral efficiency of 6 bpcu or greater.

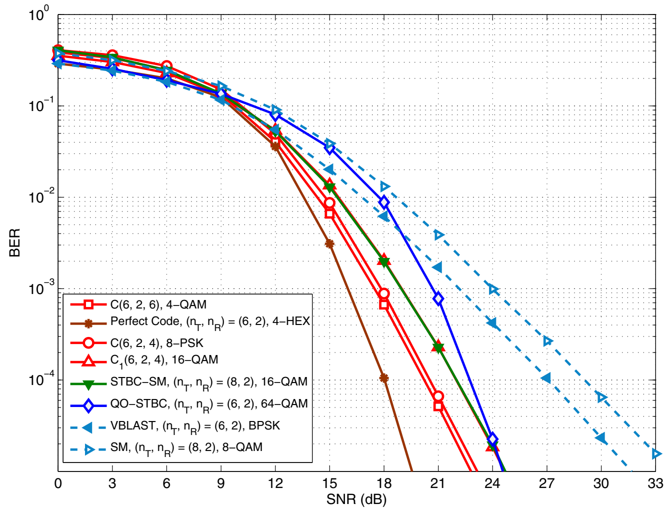

Fig. 6. BER curves of $C(6,2,6)$, $C_1(6,2,4)$, QO–STBC, V–BLAST and Perfect Code for $(n_T, n_R) = (6,2)$ and of STBC–SM and SM for $(n_T, n_R) = (8,2)$. The spectral efficiency is 6 bpcu.

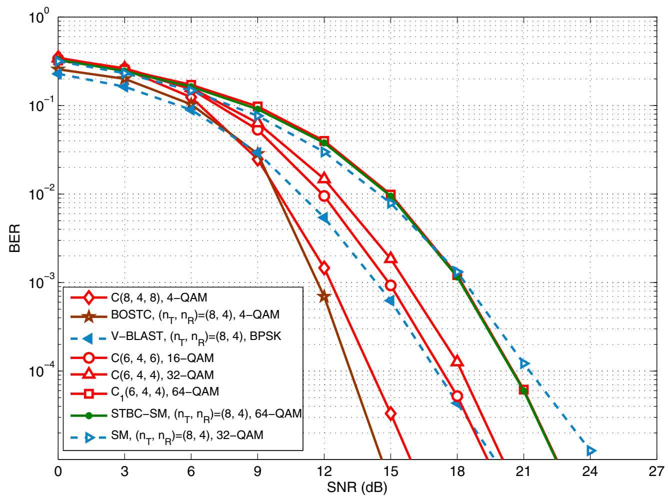

Fig. 7. BER curves of $C(6,4,6)$, $C_1(6,4,4)$, $C(6,4,4)$, $C(8,4,8)$, STBC–SM, SM, V–BLAST and BOSTC for $(n_T, n_R) = (8,4)$. The spectral efficiency is 8 bpcu.

the rate-2 Srinath STBC scheme by removing 2 complex symbols. Appropriate modulation techniques are employed in order to guarantee the same spectral efficiency of 6 bpcu. The proposed SM–OSTBC scheme outperforms all the other MIMO schemes, with the exception of the rate-1.5 Srinath STBC scheme. For example, at $\mathrm{BER} = 10^{-5}$, the SNR gain achieved by the proposed MIMO scheme is approximately 1 dB compared to the Sezginer STBC scheme, about 2 dB compared to the STBC–SM scheme[6] and the corresponding minimum determinant is equal to $\delta_{\min} = 9.03$, and about 5 dB compared to the Alamouti STBC sheme. On the other hand, the proposed $C(4,4,4)$ scheme is less than 0.7 dB worse than the Srinath STBC scheme, but it provides a lower demodulation complexity.

Figure 6 shows the BER performance of the $C(6,2,6)$, $C_1(6,2,4)$, $C(6,2,4)$, V–BLAST, QO–STBC, Perfect Code, STBC–SM, and SM schemes. All the schemes have the same spectral efficiency of 6 bpcu. The SM–OSTBC, QO–STBC, Perfect Code, and V–BLAST schemes use 4-QAM, 64-QAM, 4-HEX, and BPSK modulations, respectively, with (n_T, n_R)= $(6,2)$. On the other hand, STBC–SM and SM schemes use 16-QAM and 8-QAM modulations, respectively, with $(n_T, n_R) = (8,2)$. The V–BLAST and SM scheme provide a diversity order of 2, leading to the worse BER performance compared to the other MIMO arrangements. Among the analyzed schemes, the MIMO scheme using the Perfect Code achieves the best BER performance since it offers full diversity order of 12. The $C(6,2,6)$ and $C(6,2,4)$ SM–OSTBC schemes show a performance loss of about 3.3 dB and 3.6 dB at $\mathrm{BER} = 10^{-5}$, compared to the Perfect Code. The performance gap between the $C(6,2,6)$ scheme and the QO–STBC scheme is about 1.6 dB at BER = 10^{-5}. Compared to the STBC–SM, $C(6,2,6)$ provides approximately 1.9 dB gain at $\mathrm{BER} = 10^{-5}$ and it can also reduce the number of transmit antennas. It is also to be noted that $C_1(6,2,6)$ and STBC–SM schemes have the same BER.

Figure 7 compares the achievable BER of $C(6,4,6)$, $C_1(6,4,4)$, $C(6,4,4)$, $C(8,4,8)$, STBC–SM, V–BLAST, rate-4 BOSTC and SM for the same spectral efficiency of 8 bpcu. The STBC–SM, V–BLAST, BOSTC and SM schemes have $(n_T, n_R) = (8,4)$. The rate-4 BOSTC is obtained by removing 32 complex symbols from the original rate-8 BOSTC matrix. It can be observed that the V–BLAST scheme has the best BER performance for low SNR, since it uses the smallest constellation size (i.e., BPSK modulation). However, as the SNR increases, it is outperformed by the $C(8,4,8)$ and BOSTC schemes, due to its low diversity order. At $\mathrm{BER} = 10^{-5}$, the SNR difference between BOSTC and $C(8,4,8)$ is about 1.3 dB in favor of the BOSTC scheme. Similar to Fig. 6, we observe that the $C(6,4,6)$ SM–OSTBC scheme outperforms both the STBC–SM[7] and the SM schemes. The SNR gain is approximately 3 dB and 4.8 dB, respectively, despite the fact that the proposed scheme requires fewer transmit antennas.

As discussed in [35], future wireless transmission technologies should take into account not only conventional performance factors, such as throughput, Quality-of-Service (QoS), etc., but also power consumption and system complexity. This motivates us to evaluate the energy efficiency and detection complexity of our proposed scheme compared to other MIMO schemes. In Fig. 8, the BER performance of $C(6,4,4)$, $C(8,4,6)$, STBC–SM, Perfect Code and BOSTC are evaluated by taking into account the total power consumption, under the assumption that a urban macro base station is used. The total power consumption accounts for both the radiated power and the circuit power, with the latter being mainly determined by the number of RF chains, i.e., the power amplifiers needed for implementing the MIMO scheme [43], [44]. The simulation parameters are summarized in Table I. Perfect Code and BOSTC schemes are equipped with n_T RF chains, while the proposed schemes are equipped with $n_A = (n_T - 2)$ RF chains and the STBC–SM scheme is equipped with 2 RF chains.

[6]By using a computer search, the optimum rotation angle for the STBC–SM scheme employing 32-QAM is equal to $\theta_{opt} = 0.82$ radians.

[7]The optimum rotation angles of the STBC–SM scheme with 16-QAM are used for the STBC–SM scheme with 64-QAM.

TABLE I
SIMULATION PARAMETERS

Simulation Parameters	Values
BS Type	Urban Macro
Power Model Parameters	SOTA 2010 [49]
Carrier Frequency	2 GHz
Path Loss Model	3GPP NLOS [50]
Bandwidth	5 MHz
Temperature	298 K
Distance from mobile station to BS	200 m
Maximum Power Consumption	1350 W

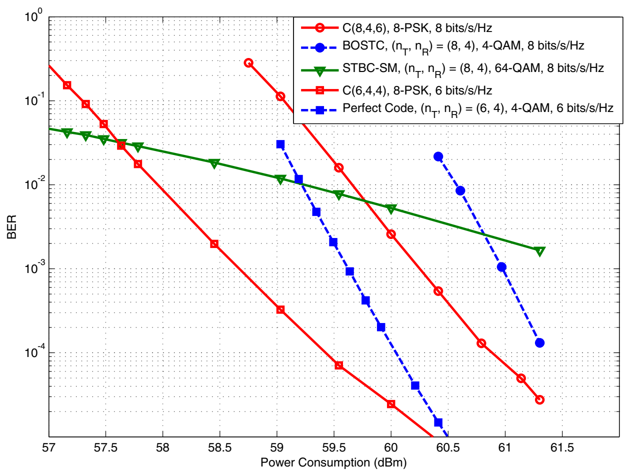

Fig. 8. BER of $C(6,4,4)$, $C(8,4,6)$, STBC–SM, Perfect Code and BOSTC schemes as a function of the total power consumption (horizontal axis) of a urban macro base station.

Figure 8 shows that the proposed schemes outperform SOTA schemes if the total power consumption is taken into account, especially if $\mathrm{BER} \geq 10^{-4}$. At $\mathrm{BER} = 10^{-4}$, our schemes allow a BS to reduce the power consumption of about 0.5 dB, which is equivalent to roughly 12% of power saving. The power saving increases by a factor of two at $\mathrm{BER} = 10^{-3}$. By contrast, the STBC–SM scheme is more power-efficient than the $C(8,4,6)$ scheme in the high BER region, i.e., for $\mathrm{BER} \geq 10^{-2}$, because it needs a smaller number of RF chains. However, the opposite holds if $\mathrm{BER} < 6 \times 10^{-3}$.

The detection complexity of $C(4,2,4)$, $C(4,4,4)$, $C(6,2,6)$, $C(8,4,8)$, rate-2 Srinath STBC, rate-1.5 Srinath STBC, Perfect Code and BOSTC schemes is studied in Fig. 9. Here, SOTA codes are decoded by using the Schnorr-Euchner SD (SE–SD) of [51]. To the best of the authors' knowledge, the SE–SD is among the best SD available in the literature, providing very low demodulation complexity. The complexity of the SE–SD decoder, which includes preprocessing and search stages, can be shown to be equal to $C = 2MLN + 2ML^2 + 2ML - \frac{L^2+L}{2} + \bar{F}_S$ (flops), where $M = 2n_R T$, $N = 2n_T T$, $L = 2K$, K is the number of transmitted symbols per code block, and $\bar{F}_S$ is the average complexity of the searching stage. The average number of flops per bit for SOTA codes is therefore equal to $C_{SOTA} = \frac{C}{mK}$, where m is the number of bits per transmitted symbol. Figure 9 highlights that, for the same MIMO setup and for the same spectral efficiency, the proposed codes require a significantly lower decoding complexity compared to SOTA codes. Specifically, at $\mathrm{SNR} = 15$ dB, the number of flops per bit required by the $C(4,2,4)$ and $C(4,4,4)$ schemes is about 7 and 5.3 times less than that of the rate–2 and rate–1.5 Srinath STBCs, respectively. The complexity ratio of $C(8,4,8)$ and BOSTC schemes is approximately 49.6, while that of $C(6,2,6)$ and Perfect Code is about 506.

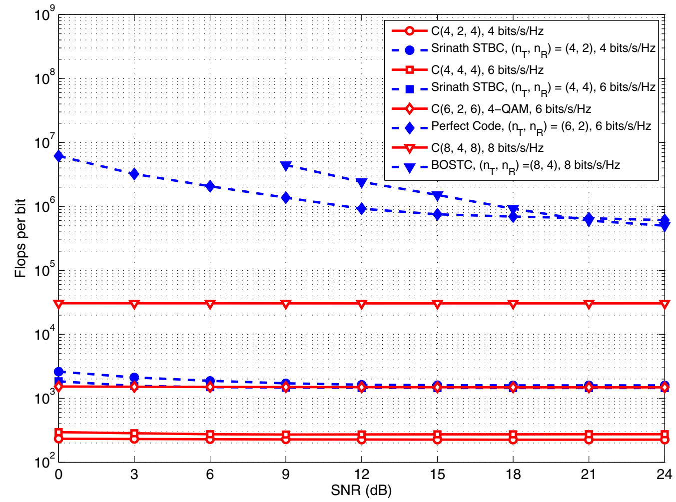

Fig. 9. Detection complexity of $C(4,2,4)$, $C(4,4,4)$, $C(6,2,6)$, $C(8,4,8)$, rate-2 Srinath STBC, rate-1.5 Srinath STBC, Perfect Code and BOSTC schemes for different antenna configurations and spectral efficiencies. The proposed codes are demodulated by using the SO-SD decoder.

From these results, we conclude that, compared to Srinath STBC, Perfect Code and BOSTC schemes, the proposed SM–OSTBC schemes provide the following advantages:

- They allow MIMO implementations with reduced total power consumption, especially for low/medium SNR, thanks to the reduced number of RF chains.
- They have a lower demodulation complexity, by still providing comparable or slightly worse BER performance.
- The demodulator can be implemented by using parallel computing technique, thereby improving the decoding speed without degrading the BER.

Finally, Fig. 10 illustrates the BER of SM–OSTBC schemes for various MIMO setups and modulations providing spectral efficiencies of 7.5 and 8 bpcu. It can be observed that, for a fixed n_A, a better BER can be obtained by increasing n_T since the modulation order can be reduced. Interestingly, the BER performance of $C(4,4,4)$ and $C_1(4,4,4)$ is almost the same in the high SNR region, although $C(4,4,4)$ provides a higher spectral efficiency compared to $C_1(4,4,4)$. This result can be explained as follows. In SM–OSTBC, bit errors occur if either the SC codeword or the Alamouti codeword or both of them are detected erroneously. When a high-order modulation scheme is utilized, e.g., 64–QAM modulation, this is the dominant factor that affects the BER performance of an SM–OSTBC system.

The simulation results of Fig. 6, 7 and 10 suggest the following conclusions.

- In order to deliver a high spectral efficiency with good BER performance, a SM–OSTBC scheme should be equipped with many transmit n_T and active n_A antennas in order to reduce the modulation size. However, the

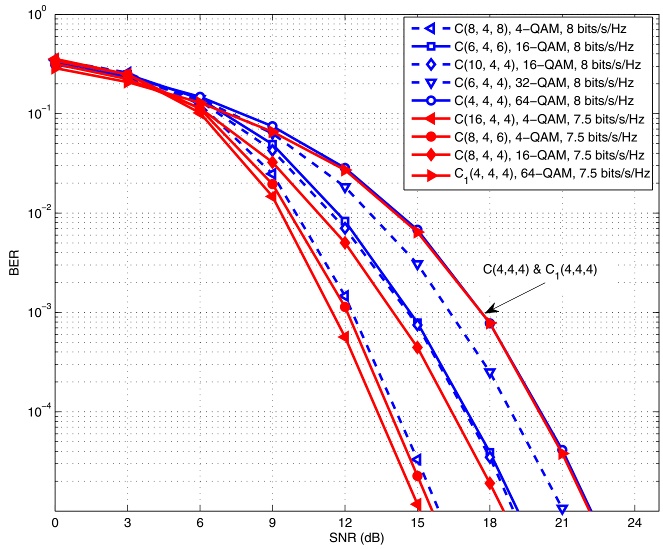

Fig. 10. BER curves of SM–OSTBC schemes as a function of n_T, n_A and the modulation.

larger n_A the more the RF chains, which affect hardware cost, system complexity and total power dissipation.

- Given the number of RF chains n_A, increasing n_T instead of M is the best option for increasing the spectral efficiency and for providing good BER performance.

VI. Conclusions

In this paper, we have proposed a novel high-rate, high-performance and low-complexity MIMO transmission scheme, which is refereed to as SM–OSTBC. In order to strike a flexible trade-off between performance, energy-efficiency and system complexity, code constructions for MIMO setups when either all or a subset of the available transmit antennas are activated have been provided. It has been proved that the proposed SM–OSTBCs attain: 1) the NVD property, 2) a transmit-diversity order equal to two, and 3) a maximum spectral efficiency of $(n_T - 2 + \log_2 M)$ bpcu. By capitalizing on a linear dispersion representation and on the orthogonality of the Alamouti STBC, we have proposed a single-stream low-complexity ML decoder and a single-stream SD with near-optimum performance. Simulation results have demonstrated that the proposed SM–OSTBC schemes are capable of outperforming many existing MIMO arrangements, including the Alamouti STBC, Sezginer STBC, SM, STBC–SM and V–BLAST schemes. Compared to Srinath STBC, Perfect Code, and BOSTC schemes, the proposed SM–OSTBC schemes offer slightly worse BER performance if only the transmit power is taken into account. On the other hand, if the total power dissipation is considered, the proposed SM–OSTBC schemes outperform the other MIMO schemes. Furthermore, they can be demodulated with much lower signal processing complexity.

Appendix

Proof of Proposition 1

Let us consider two different SM–OSTBC codewords $\mathbf{C} \neq \hat{\mathbf{C}}$. The codeword difference matrix $\mathbf{B}(\mathbf{C}, \hat{\mathbf{C}})$ and the codeword distance matrix $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ are defined as:

$$\mathbf{B}(\mathbf{C}, \hat{\mathbf{C}}) = \mathbf{C} - \hat{\mathbf{C}} = \mathbf{S}\mathbf{X} - \hat{\mathbf{S}}\hat{\mathbf{X}} \tag{40}$$

$$\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}}) = \mathbf{B}^H(\mathbf{C}, \hat{\mathbf{C}})\mathbf{B}(\mathbf{C}, \hat{\mathbf{C}}) \tag{41}$$

In this section, we will show that $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ has rank two for any $\mathbf{C} \neq \hat{\mathbf{C}}$ and that its minimum determinant is given in (6). Thee cases have to be studied: 1) $\mathbf{S} = \hat{\mathbf{S}}, \mathbf{X} \neq \hat{\mathbf{X}}$, 2) $\mathbf{S} \neq \hat{\mathbf{S}}, \mathbf{X} = \hat{\mathbf{X}}$, and 3) $\mathbf{S} \neq \hat{\mathbf{S}}, \mathbf{X} \neq \hat{\mathbf{X}}$.

1) Case 1: $\mathbf{S} = \hat{\mathbf{S}}, \mathbf{X} \neq \hat{\mathbf{X}}$

We can write:

$$\mathbf{B}(\mathbf{C}, \hat{\mathbf{C}}) = \mathbf{C} - \hat{\mathbf{C}} = \mathbf{S}(\mathbf{X} - \hat{\mathbf{X}}) \tag{42}$$

Thus, the codeword distance matrix is:

$$\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}}) = (\mathbf{X} - \hat{\mathbf{X}})^H\mathbf{S}^H\mathbf{S}(\mathbf{X} - \hat{\mathbf{X}}) \tag{43}$$

It is easy to verify that $\mathbf{G}^H(\mathbf{s})\mathbf{G}(\mathbf{s}) = \mathbf{I}_2$ for any $\mathbf{s}$, and hence $\mathbf{S}^H\mathbf{S} = \mathbf{I}_2$, where $\mathbf{I}_2$ is a 2×2 identity matrix. Equation (43) then simplifies to:

$$\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}}) = (\mathbf{X} - \hat{\mathbf{X}})^H(\mathbf{X} - \hat{\mathbf{X}}) = \sum_{n=1}^{2} |x_n - \hat{x}_n|^2\mathbf{I}_2 \tag{44}$$

Since $\mathbf{X} \neq \hat{\mathbf{X}}$, or equivalently, $(x_1, x_2) \neq (\hat{x}_1, \hat{x}_2)$, the distance matrix $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ has a full column rank of 2.

Without loss of generality, we assume that $x_1 \neq \hat{x}_1$. Then, the determinant of $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ is as follows:

$$D_1 = \left[\sum_{n=1}^{2} |x_n - \hat{x}_n|^2\right]^2 \geq |x_1 - \hat{x}_1|^4 \geq D_{1,\min} \tag{45}$$

where:

$$D_{1,\min} = \begin{cases} 16 & \text{for } M\text{-QAM} \\ 16\sin^4\frac{\pi}{M} & \text{for } M\text{-PSK} \end{cases} \tag{46}$$

2) Case 2: $\mathbf{S} \neq \hat{\mathbf{S}}, \mathbf{X} = \hat{\mathbf{X}}$

We can write:

$$\mathbf{B}(\mathbf{C}, \hat{\mathbf{C}}) = \mathbf{C} - \hat{\mathbf{C}} = (\mathbf{S} - \hat{\mathbf{S}})\mathbf{X} \tag{47}$$

Thus, the codeword distance matrix is:

$$\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}}) = \mathbf{X}^H(\mathbf{S} - \hat{\mathbf{S}})^H(\mathbf{S} - \hat{\mathbf{S}})\mathbf{X} \tag{48}$$

In general, we have:

$$\begin{aligned}(\mathbf{S} - \hat{\mathbf{S}})^H(\mathbf{S} - \hat{\mathbf{S}}) &= (\mathbf{G}(\mathbf{s}) - \mathbf{G}(\hat{\mathbf{s}}))^H(\mathbf{G}(\mathbf{s}) - \mathbf{G}(\hat{\mathbf{s}})) \\ &= \frac{1}{\Gamma^2}\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2\mathbf{I}_2 \end{aligned} \tag{49}$$

It follows that:

$$\begin{aligned}\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}}) &= \frac{1}{\Gamma^2}\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2\mathbf{X}^H\mathbf{X} \\ &= \frac{1}{\Gamma^2}\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2 \sum_{k=1}^{2} |x_k|^2\mathbf{I}_2 \end{aligned} \tag{50}$$

Since $\mathbf{S} \neq \hat{\mathbf{S}}$, or equivalently, $\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2 > 0$, the distance matrix $\mathbf{A}(\mathbf{C}, \hat{\mathbf{C}})$ has a full column rank of two.

The determinant of $\mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big)$ is as follows:

$$D_2 = \frac{1}{\Gamma^4}\left[\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2\right]^2 \left[\sum_{k=1}^{2} |x_k|^2\right]^2 \tag{51}$$

Since $\mathbf{S} \neq \hat{\mathbf{S}}$, it is reasonable to assume that $s_b \neq \hat{s}_b$, where $s_b, \hat{s}_b \in \{\pm 1, \pm j\}$ are the bth entries of vectors $\mathbf{s}$ and $\hat{\mathbf{s}}$, respectively. So, we can write:

$$\left[\sum_{n=1}^{n_T} |s_n - \hat{s}_n|^2\right]^2 \geq |s_b - \hat{s}_b|^4 \geq 4 \tag{52}$$

In addition, we have:

$$\left[\sum_{k=1}^{2} |x_k|^2\right]^2 \geq \begin{cases} 16 & \text{for } M\text{-QAM} \\ 4 & \text{for } M\text{-PSK} \end{cases} \tag{53}$$

It follows from (51), (52), and (53) that:

$$D_2 \geq D_{2,\min} \tag{54}$$

where:

$$D_{2,\min} = \begin{cases} \frac{64}{\Gamma^4} & \text{for } M\text{-QAM} \\ \frac{16}{\Gamma^4} & \text{for } M\text{-PSK} \end{cases} \tag{55}$$

3) Case 3: $\mathbf{S} \neq \hat{\mathbf{S}}$ and $\mathbf{X} \neq \hat{\mathbf{X}}$

In this case, the codeword difference matrix is:

$$\begin{aligned} \mathbf{B}\big(\mathbf{C}, \hat{\mathbf{C}}\big) &= \mathbf{C} - \hat{\mathbf{C}} \\ &= \frac{1}{\Gamma}\begin{bmatrix} c_1 - \hat{c}_1 & -c_2^* + \hat{c}_2^* \\ c_2 - \hat{c}_2 & c_1^* - \hat{c}_1^* \\ \vdots & \vdots \\ c_{n_T-1} - \hat{c}_{n_T-1} & -c_{n_T}^* + \hat{c}_{n_T}^* \\ c_{n_T} - \hat{c}_{n_T} & c_{n_T-1}^* - \hat{c}_{n_T-1}^* \end{bmatrix} \end{aligned} \tag{56}$$

Consequently, the codeword distance matrix is:

$$\begin{aligned} \mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big) &= \mathbf{B}^H\big(\mathbf{C}, \hat{\mathbf{C}}\big)\mathbf{B}\big(\mathbf{C}, \hat{\mathbf{C}}\big) \\ &= \frac{1}{\Gamma^2}\sum_{n=1}^{n_T} |c_n - \hat{c}_n|^2 \mathbf{I}_2 \end{aligned} \tag{57}$$

Since $s_1 = \hat{s}_1 = 1$, $s_2 = \hat{s}_2 = 1$ and $(x_1, x_2) \neq (\hat{x}_1, \hat{x}_2)$, the summation on the right-hand side of (57) can be written as follows:

$$\begin{aligned} \sum_{n=1}^{n_T} |c_n - \hat{c}_n|^2 &\geq \sum_{n=1}^{2} |c_n - \hat{c}_n|^2 \\ &= |x_1 - \hat{x}_1 + x_2 - \hat{x}_2|^2 \\ &\quad + |-(x_1 - \hat{x}_1) + x_2 - \hat{x}_2|^2 \\ &= 2\sum_{n=1}^{2} |x_n - \hat{x}_n|^2 > 0 \end{aligned} \tag{58}$$

From (57) and (58), we conclude that $\mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big)$ has a column rank of two.

The determinant of $\mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big)$ is equal to:

$$D_3 = \frac{1}{\Gamma^4}\left[\sum_{n=1}^{n_T} |c_n - \hat{c}_n|^2\right]^2 \tag{59}$$

By using (45), (46) and (58), we can write:

$$D_3 \geq \frac{4}{\Gamma^4}\sum_{n=1}^{2} |x_n - \hat{x}_n|^2 \geq D_{3,\min} \tag{60}$$

where:

$$D_{3,\min} = \frac{4}{\Gamma^4} D_{1,\min} = \begin{cases} \frac{64}{\Gamma^4} & \text{for } M\text{-QAM} \\ \frac{64}{\Gamma^4}\sin^4\frac{\pi}{M} & \text{for } M\text{-PSK} \end{cases} \tag{61}$$

Therefore, the minimum determinant for the proposed SM–OSTBC scheme is:

$$\delta_{\min} = \min\left(D_{1,\min}, D_{2,\min}, D_{3,\min}\right). \tag{62}$$

For an M-QAM constellation, we have:

$$\delta_{\min,QAM} = \frac{64}{\Gamma^4} \tag{63}$$

which is independent of M.

For BPSK (2-PSK) and QPSK (4-PSK), we have

$$\delta_{\min,BPSK} = \delta_{\min,QPSK} = \frac{16}{\Gamma^4} \tag{64}$$

For M-PSK with $M \geq 8$, with the aid of simulation results, we conclude that $D_{3,\min} \geq 16\sin^4\frac{\pi}{M}$. Consequently, the minimum determinant for M-PSK with $M \geq 8$ is equal to

$$\delta_{\min,M-PSK} = 16\sin^4\frac{\pi}{M} \tag{65}$$

In conclusion, $\mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big)$ has a full column rank of 2 for any pair of distinct SM–OSTBC codewords $\big(\mathbf{C}, \hat{\mathbf{C}}\big)$. Therefore, the proposed SM–OSTBC scheme is capable of achieving a transmit diversity order of 2. In addition, the minimum determinant of $\mathbf{A}\big(\mathbf{C}, \hat{\mathbf{C}}\big)$ for different modulations are given in (63)-(65) and (6). This concludes the proof.

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