LOW-COMPLEXITY DETECTION AND TRANSMIT ANTENNA SELECTION FOR SPATIAL MODULATION

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Abstract: In this paper, we propose a novel low-complexity detection technique for conventional Spatial Modulation (SM). The proposed scheme is compared to SM with optimal detection (SM-OD), SM with signal vector based detection (SM-SVD) and another reduced complexity detection technique, presented in the literature. The numerical results show that the proposed scheme can match the error performance of SM-OD very closely, even at low bit-error rate (BER). The computational complexity at the receiver, is shown to be independent of the symbol constellation size, and hence, offers a much lower complexity compared to existing schemes.
To further improve the error performance, we then propose two closed-loop transmit antenna selection (TAS) schemes for conventional SM. We assume the receiver has knowledge of the channel and a perfect low-bandwidth feedback path to the transmitter exists. From evaluation of the BER versus normalized signal-to-noise ratio (SNR) and the complexity analysis, the proposed schemes exhibit a significant improvement over SM-OD and other improved SM schemes in terms of error performance and/or complexity.

Keywords: Spatial modulation, transmit antenna selection, multiple-input multiple-output, adaptive spatial modulation, reduced complexity detection, signal vector based detection, optimal detection.

I. INTRODUCTION

The use of multiple transmit and receive antennas in mobile communication systems will become more prominent in 4G systems and beyond. These multiple-input multiple-output (MIMO) systems allow unprecedented improvements (in terms of throughput and error performance) compared to their single-input single-output (SISO) counterparts due to diversity, multiplexing or antenna gain (smart antennas) or a combination of these. The primary inhibiting factors for the implementation of these MIMO techniques have been the system complexity and cost.

SM is an attractive MIMO technique that has been presented fairly recently in the literature, and holds the promise of reduced system complexity and cost [1]. The key idea behind SM is to select only one antenna from the available transmit antennas for transmission of the symbol in the transmission interval. Thus, only one transmit antenna is active at any particular transmission instant.

The SM symbol has two parts. The first part of the symbol represents the chosen antenna index, while the second part of the symbol is based on the amplitude and/or phase modulation (APM), e.g. quadrature amplitude modulation (QAM) and phase shift keying (PSK). It is important to note that the first part of the symbol is not transmitted explicitly, but rather by the particular antenna index. This implies that the active antenna index must be detected at the receiver to estimate the complete symbol.

The advantages of SM compared to other MIMO schemes, e.g. vertical Bell laboratories layered space-time architecture (V-BLAST), and Alamouti schemes [1], [2] are, a) avoids inter-channel interference (ICI) at the receiver [1], b) reduced receiver complexity [1], [3], c) inter-antenna synchronization (IAS) at the transmitter is not required, due to the single active antenna per signaling interval [1], d) high spectral efficiency with no ICI or IAS [4], e) SM requires a single radio frequency (RF) chain [1], and f) SM has been shown to outperform many existing MIMO schemes including V-BLAST [1], [2]. All of these also contribute to g) lower cost. These important advantages make SM an important technique for future high performance wireless communications.

The detection scheme employed for SM has much importance, and has been investigated intensively. In [1], Mesleh *et al*. proposed the use of iterative maximal ratio combining (*i*-MRC). The *i*-MRC technique iteratively computes the product of the channel gain and received signal vectors to estimate the active transmit antenna index. Then, by using the active transmit antenna index estimate, the transmit symbol is estimated. However, this method was shown to be sub-optimal [2].

In [2], Jeganathan *et al*. derived the optimal detection rule by means of the maximum-likelihood (ML) principle. The proposed detector exhibited a gain of approximately 4 dB (constrained channels [2]) at a BER of 10^{-5} [2], compared to the sub-optimal technique [1], and was further supported by analytical results. However, the complexity analysis in [2] showed that both the optimal

and sub-optimal detection techniques exhibit a similar level of complexity, which is still relatively high.

Meanwhile, [3] showed that for high-order modulation schemes ($M \geq 16$), the optimal detector [2] complexity can increase by approximately 125% in comparison to the sub-optimal scheme [1], [2]. Based on this observation, [3] presented a sub-optimal multiple stage detection technique. In [3], in the first stage the original SM detector for conventional channels [2] is used to determine the most probable estimates of the active transmit antenna index. A final estimate of the active transmit antenna index and transmit symbol is then computed by means of the ML principle [2]. This method showed a decrease in the receiver computational complexity, compared to the schemes in [1] and [2]. In addition, the error performance closely matched that of the optimal detector.

To further reduce complexity, Xu [4] presented a simplified ML based detection scheme for multilevel-quadrature amplitude modulation (M-QAM) SM; this scheme was based on searching partitioned signal sets for the active transmit antenna index and transmit symbol, thus reducing the computational complexity. Xu [4] then applied the method to the multiple stage detection proposed in [3], which again exhibited a further reduction in complexity, compared to existing work.

In [5], [6], signal vector based detection (SVD) was proposed. SVD determines the active transmit antenna index at the receiver by searching for the smallest angle between the channel gain vectors and the receive vector. The transmit symbol is then estimated using the ML principle [2], [5], [7]. Wang *et al*. [5] then motivated that the SVD detector BER matched that of the optimal detector very closely at a much reduced complexity. However, in [6], it was shown that for moderate to high SNRs the error performance does not match that of SM-OD.

To improve on the error performance of SVD, Zheng presented SVD-List detection [7]. However, the complexity was higher [7] compared to SVD [5], due to the selection and search of a list of candidate antennas based on the L smallest angles between the channel gain vectors and receive vector. In [8], a reduced complexity scheme (we will refer to as SM-RC) for achieving optimal error performance for SM was presented. In [8], the transmit antenna index and symbol are detected separately; however, by taking the correlation into account (since the antenna index and transmit symbol fade together) the scheme retains the error performance of SM-OD. However, the scheme works with demodulating bits rather than symbols, and the computational complexity is a function of the symbol constellation size (as with all the existing detection schemes for SM), meaning that for large constellation dimension the complexity will still be relatively high.

Sphere decoding for SM has also been presented in the literature and exhibits good BER performance [9], [10]; however, it has been shown that the lower bound of the computational complexity, at the receiver, is once again a function of the symbol constellation size.

Motivated by this, in this paper, we first propose a low-complexity detection technique for conventional SM (SM-LCD), which has a computational complexity independent of the symbol constellation size. The motivation for our proposed scheme is based on a single antenna transmission system and is as follows: For a single antenna transmission system, it is relatively straightforward to compute the equalized symbol, and consequently, the estimate of the transmit symbol [11].

However, for SM, although only one transmit antenna is active per transmission instant (the active antenna is selected by mapping the first $\log_2 N_T$ symbol bits to the antenna index), at the receiver, knowledge of the active antenna is unknown. Thus, in the proposed detection scheme, the estimate of the transmit symbol for each transmit antenna in the array is determined using the corresponding equalized symbol estimate. Finally, the estimated transmit symbol for each of the transmit antennas are compared to the received signal vector and the closest estimate (based on the ML principle) is chosen as the final estimate of the antenna index/transmit symbol. The optimal detector performs the search over M symbol constellation points for each antenna [2]. On the contrary, in SM-LCD there is only a single constellation point for each transmit antenna, hence reducing the computational complexity at the receiver, especially for high-order symbol constellations. In Section III, the proposed SM-LCD scheme is explained in more detail.

As noted earlier, [2] demonstrated how optimal error performance can be achieved for SM using the ML principle. However, the error performance of SM is still dependent on the selected transmit antenna and the resulting propagation paths between the transmitter and receiver. Several works have considered antenna selection for MIMO to improve performance. In [12], Molisch *et al*. motivated that optimum selection of the antenna elements requires an exhaustive search of all possible combinations for either the best SNR (spatial diversity) or capacity (spatial multiplexing). This requires computation of determinants for each channel realization, thus making this impractical [12]. On this note, Molisch *et al*. [12] motivated for antenna selection at only one end of the MIMO system.

Several works have investigated the application of single-ended transmit or receive antenna selection for MIMO systems. In [13], [14], the authors investigate TAS to reduce energy consumption in MIMO systems. Single TAS based on selecting the antenna with the highest equivalent receive SNR was proposed in [13]. Multiple TAS was investigated in [14], where the antennas corresponding to the highest channel gains were selected

to form the active subset of the antenna array. In this paper, we refer to this method as maximal gain TAS (MG-TAS). A correlation-based method (CBM) was proposed in [15] for antenna selection at the receiver for MIMO systems, and showed improved performance. However, none of these schemes have been applied to SM.

Furthermore, recently several works have shown that the BER performance of SM can be improved through closed-loop design. In [16], an adaptive SM scheme was proposed to achieve an improved system performance for fixed data rates. The key idea behind the scheme presented in [16] was to select the optimum modulation candidate (the candidate that maximizes the received minimum distance) for transmission. The candidate was chosen by searching among a set of all possible candidates. The transmitter then employs the chosen modulation orders in the next channel use. On the same note, in [17], the authors presented two more adaptive closed-loop SM schemes, which additionally use transmit-mode switching to further improve on the performance. The first scheme, optimal-hybrid SM (OH-SM) proposed in [17], uses both adaptive modulation and transmit-mode switching. The second scheme, concatenated SM (C-SM) was then proposed in [17], and used a multistage adaptation technique to balance the trade-off between the computational complexity and error performance. However, the existing closed-loop schemes have serious drawbacks. The first drawback of the schemes proposed in [16] and [17] was the very high complexity, e.g. for 4 bits/s/Hz the complexity of OH-SM was approximately 40,000 real multiplications, while the complexity of C-SM was approximately 20,000 real multiplications [17].

Secondly, in SM the antenna indices carry information; however, in the proposed closed-loop schemes [16], [17], the antenna indices also carry information about the transmission mode and there is no method presented to take care of the errors when the information at the receiver is inconsistent with that at the transmitter and the transmission mode is unknown.

In addition, in SM, based on SVD [5], the BER of SM not only depends on the amplitude of the channel gain vectors but also on the angles between the j channel gain vectors. Thus, based on these factors as well as the drawbacks of the existing closed-loop designs for SM [16], [17], we propose a more practical closed-loop design for SM, which uses antenna selection at the transmitter.

Specifically, we propose two TAS schemes for SM which exhibits much lower complexity and/or an improved error performance compared to the existing schemes. The first scheme selects transmit antennas based on the largest channel gain amplitudes, then the largest angles between channel gain vectors. The second scheme selects transmit antennas based on the channel gain vectors which have the largest amplitudes followed by the smallest correlations. The new schemes are also compared to MG-TAS, which we have applied to SM (MG-TAS-SM).

The structure of the remainder of this paper is as follows: In Section II, the system model for conventional SM is presented and supported with brief background theory. In Section III, we present the proposed SM-LCD scheme and the complexity analysis, where we show SM-LCD has a computational complexity, at the receiver (in terms of the number of complex multiplications and additions), independent of the symbol constellation size. The TAS algorithms for SM are then presented in Section IV and the complexity analysis is included. Section V presents the numerical analysis for the proposed schemes and draws comparisons with existing schemes. Conclusions are finally drawn in Section VI.

II. System Model & Background theory

Consider a MIMO communication system with transmit antenna array of length N_T and receive antenna array of length N_R. The received signal vector $\boldsymbol{y}$ can be denoted as [1], [2],

$$\boldsymbol{y} = \sqrt{\rho}\boldsymbol{H}\boldsymbol{X} + \boldsymbol{\eta} \tag{1}$$

where $\boldsymbol{X}$ is an $N_T \times 1$ vector, with one non-zero entry corresponding to the transmit symbol (we consider M-QAM with Gray mapping). $\boldsymbol{H}$ is an $N_R \times N_T$ complex channel gain matrix with independent and identically distributed (i.i.d) entries with distribution $CN(0,1)$. ρ represents the average SNR per receive antenna and $\boldsymbol{\eta}$ represents an $N_R \times 1$ complex additive white Gaussian noise (AWGN) vector with i.i.d entries with distribution $CN(0,1)$.

In [1], [2], the sub-optimal detection rule for conventional channels [2] was defined as,

$$\hat{j} = \underset{j}{\operatorname{argmax}} \left(\frac{|\boldsymbol{h}_j^{\dagger}\boldsymbol{y}|}{\|\boldsymbol{h}_j\|_F^2} \right) \tag{2}$$

$$\hat{q} = \mathcal{D}\left(\frac{\boldsymbol{h}_{\hat{j}}^{\dagger}\boldsymbol{y}}{\|\boldsymbol{h}_{\hat{j}}\|_F^2} \right) \tag{3}$$

where $\hat{j}$, $\hat{q}$ represent the estimates of the active transmit antenna index j and symbol index q, respectively. Note that $j \in [1{:}N_T]$ and $q \in [1{:}M]$. $\boldsymbol{h}_j$ represents a column vector of $\boldsymbol{H}$ and is defined as $\boldsymbol{h}_j = [h_{1j} \quad h_{2j} \quad \dots \quad h_{N_R j}]^T$. $\dagger$ represents the transpose conjugate operator, $\underset{j}{\operatorname{argmax}}(\cdot)$ represents the argument of the maximum with respect to j, $\|\cdot\|_F$ represents the Frobenius norm, $|\cdot|$ represents the magnitude operation, $\mathcal{D}(\cdot)$ represents the constellation demodulator or slicing function, which extracts the in-phase and quadrature-phase components from the symbol argument, and $[\cdot]^T$ represents the matrix or vector transposition.

Jeganathan *et al*. [2] presented the optimal joint detection rule based on the ML principle. The detection rule was defined as [2],

$$[\hat{j},\hat{q}] = \underset{j,q}{\operatorname{argmin}} \sqrt{\rho}\left\|\boldsymbol{l}_{jq}\right\|_F^2 - 2Re\{\boldsymbol{y}^{\dagger}\boldsymbol{l}_{jq}\} \tag{4}$$

where $\boldsymbol{l}_{jq} = \boldsymbol{h}_j x_q$ and $Re\{\cdot\}$ represents the real part of the complex argument. Note that (4) is a joint detection rule and M symbol constellation points are searched for each transmit antenna j, thus the method has a high computational complexity at the receiver, and yields optimal error performance [2], [3].

To reduce the high complexity due to joint detection of antenna and symbol indices, Xu *et al*. [8] proposed a detection technique where the active transmit antenna index and symbol are detected separately using decorrelating vectors [8]. To detect the active antenna index, the maximum metric [8] for the known constellation is searched over N_T candidates. Finally, the $\log_2 M$ symbol bits are demodulated based on a decision rule involving the elements of the decorrelating vectors [8]. This method was shown to reduce complexity at the receiver and retain the error performance of SM-OD [8].

In Section III, we propose a new low-complexity detection technique for SM. We refer to the scheme as SM-LCD. In addition, the complexity analysis of the proposed SM-LCD scheme is included. The complexity analysis shows that the computational complexity, at the receiver, for the proposed detector is independent of the symbol constellation size. This significantly reduces the complexity, compared to the existing schemes (from the literature survey we found that the computation complexity for all existing detection schemes for SM is a function of the symbol constellation size).

Motivated by the work performed for antenna selection in MIMO systems [12-15], the potential improvement in error performance by employing closed-loop design for SM [16-17] and the drawbacks of the existing closed-loop designs for SM, in Section IV we propose TAS for SM. In the open literature, MG-TAS has been proposed for single or multiple TAS for MIMO to decrease the average energy consumption [13], [14]. In this paper, we will denote the configuration of MG-TAS as (N_{TOTAL}, N_T, N_R), where N_{TOTAL} is the total number of antennas in the transmit array and $N_{TOTAL} > N_T$.
The MG-TAS scheme [14] proceeds as follows:

Step 1: Compute the channel power gains,

$$P_x = \|\boldsymbol{h}_x\|^2 \qquad 1 \le x \le N_{TOTAL} \tag{5}$$

Step 2: Arrange the channel power gains in descending order,

$$\tilde{P}_1 \ge \tilde{P}_2 \ge \cdots \ge \tilde{P}_{N_{TOTAL}} \tag{6}$$

Step 3: Select the antennas for transmission corresponding to the first N_T elements.

This algorithm has not been applied to SM or to improve error performance for MIMO systems. In Section V (Numerical results), we apply the algorithm to SM (MG-TAS-SM) and utilize it for comparison with the proposed schemes.

The detection of SM symbols requires detection of the active transmit antenna index and the APM symbol at the receiver. Optimal detection of SM was presented in [2]; however, the complexity was relatively high [3]. In [5], Wang *et al*. proposed SVD, a low-complexity technique for detecting the active transmit antenna index at the receiver, for SM. Once the active transmit antenna index was estimated, the ML principle was applied to estimate the APM symbol. Estimation of the active transmit antenna index at the receiver was based on computing the angle between the jth channel gain vector and the received signal vector [5]. The antenna j corresponding to the smallest angle was then selected as the active antenna. This choice was based on the motivation that the received signal vector lies within a sphere of the channel gain vector due to the summation of the scalar product of channel gain vector and QAM symbol and the noise/interference [5]. Motivated by this reasoning, for the first of the proposed TAS algorithms we propose to use in part, a similar technique [5], not for detection but rather to determine the angles between all pairs of channel gain vectors at the transmitter (we assume full channel knowledge at the receiver and a perfect low-bandwidth feedback path to the transmitter). By eliminating the transmit antennas corresponding to the smallest channel gain amplitudes followed by the smallest angles between the channel gain vectors the error performance can be improved by using the reduced transmit antenna array for SM. The second algorithm is a CBM-based method for SM for multiple TAS and involves selecting antennas based on amplitude of the channel gain vectors and correlation between pairs of channel gain vectors.

III. Low-complexity Detection for Spatial Modulation

A. Detection of SM using a low-complexity approach

Assume a SISO system, and suppose the received signal is defined by $y = hx + n$, where h is the channel gain, x is the transmit symbol and n is the AWGN component, then the equalized transmit symbol is given by $\tilde{x}_e = \frac{h^* y}{|h|^2}$ where $(\cdot)^*$ represents the conjugate operator. The final estimate of the transmit symbol is given by $\tilde{x} = \mathcal{D}(\tilde{x}_e)$. In the SM system, the selected transmit antenna used to transmit the symbol is unknown. Thus, the estimate of the transmit symbol for each possible transmit antenna is needed. A comparison of each of the estimated transmit symbols with the receive vector is then performed to determine the most probable final estimate of the antenna index/transmit symbol.

Following this logic, the algorithm for the proposed low-complexity detection of SM proceeds as follows:

Step 1: Compute the equalized symbol for each antenna,

$$z_j = \frac{\boldsymbol{h}_j^{\dagger}\boldsymbol{y}}{\|\boldsymbol{h}_j\|_F^2} \qquad j \in [1{:}N_T] \tag{7}$$

Step 2: Demodulate the estimated symbol for each transmit antenna using,

$$q_j = \mathcal{D}(z_j) \tag{8}$$

where the constellation demodulator function extracts the in-phase and quadrature-phase components from the equalized symbol.

Step 3: Perform detection using the ML principle [2] to estimate the antenna index and transmit symbol,

$$[\hat{j}, q_{\hat{j}}] = \underset{j}{\operatorname{argmin}} \sqrt{\rho}\|\boldsymbol{g}_j\|_F^2 - 2Re\{\boldsymbol{y}^{\dagger}\boldsymbol{g}_j\} \tag{9}$$

where $\boldsymbol{g}_j = \boldsymbol{h}_j q_j$.

It is evident from Step 3, that the search is performed only over a single symbol constellation point for each transmit antenna, compared to the M constellation points searched for each transmit antenna in SM-OD [2], thus reducing complexity. In the next sub-section, the complexity analysis for SM-LCD is presented.

B. COMPLEXITY ANALYSIS FOR SM-LCD

In this sub-section, we represent the computational complexity using the same approach as [1] and [4], i.e. in terms of complex multiplications and additions.

Table 1 presents the complexity at the receiver for each of the existing schemes that are used for comparison with SM-LCD.

Table 1: Computational complexity for SM detection schemes

Scheme	Computational complexity
SM-OD [4]	$\delta_{SM-OD} = (3N_R + M - 1)N_T + M$ (10)
SM-SVD	$\delta_{SM-SVD} = 3N_R N_T + N_R + 2M$ (11)
SM-RC [8]	$\delta_{SM-RC} \in O(\sqrt{M}N_T + 4)$ (12)

Next, we formulate the complexity of the proposed scheme SM-LCD. Step 1 involves the computation of the equalized symbol (7). Assuming a single transmit antenna. The numerator of (7) $\boldsymbol{h}_j^{\dagger}\boldsymbol{y}$ is the multiplication of a $1{\times}N_R$ row vector and the $N_R{\times}1$ received signal vector. This equates to N_R complex multiplications and $N_R - 1$ complex additions. The denominator $\|\boldsymbol{h}_j\|_F^2$ requires N_R complex multiplications and zero complex additions for a single transmit antenna. Thus, for N_T transmit antennas the complexity of Step 1 is given by,

$$\delta_{SM-LCD\_Step\_1} = 3N_R N_T - N_T \tag{13}$$

Step 2 demodulates the estimated symbol using the constellation demodulator function [2], and similarly to [2], [4] we can ignore the complexity of this stage, since the mapping is one-to-one and does not involve any complex multiplications or additions. Therefore, $\delta_{SM-LCD\_Step\_2} = 0$. Step 3 involves the final estimation of the antenna index and transmit symbol and is based on the detection rule (9), similar to [2]. However, as discussed earlier, in this scenario instead of searching over M symbol constellation points for each transmit antenna, the search is performed only over a single constellation point for each transmit antenna. In addition, both $\|\boldsymbol{h}_j\|_F^2$ and $\boldsymbol{y}^{\dagger}\boldsymbol{h}_j$ have been computed in Step 1 and could be stored, thus not further adding to the complexity. Therefore, it can be verified that the complexity of Step 3 is defined by,

$$\delta_{SM-LCD\_Step\_3} = 2N_T \tag{14}$$

Hence the total complexity for SM-LCD can be defined as,

$$\delta_{SM-LCD} = 3N_R N_T + N_T \tag{15}$$

From examination of the expressions in Table 1 and (15), it is immediately clear that the computational complexity of the proposed SM-LCD is not a function of M compared to SM-OD [2], SM-SVD [5] and SM-RC (lower bound) [8]. Thus, we can conclude that the complexity will be significantly reduced, especially as the symbol constellation order increases. The numerical results for the SM-LCD scheme are presented in Section V and compared with existing schemes. In the next section, the proposed antenna selection schemes for SM are presented.

IV. TRANSMIT ANTENNA SELECTION FOR SPATIAL MODULATION

A. TAS-SM

In this sub-section, we present the proposed algorithms for multiple TAS for SM. We refer to the first scheme as TAS-SM and denote its configuration as $(N_{TOTAL}, N_{SA}, N_T, N_R)$, where N_{SA} is the number of antennas in a subset of the array. Note that $N_{TOTAL} > N_{SA} > N_T$. The algorithm begins by selecting N_{SA} out of N_{TOTAL} antennas corresponding to the maximum amplitudes of the respective channel gain vectors [14], and follows by discarding $N_{SA} - N_T$ antennas corresponding to the smallest angles for all possible antenna pairs. The algorithm proceeds as follows:

Step 1: Arrange the channel gain vectors for N_{TOTAL} antennas into the matrix

$$\boldsymbol{H} = [\boldsymbol{h}_1 \quad \boldsymbol{h}_2 \quad \dots \quad \boldsymbol{h}_{N_{TOTAL}}] \tag{16}$$

Step 2: Compute the sum of the magnitudes of each element in the column vectors to yield,

$$\tilde{\boldsymbol{H}} = [\tilde{h}_1 \quad \tilde{h}_2 \quad \dots \quad \tilde{h}_{N_{TOTAL}}] \tag{17}$$

where $\tilde{h}_x = \sum_{i=1}^{N_R} |h_{ix}|$.

Step 3: Sort the N_{TOTAL} elements in matrix $\tilde{\boldsymbol{H}}$ in descending order to form the vector,

$$\bar{\boldsymbol{H}} = [\bar{h}_1 \quad \bar{h}_2 \quad \dots \quad \bar{h}_{N_{TOTAL}}] \tag{18}$$

where $\bar{h}_1 \geq \bar{h}_2 \geq \cdots \geq \bar{h}_{N_{TOTAL}}$.

Step 4: Select the first N_{SA} elements of $\bar{\boldsymbol{H}}$ to form the vector

$$\acute{\boldsymbol{H}} = [\bar{h}_1 \quad \bar{h}_2 \quad \dots \quad \bar{h}_{N_{SA}}] \tag{19}$$

Step 5: Let $\bar{\bar{\boldsymbol{H}}} = [\bar{\bar{\boldsymbol{h}}}_1 \quad \bar{\bar{\boldsymbol{h}}}_2 \quad \dots \quad \bar{\bar{\boldsymbol{h}}}_{N_{SA}}]$ represent the matrix of N_{SA} channel gain vectors from $\boldsymbol{H}$ (Step 1) chosen and ordered according to the elements of matrix $\acute{\boldsymbol{H}}$ (Step 4). Generate N_{SA} combinations in 2 ways for the vectors of $\bar{\bar{\boldsymbol{H}}}$ using the binomial coefficient $\binom{N_{SA}}{2}$. We assume the pairs are of the form $(\bar{\bar{\boldsymbol{h}}}_y, \bar{\bar{\boldsymbol{h}}}_z)$.

Step 6: Compute the angle ϑ_n between the vectors $\bar{\bar{\boldsymbol{h}}}_y$ and $\bar{\bar{\boldsymbol{h}}}_z$

$$\vartheta_n = \cos^{-1}\left[\frac{\left\|\bar{\bar{\boldsymbol{h}}}_y^{\dagger}\bar{\bar{\boldsymbol{h}}}_z\right\|_F}{\left\|\bar{\bar{\boldsymbol{h}}}_y\right\|_F \left\|\bar{\bar{\boldsymbol{h}}}_z\right\|_F}\right] \tag{20}$$

Step 7: Arrange the n elements of $\boldsymbol{\vartheta}$ in ascending order.

Step 8: The vectors of the pair $(\boldsymbol{g}_y, \boldsymbol{g}_z)$ corresponding to the first element of $\boldsymbol{\vartheta}$ are compared in the following manner,

IF $\|\boldsymbol{g}_y\|_F < \|\boldsymbol{g}_z\|_F$ **THEN**
Discard antenna corresponding to $\boldsymbol{g}_y$
ELSE
Discard antenna corresponding to $\boldsymbol{g}_z$
ENDIF

Step 9: At this point there are $N_{SA} - 1$ channel gains. The remainder of the algorithm proceeds as follows,

IF $N_T = N_{SA} - 1$ **THEN**
Stop
ELSE
Repeat algorithm from Step 5
until $N_T = \acute{N}_{SA} - 1$
ENDIF

where $\acute{N}_{SA}$ is the new antenna subset size and replaces N_{SA}.

The final N_T antennas are then utilized as the antennas for transmission with SM.

B. CBM-SM

The second proposed algorithm will be referred to as CBM-SM. Once again, the configuration is denoted in the form $(N_{TOTAL}, N_{SA}, N_T, N_R)$. All steps are identical to the preceding scheme with the exception of Step 6 and 7. Step 6 and 7 for CBM-SM are as follows:

Step 6: For all combinations of the channel gain vectors find the correlation,

$$\rho_n = \left|\langle \bar{\bar{\boldsymbol{h}}}_y, \bar{\bar{\boldsymbol{h}}}_z \rangle\right| \tag{21}$$

where $\langle \boldsymbol{a}, \boldsymbol{b} \rangle$ represents the inner product between vectors $\boldsymbol{a}$ and $\boldsymbol{b}$.

Step 7: Arrange the n elements of $\boldsymbol{\rho}$ in descending order.

Step 8 then proceeds by using $\boldsymbol{\rho}$ instead of $\boldsymbol{\vartheta}$.

C. COMPLEXITY ANALYSIS FOR TAS-SM AND CBM-SM

It can be verified that the complexity of the TAS-SM selection algorithm in terms of real multiplications (we use real multiplications here to match the complexity analysis in [16], [17]) is given by,

$$\delta_{TAS-SM} = 2N_R N_{TOTAL} + 3 \sum_{\ell=0}^{N_{SA}-N_T-1} \binom{N_{SA} - \ell}{2} \tag{22}$$

Similarly, for the CBM-SM scheme it can be verified that the complexity is defined by,

$$\delta_{CBM-SM} = 2N_R N_{TOTAL} + 2 \sum_{\ell=0}^{N_{SA}-N_T-1} \binom{N_{SA} - \ell}{2} \tag{23}$$

From (22), (23) it is immediately evident that the proposed antenna selection schemes for SM have much lower complexity compared to the schemes proposed in [16] and [17].

V. NUMERICAL ANALYSIS

A. LOW-COMPLEXITY DETECTION OF SM

In this sub-section, the proposed SM-LCD method is compared to SM-OD [2] and SM-SVD [5]. Since SM-RC [8] closely matches the error performance of SM-OD, we do not include the simulation results for comparison. The lower bound for SM was also evaluated using Eq. (9) from [3], and has been included for comparison.

In the model, we have considered an uncoded SM system. QAM modulation has been used and two constellation sizes have been considered, viz. $M = 16, 64$. The channel considered is a quasi-static Rayleigh flat fading channel [2].

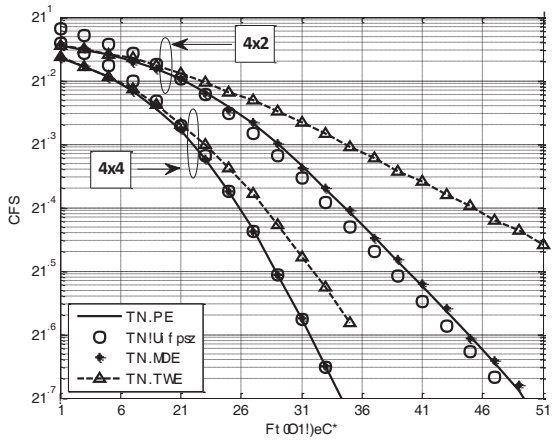

Figure 1: BER performance comparison of SM-OD including theoretical result, SM-LCD and SM-SVD for 16-QAM.

Figure 1 depicts the results for 16-QAM, where the 4×2 and the 4×4 antenna configurations have been considered. It is evident from the results, that the proposed SM-LCD technique matches the SM-OD almost perfectly for the 4×4 configuration, and for the 4×2 system there is less than 0.5 dB difference at a BER of 10^{-6}. These results have also been compared to SM-SVD. It is clear that the SM-SVD scheme only matches the SM-OD scheme at low SNRs (as is consistent with the discussion in [6]).

Figure 2 presents similar results for 64-QAM. Once again, it is evident that the proposed SM-LCD simulation curves match the SM-OD simulation results very closely.

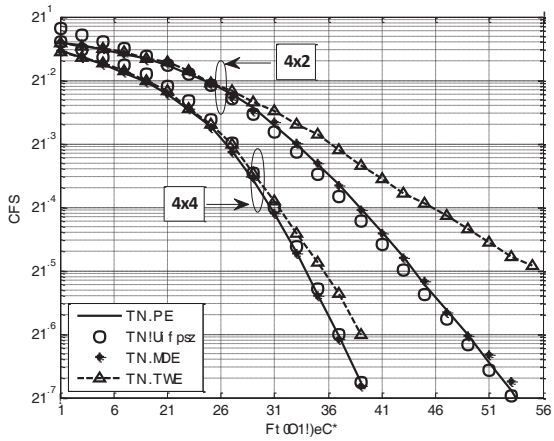

Figure 2: BER performance comparison of SM-OD including theoretical result, SM-LCD and SM-SVD for 64-QAM.

To further demonstrate the significance of the proposed SM-LCD scheme, the numerical complexity comparison of SM-LCD with SM-OD and SM-SVD is presented in Table 2 for several configurations.

Table 2: Comparison of computational complexity for SM-OD, SM-SVD and the proposed SM-LCD scheme

Configuration	Computational Complexity		
	SM-OD (10)	SM-SVD (11)	SM-LCD (Proposed) (15)
$N_T = 4, N_R = 2, M = 16$	100	58	28
$N_T = 4, N_R = 4, M = 16$	124	84	52
$N_T = 4, N_R = 2, M = 64$	340	154	28
$N_T = 4, N_R = 4, M = 64$	364	180	52

For $N_T = 4$, $N_R = 2$ and $M = 16$, SM-LCD yields a percentage drop of 72% compared to SM-OD and 52% compared to SM-SVD. When the number of receive antennas are increased to $N_R = 4$ for the same configuration this percentage drop becomes 58% and 38%, respectively. For $N_T = 4$, $N_R = 2$ and $M = 64$, SM-LCD yields a percentage drop of 92% and 82% compared to SM-OD and SM-SVD. Finally, when N_R is increased to 4 antennas for the same configuration, the percentage drops are 86% and 71%, respectively. Thus, it is clear that due to the computational complexity of the proposed SM-LCD scheme being independent of the constellation size, the imposed complexity is significantly reduced. Thus, we can conclude that the SM-LCD scheme can match the error performance of SM-OD very closely at a much lower computational complexity.

B. TRANSMIT ANTENNA SELECTION FOR SM

In this sub-section, the proposed multiple TAS-SM and CBM-SM schemes are evaluated. The analytical and simulation results for SM-OD [2], [3] are included for comparison. In addition, we have applied MG-TAS to SM. The curves have been included. In our model, we have considered a SM system with no coding. Optimal detection has been used at the receiver [2]. In each of the comparisons we have kept the primary number of antennas and the data rate the same. M-QAM modulation with Gray coded constellation is used. A quasi-static Rayleigh flat fading channel has been assumed. Figure 3 presents the results for data rates of a) 3 bits/transmission and b) 4 bits/transmission, for two receive antennas. The TAS-SM scheme has been compared to SM-OD [2] and the APM result for AWGN, similarly to [16], [17].

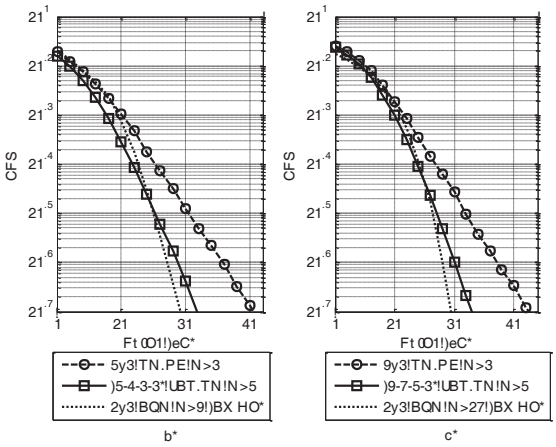

Figure 3: BER performance comparison for a) 3 bits/transmission and b) 4 bits/transmission for optimal SM, TAS-SM and APM maximum ratio receive combining in AWGN.

It is evident that for both data rates the TAS-SM scheme exhibits a significant improvement in the error rate. Example, at a BER of 10^{-6} an improvement of approximately 7.5 dB is realized in both instances and is of great significance.

Consider Figure 3 in [17]; comparing the scheme OH-SM (high complexity) to APM AWGN at a BER of 10^{-4} the OH-SM is approximately 1.6 dB worse for 4 bits/transmission and C-SM (complexity lower than OH-SM but still relatively high) is approximately 2 dB worse. In comparison to the same APM AWGN curves for 4bits/transmission the proposed scheme TAS-SM is approximately 0.5 dB worse. From this observation, we can conclude that the proposed TAS-SM offers an improved error performance compared to the closest competing schemes from [16] and [17].

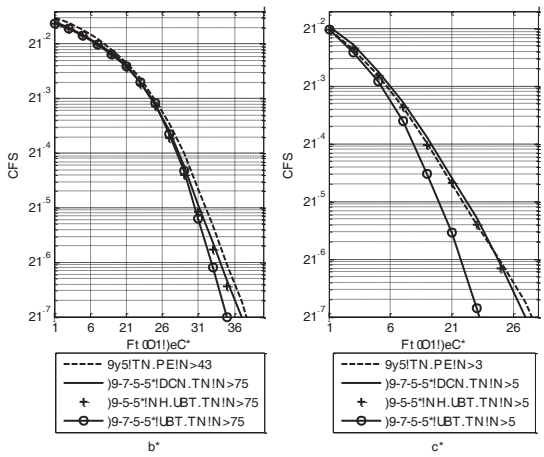

Figure 4: BER performance comparison for 4 bits/transmission and 8 bits/transmission for optimal SM, TAS-SM (8,6,4,4), CBM-SM (8,6,4,4) and MG-TAS-SM (8,4,4).

Figure 4 compares TAS-SM to optimal SM, CBM-SM and MG-TAS-SM for 4 and 8 bits/transmission and 4 receive antennas. For 4 bits/transmission, at a BER of 10^{-6} CBM-SM and MG-TAS-SM have a gain of approximately 0.5 dB compared to optimal SM, however TAS-SM has a gain of approximately 3.5 dB. For 8 bits/transmission and BER 10^{-6} CBM-SM exhibits approximately 1 dB improvement over optimal SM, slightly lower than MG-TAS-SM. On the other hand, as expected, TAS-SM exhibits an improvement of approximately 2.8 dB.

Thus, the proposed antenna selection scheme for SM can yield a significant improvement in the error performance.

VI. Conclusion

In this paper, we first proposed a low-complexity detection scheme for conventional SM. From comparisons with the existing schemes it is evident that the proposed SM-LCD scheme matches the optimal detector error performance very closely down to low BER. The receiver computational complexity analysis also shows that SM-LCD exhibits very low complexity, especially for large symbol constellation size, compared to the competing schemes, due to the complexity being independent of the symbol constellation size.

Secondly, two TAS schemes to improve the error performance of optimal SM were proposed. From comparisons with the existing schemes, CBM-SM has shown a small improvement in BER at a much lower complexity and TAS-SM has shown a significant improvement in BER at a much lower complexity.

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