Signed Quadrature Spatial Modulation for MIMO Systems

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Abstract—Quadrature spatial modulation (QSM) was recently proposed to increase the spectral efficiency of spatial modulation. In QSM, the real and imaginary parts of a symbol are transmitted independently from different antennas. A new spatial modulation technique, termed signed quadrature spatial modulation (SQSM), is introduced in this paper. Unlike QSM, SQSM creates four-dimensional spatial constellation, and hence results in higher system throughput. This technique extends the traditional spatial constellation dimension into ± in-phase and quadrature-phase dimensions. The ensuing constellation consists of four antenna indices, which are chosen independently in order to transmit the real and imaginary parts of the modulated symbols and their inverse. The performance of the proposed technique is compared to the most-recent SM techniques, namely, QSM, double SM, and improved QSM (IQSM). Numerical results demonstrate that SQSM provides significant performance gains (3-7dB), particularly when a large number of transmit antennas is used. This comes with a marginal increase in computational complexity compared to QSM and SM. Also, while IQSM can be competitive with SQSM in some cases, SQSM has a much lower computational complexity, e.g., 46% reduction in complexity in the case of 16 transmit antennas and 16 bps/Hz.

Index Terms—MIMO, quadrature spatial modulation, spatial modulation, spectral efficiency.

I. INTRODUCTION

In the last few years, there have been remarkable research efforts in developing multi-antenna transmission techniques with high spectral efficiency [1] or increased link reliability [2]. The vertical Bell labs layered space-time coding (V-BLAST) of [1] has a high spectral efficiency, but at the expense of high complexity and stringent restrictions on the number of receiver antennas needed to achieve minimal error rates [2].

Index modulation schemes, such as spatial modulation (SM), offer a trade-off between spectral efficiency and computational complexity. These schemes have higher spectral efficiency and lower decoding complexity compared to V-BLAST [3]. SM utilizes the indices of the transmit antennas to convey information bits in addition to those bits mapped with conventional M-ary amplitude phase modulation (APM) [4]. A special case of SM is space-time shift keying (STSK) [5], where the modulation is limited to the antennas’ indices and, hence, lower spectral efficiency is obtained. Different schemes of generalized spatial modulation (GSM) [6], [7] were reported in the literature to increase the spectral efficiency by allowing multiple transmit antennas to be simultaneously active during transmission.

Quadrature spatial modulation (QSM) is another novel technique that surpasses the performance of GSM [8]. In QSM, an M-ary APM symbol is decomposed into its real (in-phase, I) and imaginary (quadrature, Q) components, which are transmitted independently using selected antennas. The QSM technique has the same computational complexity of SM [8]. A generalized framework for QSM (GQSM) which can achieve higher spectral efficiency but at the cost of higher complexity is introduced in [9]. Also, a special case of GQSM, the improved quadrature spatial modulation (IQSM), is presented in [10]. In the latter scheme, the real and imaginary parts of two modulated symbols are transmitted simultaneously. The enhanced spatial modulation (ESM) scheme introduced in [11], [12] allows transmission from two independent antennas. These antennas transmit two different symbols which are drawn from basic and secondary symbol constellations. The secondary constellation is obtained by using single geometric interpolation. Accordingly, ESM is not a straightforwardly scalable scheme and is reported only for limited network configurations.

On the other hand, in double spatial modulation (DSM) [13] and complex quadrature spatial modulation (CQSM) [14], two symbols are transmitted from one/two independent antennas. The rotation angle between the symbol constellations is used to distinguish between two symbols if they are to be transmitted from the same indexed antenna. The minimum Euclidian distance between the two sets of the signal constellation is maximized by a simple optimization approach. However, this minimum distance becomes very small, and almost impractical for high-order modulation schemes. As for IQSM, a lookup table is needed for binary assignment, as the binary input cannot be divided and assigned to different dimensions.

In summary, most existing SM schemes use a single dimension of the spatial domain, e.g., SSK and SM, or a maximum of two dimensions, such as in DSM and QSM. Taking this feature as a starting point, the main contribution of this paper is to introduce a four-dimensional spatial space by using the sign of the I and Q components of the transmitted symbol.
This is expected to yield higher spectral efficiency even with zero or low APM modulation orders.

In fact, significant performance improvements can be achieved by the proposed signed quadrature spatial modulation (SQSM) technique, as shown analytically and numerically in the paper, particularly when a large number of transmit antennas is used. The gain of SQSM also comes with significant reductions in receiver-computational complexity compared to DSM and IQSM, or with a small cost compared to QSM. Furthermore, though IQSM can compete with the performance of SQSM when the number of transmit antennas is small, it is significantly more demanding in terms of receiver computational requirements. Moreover, the proposed SQSM has the same advantage of QSM in that a single radio-frequency (RF) chain is needed to produce the I/Q components of the symbol with an additional 180°-phase shifter. This would then produce the real and imaginary parts of the symbol and its inverse. On the other hand, spatial multiplexing systems (SMUX) such as V-BLAST require a dedicated RF chain for each transmit antenna while IQSM needs two RF chains at the transmitter. Moreover, transmitting the same data symbol from different antennas and the orthogonality between the I/Q components maintain a key advantage of SM, which is the avoidance of inter-channel interference (ICI) at the receiver. Nevertheless, transmit antenna synchronization is needed to avoid inter-symbol interference at the receiver.

The following content of the paper is organized as follows. In Section II, the description of the SQSM transmission technique is provided. Analysis of the bit-error probability performance is introduced in Section III. Section IV investigates the receiver complexity of SQSM in comparison with state-of-art schemes. Monte-Carlo simulation results and comparisons are provided in Section V, and the paper is concluded in Section VI.

II. SIGNED QUADRATURE SPATIAL MODULATION

Consider a multiple-input multiple-output (MIMO) communication system with \( N_t \) transmit antennas and \( N_r \) receive antennas. In general, the proposed SQSM scheme transmits a symbol \( x \) and its inverse \(-x\) within the same time period. Symbol \( x \) is drawn from the first quadrant of an \( M \)-ary APM. The real and imaginary parts of \( x \) and of its inverse \(-x\) are transmitted from different antennas.

Specifically, the incoming information bits are divided into five parts: \( \log_2 M \) bits are modulated to determine the signal constellation symbol \( x \), and four parts of length \( \log_2 N_t \) bits each are used to select the individual four transmit antennas. The selected transmit antennas are used to send the real and the imaginary components of \( x \) and \(-x\). It should be noted that \( \sqrt{2} x R \) is sent if the indices of the antennas chosen to transmit the real parts of symbols \( xR \) and \(-xR\) are identical. A similar concept is applied when the selected indices for the imaginary parts of symbols \( jx\alpha \) and \(-jx\alpha\) are identical, in which case \( \sqrt{2} jx\alpha \) is sent. Illustration of the SQSM concept with QPSK and \( N_t = 4 \) transmit antennas is shown in Fig. 1.

The received signal can be expressed as,

\[
y = Hz + n, \tag{1}
\]

where \( y \) is the \( N_r \times 1 \) receive signal vector, \( z \) is the \( N_t \times 1 \) transmit signal vector, \( n \) is the \( N_t \times 1 \) white noise vector, \( H \) is the \( N_t \times N_t \) channel matrix. Given that \( \sigma_n^2 \) is the noise power, the entries of \( H \) and \( n \) are drawn from complex Gaussian distributions \( \mathcal{CN}(0,1) \) and \( \mathcal{CN}(0,\sigma_n^2) \), respectively.

Assuming that \( l_{R_R}, l_{R_A}, l_{A_A} \) and \( l_{A_R} \) are the indices of the activated antennas used to transmit \( x_R, -x_R, jx\alpha \), and \(-jx\alpha\), respectively, the received signal of (1) can be equivalently expressed as

\[
y = a(x, l_{R_R}, l_{R_A}, l_{A_A}, l_{A_R}) + n, \tag{2}
\]

where

\[
a(x, l_{R_R}, l_{R_A}, l_{A_A}, l_{A_R}) = \begin{cases} x_R (\mathbf{h}_{l_{R_R}} - \mathbf{h}_{l_{R_A}}) + j x_\alpha (\mathbf{h}_{l_{A_A}} - \mathbf{h}_{l_{A_R}}) & \text{if } l_{R_R} \neq l_{R_A} \text{ and } l_{A_A} \neq l_{A_R} \\ \sqrt{2} \mathbf{h}_{l_{R_R}} x_R + j x_\alpha (\mathbf{h}_{l_{A_A}} - \mathbf{h}_{l_{A_R}}) & \text{if } l_{R_R} = l_{R_A} \text{ and } l_{A_A} \neq l_{A_R} \\ x_R (\mathbf{h}_{l_{R_R}} - \mathbf{h}_{l_{R_A}}) + \sqrt{2} \mathbf{h}_{l_{A_A}} x_\alpha & \text{if } l_{R_R} \neq l_{R_A} \text{ and } l_{A_R} = l_{A_A} \\ \sqrt{2} (\mathbf{h}_{l_{R_R}} x_R + j \mathbf{h}_{l_{A_A}} x_\alpha) & \text{if } l_{R_R} = l_{R_A} \text{ and } l_{A_R} = l_{A_A}
\end{cases} \tag{3}
\]

in which \( \mathbf{h}_{l_{R_R}} \) denotes the \( l_{R_R} \)th column of \( H \), i.e., \( [\mathbf{h}_{l_{R_R}} = \mathbf{h}_{l_{R_R}}, \ldots, \mathbf{h}_{l_{R_R}}, \ldots] \). It is assumed that the variance of the transmitted signal vector is normalized to unity, or

\[
\mathbf{E}(a(x, l_{R_R}, l_{R_A}, l_{A_A}, l_{A_R}) a^*(x, l_{R_R}, l_{R_A}, l_{A_A}, l_{A_R})) = 1, \tag{4}
\]

where \( \mathbf{E}(\cdot) \) is the expected value function. Therefore, the signal-to-noise ratio (SNR) is defined as \( \text{SNR} = 1 / \sigma_n^2 \).

For illustration purposes, a \( 4 \times 4 \) MIMO system with 8-QAM modulation is considered. The bit rate for each channel used is 9 bits/s. The first bit is used to choose between the two symbols of the 8-QAM’s first quadrant: 0 for \( 1+j \) and 1 for \( 1+3j \). The remaining four 2-bits are used to index the real and imaginary parts on antennas. For example, assume that
the incoming information bits are $[01001110]$. Thus, $x = 1 + j$, and its inverse is $-x = -1 - j$. The $x_R = 1$ will be transmitted from the $3^{rd}$ antenna, $jx_Q = j$ from the $2^{nd}$ antenna, $-x_R = -1$ from the $4^{th}$ antenna, and $-jx_Q = -j$ from the $1^{st}$ antenna. The transmitted vector is then given by $z = [-j, j, 1, -1]$. In the same manner, say $[101011110]$ is the input information, then the transmit vector will be $z = [0, \sqrt{2}, -j, 3, j, 3]$. The maximum-likelihood (ML) detection at the SQSM receiver consists in finding the estimates $[\tilde{x}, \tilde{l}_R, \tilde{l}_I, \tilde{l}_Q, \tilde{l}_L, \tilde{l}_N]$ according to,

$$
\begin{align*}
\arg \min_{x,l_R,l_I,l_Q,l_L,l_N} ||y - a(x,l_R,l_I,l_Q,l_L,l_N)||^2.
\end{align*}
$$

where $a$ represents the channel estimation error, modeled by a complex Gaussian variable with zero mean and variance $\sigma^2$. Let $a = h_{R,v}x_R - h_{I,v}x_I + jh_{I,v}x_Q - jh_{R,v}x_Q$ and $\tilde{a} = \tilde{h}_{R,v}\tilde{x}_R - \tilde{h}_{I,v}\tilde{x}_I + j\tilde{h}_{I,v}\tilde{x}_Q - j\tilde{h}_{R,v}\tilde{x}_Q$ be two distinct codewords assuming that $a$ is estimated at the receiver given that $a$ is transmitted. Then, the pairwise error probability (PEP) is given by,

$$
\begin{align*}
P_{pe}[a \rightarrow \tilde{a} | H] &= Q\left(\sqrt{\frac{||a - \tilde{a}||^2}{2\sigma^2}}\right) \\
&= Q\left(\sqrt{\tilde{\zeta}}\right),
\end{align*}
$$

where $Q(.)$ is the Q-function, and

$$
\tilde{\zeta} = \frac{1}{2\sigma^2}(a - \tilde{a})^H(a - \tilde{a}),
$$

where $(.)^H$ denotes the Hermitian operator. The PEP in [7] can be easily generalized to include all cases of transmitted vectors defined in [8]. The average PEP can be written as [15].

$$
\tilde{P}_e(a \rightarrow \tilde{a}) = \frac{1}{2} \left(1 - \sqrt{\frac{\tilde{\zeta}^2}{1 + \tilde{\zeta}^2}}\right),
$$

(9)

where $\tilde{\zeta}$ is the expected value of $\zeta$, given by

$$
\tilde{\zeta} = \frac{1}{2\sigma^2 + 4\sigma^2} \left(\sum_{k=1}^{4} D_k \right),
$$

(10)

where $D_1 = (|x_{R,v}|^2 + |\tilde{x}_{R,v}|^2) A_1 + (|x_{Q,v}|^2 + |\tilde{x}_{Q,v}|^2) B_1$, and $D_2 = |x_{R,v} - \tilde{x}_{R,v}|^2 A_2 + |x_{Q,v} - \tilde{x}_{Q,v}|^2 B_2$, with

$A_1 = 0, A_2 = 1$ 
$B_1 = 0, B_2 = 1$

$D_3$ and $D_4$ are the same as $D_1$ and $D_2$ after replacing the subscript $v$ with $n$ in all symbols. The bit error rate (BER) of the SQSM system is upper bounded (asymptotically tight bound) by the following average bit-error probability (ABEP),

$$
P_b = \frac{2}{m2^m} \sum_{i=1}^{2^m} \sum_{k=1}^{2^m} \tilde{P}_e(a \rightarrow \tilde{a}) e_{i,k},
$$

(11)

where $m$ is the spectral efficiency given in bps/Hz and $e_{i,k}$ is the number of bit errors related with $P_e(a \rightarrow \tilde{a})$. In the ideal case of perfect channel estimation (i.e. $\sigma^2 = 0$), $\tilde{\zeta}$ (10) reduces to

$$
\tilde{\zeta} = \frac{1}{2\sigma^2} \sum_{k=1}^{4} D_k.
$$

With $N_r$ receive antennas, the PEP is given by

$$
P_{pe}[a \rightarrow \tilde{a} | H] = Q\left(\sqrt{\frac{N_r}{\sum_{i=1}^{N_r} \zeta_i}}\right)
$$

(13)

which results in $Z$ being a chi-squared random variable with a probability density function (PDF) given by

$$
P_Z(Z) = \frac{1}{\Gamma(N_r)\zeta^{N_r-1}} Z^{N_r-1} \exp \left(-\frac{Z}{\zeta}\right),
$$

(14)
where $\Gamma(\cdot)$ denotes the Gamma function. The average unconditional PEP is given by [16], [17],

$$
\bar{P}_e (\mathbf{a} \rightarrow \hat{\mathbf{a}}) = \gamma^{N_r} \sum_{k=0}^{N_r-1} \left( \frac{N_r - 1 + k}{k} \right) [1 - \gamma]^k,
$$

(15)

where $\gamma = \frac{1}{2} \left( 1 - \sqrt{\frac{\zeta}{1 + \zeta}} \right)$. The PDF of $Z$ around the origin can be approximated using Taylor series as [18],

$$
P_Z(Z) = \frac{1}{\Gamma(N_r)\zeta^{N_r}} Z^{N_r-1} \left( 1 + O \right),
$$

(16)

where $O$ represents the ignored higher terms. The average PEP is computed as [18],

$$
\bar{P}_e (\mathbf{a} \rightarrow \hat{\mathbf{a}}) = \int_0^\infty Q(\sqrt{Z}) \frac{1}{\Gamma(N_r)\zeta^{N_r}} Z^{N_r-1} dZ
$$

$$
\approx \frac{2^{N_r-1} (N_r + 0.5)}{\sqrt{\pi} (N_r)!} \left( \frac{1}{\zeta} \right)^{N_r}.
$$

(17)

It is clear that the SQSM system can achieve a diversity gain of $N_r$.

**IV. RECEIVER COMPLEXITY ANALYSIS**

The computational complexity of SQSM is analyzed here. It equals the total result of the number of real-addition operations and the number of real-multiplication operations multiplied by the number of ML searches. The detection process in (5) requires $4N_t^4 N_r$ real multiplications, $4N_t^2 (N_r - 1)$ average real additions, and ML searches over $M/4$-dimensional modulation space.

For comparison purpose, the complexity of recent SM techniques is shown in Table I. In Fig. 2, the computational complexities of all systems under consideration are compared using $16 \times 16$ MIMO configuration and achieving $m = 16$ bps/Hz (with different constellation size $M$ for each scheme to guarantee a fair comparison).

As observed from Table I, the SQSM scheme offers an extra-considerable complexity reduction compared to SMUX and IQSM systems. This can be easily seen from Fig. 2. Clearly, there is a complexity reduction of $44\%$ compared to IQSM. Although the SQSM scheme requires an acceptable increase in complexity ($\leq 24\%$) compared to SM and QSM, it provides significant BER improvement as will be seen in the next section. Moreover, for large-scale MIMO systems, suboptimal low decoding complexity schemes can be modified to suit the SQSM [19], [20], [21].

**V. PERFORMANCE COMPARISONS**

This section provides simulation results to evaluate the performance of the proposed SQSM, in terms of BER, which is a function of the received SNR per antenna. Both analytical (upper bound) and Monte-Carlo numerical results are shown for different spectral efficiency values. Results show a close-match between the analysis and simulations for a wide and pragmatic range of SNR. Specifically, SQSM is compared to the most recent state-of-art spatial modulation schemes,
namely, QSM, IQSM and DSM. It should be noted that the ESM scheme of [10] is ignored in the comparisons due to its lack of simple scalability. In contrast, DSM is adequate for comparison due to its superior BER performance compared to ESM [11]. Simulations were conducted assuming Rayleigh fading channels. Perfect channel estimation at the receivers is assumed unless otherwise stated.

In Fig. 3, the performance of SQSM is further analyzed for a 16 × 16 MIMO system with a rate of 16 bps/Hz. Significant enhancement is achieved over QSM, DSM and SMUX, where gains of 14.55 dB, 9.1 dB, and 1.05 dB are respectively obtainable at a BER of 10⁻⁴. It is worth to note that the SQSM technique starts to have a better performance than IQSM while reducing the complexity by 44%.

In order to show SQSM effectiveness using different modulation schemes, Fig. 4 shows the performance of 4 × 4 SQSM system with data rates of 8, 9, and 10 bps/Hz, using PSK and QAM constellations. It is worth noting that, since SQSM works with non-zero real and imaginary components of constellation points, the M-PSK constellation is rotated with \( \pi \times M \). At a BER of 10⁻⁴, the gain losses of using PSK over QAM are 0.3 dB and 3 dB at data rates of 9 bps/Hz and 10 bps/Hz, respectively. This can be easily justified by the fact that the minimum Euclidian distance of PSK constellation is smaller than its QAM counterpart.

In Fig. 6, the impact of imperfect channel knowledge at the receiver for SQSM and QSM is evaluated and compared to the case of perfect channel knowledge of 4 × 4 system for 10 bps/Hz. Two values of the channel estimation error are considered, namely, the error variance is proportional to 1/SNR and when it is fixed to -15 dB. Analytical results for SQSM are also depicted and compared to simulation results. The analytical results shows a good match with the simulation ones at high SNR. In case of \( \sigma_e^2 = 1/SNR \), a degradation of 2.3 dB can noticed for SQSM and 3 dB for QSM as in [15], while SQSM has the superiority in performance with gain of 2.3 dB at a BER of 10⁻⁴. When \( \sigma_e^2 = -15 \text{ dB} \), the error probability at high SNR is fixed to \( 1.5 \times 10^{-3} \) for SQSM and \( 4 \times 10^{-3} \) for QSM.

The results demonstrate that SQSM is better suited to MIMO technology compared to the most recent SM schemes (QSM, IQSM and DSM) in terms of both performance and computational complexity. The results show that SQSM has stronger immunity to channel uncertainties compare to QSM [15] while at the same time giving significant enhancement in performance.

VI. CONCLUSION

This paper introduced a new spatial modulation scheme that uses APM symbol from the positive quadrant and its inverse.
This was done by splitting a symbol and its inverse into in-phase and quadrature-phase components; each transmitted from an independently selected antenna. This design allows for four-dimensional spatial constellations, unlike conventional SM, QSM and DSM. Moreover, SQSM has a straightforward scalability, unlike ESM. Therefore, SQSM has a higher spatial efficiency compared to QSM under the same number of transmit antennas.

Performance evaluation of the proposed SQSM scheme was carried out using analysis and simulations. In particular, it was shown that SQSM has superior performance compared to QSM and DSM. Also, SQSM has a much lower decoding complexity than IQSM (56% in the setting corresponding to Fig. 2), while at the same time achieving a performance enhancement over IQSM when used as a signed quadrature shift keying ($M = 4$).

REFERENCES


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