A High-Rate Full-Diversity 2x2 Space-Time Code with Simple Maximum Likelihood Decoding

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Abstract— *Multiple-input multiple-output (MIMO) techniques have become an essential part of broadband wireless communications systems. For example, the recently developed IEEE 802.16e specifications for broadband mobile wireless access include three MIMO profiles employing 2×2 space-time codes (STCs) and two of those, namely, Alamouti's STC for transmit diversity, and the 2x2 spatial multiplexing (SM) scheme, are mandatory on the downlink of Mobile WiMAX systems. The first of these has full diversity but it is only half rate, while the second is full rate but suffers from diversity loss. In this paper, we develop a full-diversity STC of size 2×2, which represents an interesting trade-off between these two schemes. More specifically, the presented code is of rate 3/4, it has full diversity, and it leads to simple maximum-likelihood (ML) detection whose complexity grows only linearly with the size of the signal constellation. This makes it a suitable candidate for future evolutions of broadband wireless systems toward higher data rates and performance.*

Keywords— *MIMO systems, maximum likelihood detection, space-time codes, Mobile WiMAX.*

I. INTRODUCTION

Multiple-input multiple-output (MIMO) techniques based on using multiple antennas at both transmitter and receiver can provide spatial diversity, multiplexing gain, interference suppression, and make various tradeoffs between them. These techniques have been incorporated in all of the recently developed wireless communications system specifications including the IEEE 802.11n standard for local area networks and the IEEE 802.16e-2005 standard [1] for mobile broadband wireless access.

From the MIMO schemes included in the IEEE 802.16e specifications, the WiMAX Forum has specified two mandatory profiles for use on the downlink. The first of them, referred to as Matrix A, is based on the space–time code (STC) proposed by Alamouti for transmit diversity [2]. This code provides perfect second-order diversity when used with a single receive antenna and fourth-order diversity when used with two antennas at the receiver. But it is only half-rate, because it only transmits two symbols using two time slots and two transmit antennas. The second profile, referred to as Matrix B, is spatial multiplexing (SM), which uses two transmit antennas to transmit two independent data streams. This scheme is full-rate, but it does not benefit from any spatial diversity at the transmitter.

For future evolutions of the WiMAX standard, it is highly desirable to include new codes combining the respective advantages of the Alamouti code and the SM while avoiding their drawbacks. Such a code, referred to as Matrix C, actually exists in the IEEE 802.16e specifications. This code is a simple variant of the Golden code [3], which is known to be one of the best 2×2 space-time codes, but the problem of this code is its detection complexity, which grows with the fourth-power of the signal constellation size. To alleviate the detection complexity problem, a new 2×2 STC was recently proposed in [4]. This 2×2 STC is both full-rate, full-diversity, and also has the interesting property that its optimum detection complexity grows only quadratically with the size of the signal constellation. In this paper, we present a subclass of this code which provides a further trade-off between rate and detection complexity.

The rest of the paper is organized as follows. First, in the next section, we review Alamouti's STC and SM, and briefly recall the general design criteria for STCs. Next, in Section III, we describe the proposed STC and discuss its advantages. Finally, we present some performance results in Section IV, and give our conclusions in Section V.

II. BACKGROUND

A. Alamouti's STC

The first multiple antenna profile included in the IEEE 802.16e-2005 specifications is the simple STC proposed by Alamouti [2] for transmit diversity. Originally, Alamouti's STC was proposed to avoid the use of receive diversity on the downlink and keep the subscriber stations simple. In OFDMA-based WiMAX systems, this technique is applied subcarrier by subcarrier and can be described as follows:

Suppose that (s_1, s_2) represents a group of two consecutive symbols in the input data stream to be transmitted. During a first symbol period t_1, transmit (Tx) antenna 1 transmits symbol s_1 and Tx antenna 2 transmits symbol s_2. Next, during the second symbol period t_2, Tx antenna 1 transmits symbol s_2^* and Tx antenna 2 transmits symbol $-s_1^*$. Denoting the channel response from Tx1 to the receiver (Rx) by h_1 and the channel response from Tx2 to the Rx by h_2, the received signal samples corresponding to symbol periods t_1 and t_2 can be respectively written as:

$$r_1 = h_1 s_1 + h_2 s_2 + n_1 \,, \quad (1.a)$$
$$r_2 = h_1 s_2^* - h_2 s_1^* + n_2 \,, \quad (1.b)$$

where the n_i's are independent circularly symmetric additive Gaussian noise terms having a spectral density N_0.

The receiver computes the following signals to estimate the symbols s_1 and s_2:

$$z_1 = h_1^* r_1 - h_2 r_2^* = \left(|h_1|^2 + |h_2|^2\right) s_1 + h_1^* n_1 - h_2 n_2^* \,, \quad (2.a)$$
$$z_2 = h_2^* r_1 + h_1 r_2^* = \left(|h_1|^2 + |h_2|^2\right) s_2 + h_2^* n_1 + h_1 n_2^* \,. \quad (2.b)$$

These expressions clearly show that z_1 (resp. z_2) can be sent to a threshold detector to estimate symbol s_1 (resp. symbol s_2) without interference from the other symbol. Moreover, since the useful signal coefficient is the sum of the squared moduli of two independent fading channels, these estimations benefit from perfect second-order diversity, which is equivalent to Rx diversity with maximum-ratio combining (MRC).

Alamouti's transmit diversity can also be combined with MRC when $N_R > 1$ antennas are used at the Rx. In the multiple Rx antenna case, it can be easily seen that this MIMO scheme has a diversity order of $2N_R$. In other words, Alamouti's STC achieves the maximum available diversity with a simple Rx structure, but it does not give any spatial multiplexing gain. Indeed, if we define the rate as the number of symbols transmitted per antenna use, this MIMO scheme leads to a transmission rate of 1/2 which counterbalances its other attractive features.

B. *Spatial Multiplexing*

The second MIMO profile included in WiMAX system specifications is the SM based on the so-called matrix **B** = $[s_1 \; s_2]^T$. This system is full-rate, as it transmits one symbol per antenna use. But it does not offer any diversity gain from the Tx side, because each symbol is transmitted from one antenna only. In a 2×2 MIMO system, it offers a diversity gain of 2 on the receiver side provided it uses maximum-likelihood (ML) detection [5].

To describe the SM, we omit the time and frequency dimensions, leaving only the space dimension. The symbols transmitted by Tx1 and Tx2 in parallel are denoted as s_1 and s_2, respectively. Denoting by h_{ji} the channel response from Tx i to Rx j (i, j = 1, 2), the signals received by the two Rx antennas are given by

$$r_1 = h_{11} s_1 + h_{12} s_2 + n_1 \quad (3.a)$$
$$r_2 = h_{21} s_1 + h_{22} s_2 + n_2 \quad (3.b)$$

which can be written in matrix form as:

$$\begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}. \quad (4)$$

The ML detector makes an exhaustive search over all possible values of the transmitted symbols and decides in favor of (s_1, s_2) which minimizes the Euclidean distance:

$$D(s_1, s_2) = \left\{ |r_1 - h_{11}s_1 - h_{12}s_2|^2 + |r_2 - h_{21}s_1 - h_{22}s_2|^2 \right\}. \quad (5)$$

As an alternative to Alamouti's STC and the SM, we present in this paper a full-diversity 2×2 STC which provides an interesting trade-off between rate and complexity.

C. *STC Design Criteria*

We now give a brief discussion on the most common design criteria for STCs when the transmitter does not have any channel state information. For a MIMO transmission with 2 Tx antennas, we take

$$\mathbf{X} = \begin{bmatrix} x_1 & x_3 \\ x_2 & x_4 \end{bmatrix}, \quad (6)$$

as the 2×2 codeword matrix whose elements take values from the codebook $\mathcal{X}$. Code optimization generally relies on the analysis of pairwise error probability (PEP) $P(\mathbf{X} \to \hat{\mathbf{X}})$, which is the probability that $\hat{\mathbf{X}}$ is detected while $\mathbf{X}$ is transmitted. The goal is to minimize the error probability, but the analysis of exact error performance is difficult, and therefore one resorts to the union bound

$$P_s \le \frac{1}{|\mathcal{X}|} \sum_{\mathbf{X} \in \mathcal{X}} \sum_{\hat{\mathbf{X}} \neq \mathbf{X}} P(\mathbf{X} \to \hat{\mathbf{X}}) \,, \quad (7)$$

where $|\mathcal{X}|$ denotes the cardinality of the codebook $\mathcal{X}$. A Chernoff bound analysis of the PEP was performed in [6], and this analysis led to the following design criteria:

1) Rank Criterion: In order to achieve maximum diversity, the diversity gain

$$d(\mathcal{X}) = \min_{\substack{\mathbf{X}, \hat{\mathbf{X}} \in \mathcal{X} \\ \mathbf{X} \neq \hat{\mathbf{X}}}} \operatorname{rank}\left[(\mathbf{X} - \hat{\mathbf{X}})(\mathbf{X} - \hat{\mathbf{X}})^H \right] \quad (8)$$

should be maximized. If $(\mathbf{X} - \hat{\mathbf{X}})(\mathbf{X} - \hat{\mathbf{X}})^H$ is full rank for all code word pairs, then the code is said to have *full diversity*.

2) Determinant Criterion: The coding gain is defined as

$$\delta = \min_{\substack{\mathbf{X}, \hat{\mathbf{X}} \in \mathcal{X} \\ \mathbf{X} \neq \hat{\mathbf{X}}}} \det\left[(\mathbf{X} - \hat{\mathbf{X}})(\mathbf{X} - \hat{\mathbf{X}})^H \right]. \quad (9)$$

In order to obtain the best performance, the coding gain should be maximized for a given average transmit power.

It should be noted that, for high values of the signal-to-noise ratio (SNR), the most important parameter is the diversity gain which dominates the steepness of the bit-error rate (BER) curve. Afterwards, it is the coding gain which should be maximized. In the sequel, we will directly optimize the proposed code such that the coding gain is maximized. In light of the above design criteria, this directly ensures that the optimized code will have full diversity.

III. RATE-3/4 2x2 STC WITH SIMPLE ML DECODING

In [4], the following full-rate 2×2 STC was proposed and shown that it maximizes the diversity gain with a high coding gain, while leading to an optimum detector whose complexity grows only quadratically with the signal constellation size. In this STC, a group of 4 data symbols (s_1, s_2, s_3, s_4) is transmitted as

$$\mathbf{X}_{full} = \begin{bmatrix} as_1 + bs_3 & -cs_2^* - ds_4^* \\ as_2 + bs_4 & cs_1^* + ds_3^* \end{bmatrix}, \tag{10}$$

where *a*, *b*, *c*, and *d* are complex-valued design parameters to be optimized and $(\cdot)^*$ designates complex conjugate. In this matrix representation, the first column represents the symbol combinations transmitted during a first symbol interval t_1 and the second column represents the symbol combinations transmitted during a second symbol interval t_2. The first row of the matrix gives the symbol combinations transmitted from the first Tx antenna, and second row of the matrix gives the symbol combinations transmitted from the second Tx antenna. As shown in [4], the *a, b, c, d* parameters must have the same magnitude to obtain the desired properties. Furthermore, without any loss of generality, the parameters *a* and *c* can be constrained to have the same value in order to decrease the number of unknown parameters. Then, the complex-valued parameters *b* and *d* can be optimized such that the resulting STC has a large coding gain (equivalently it has full diversity).

For the optimized parameters, the full-rate 2×2 STC given in (10) is a full-diversity code whose optimum Rx has a complexity that is *only* proportional to M^2, where M is the size of the signal constellation. On the other hand, the optimum Rx for the existing full-rate full-diversity STCs has a complexity proportional to M^4. Therefore, (10) leads to a reduction in the number of Euclidean distance computations by a factor of $16^2 = 256$ (resp. $64^2 = 4{,}096$) for the 16-QAM (resp. 64-QAM) signal constellation. Consequently, this code makes the implementation of full-rate, full-diversity 2×2 STCs with optimum receiver more realistic.

We now modify the STC given in (10) for a further reduction of the ML decoder complexity. More specificallly, we set $s_4 = s_3$ in the coding matrix and scale the power of this symbol. This leads to the following 2×2 code of rate 3/4:

$$\mathbf{X}_{3/4} = \begin{bmatrix} as_1 + bs_3/\sqrt{2} & -(cs_2^* + ds_3^*/\sqrt{2}) \\ as_2 + bs_3/\sqrt{2} & cs_1^* + ds_3^*/\sqrt{2} \end{bmatrix}. \tag{11}$$

For the sake of simplicity, we only consider the 2 Rx antenna case to describe the detector. The results can be easily extended to any number of Rx antennas.

On the first Rx antenna, let the two signals received at the first and second symbol intervals be

$$r_1 = h_{11}\left(as_1 + bs_3/\sqrt{2}\right) + h_{12}\left(as_2 + bs_3/\sqrt{2}\right) + n_1 \tag{12.a}$$

$$r_2 = h_{11}\left(-cs_2^* - ds_3^*/\sqrt{2}\right) + h_{12}\left(cs_1^* + ds_3^*/\sqrt{2}\right) + n_2\,. \tag{12.b}$$

Similarly, on the second Rx antenna we have

$$r_3 = h_{21}\left(as_1 + bs_3/\sqrt{2}\right) + h_{22}\left(as_2 + bs_3/\sqrt{2}\right) + n_3 \tag{13.a}$$

$$r_4 = h_{21}\left(-cs_2^* - ds_3^*/\sqrt{2}\right) + h_{22}\left(cs_1^* + ds_3^*/\sqrt{2}\right) + n_4\,, \tag{13.b}$$

where the n_i 's are circularly symmetric additive Gaussian noise terms with spectral density N_0.

The ML detector makes an exhaustive search over all possible values of the transmitted symbols and decides in favor of the symbol triplet (s_1, s_2, s_3), which minimizes the Euclidean distance denoted by $D(s_1, s_2, s_3)$. Specifically, this exhaustive search involves the computation of M^3 metrics and M^3-1 comparisons, which is excessive for the 16-QAM and 64-QAM signal constellations. Now, we show that the particular structure of the proposed code leads to an ML detector whose complexity grows only linearly with the constellation size.

From the received signal samples (r_1, r_2, r_3, r_4), let us compute the following intermediate signals:

$$z_1 = r_1 - b(h_{11} + h_{12})s_3/\sqrt{2} \tag{14.a}$$

$$z_2 = r_2 - d(h_{11} + h_{12})s_3^*/\sqrt{2} \tag{14.b}$$

$$z_3 = r_3 - b(h_{21} + h_{22})s_3/\sqrt{2} \tag{14.c}$$

$$z_4 = r_4 - d(h_{21} + h_{22})s_3^*/\sqrt{2} \tag{14.d}$$

for a given value of the symbol s_3. Then, from (12) – (14), we have

$$z_1 = a\left(h_{11}s_1 + h_{12}s_2\right) + n_1 \tag{15.a}$$

$$z_2 = c\left(h_{12}s_1^* - h_{11}s_2^*\right) + n_2 \tag{15.b}$$

$$z_3 = a\left(h_{21}s_1 + h_{22}s_2\right) + n_3 \tag{15.c}$$

$$z_4 = c\left(h_{22}s_1^* - h_{21}s_2^*\right) + n_4 \tag{15.d}$$

From these intermediate signals, we next compute:

$$h_{11}^* z_1 = a\left(|h_{11}|^2 s_1 + h_{11}^* h_{12} s_2\right) + h_{11}^* n_1 \tag{16.a}$$

$$h_{12} z_2^* = c^* \left(|h_{12}|^2 s_1 - h_{11}^* h_{12} s_2 \right) + h_{12} n_2^* \quad (16.b)$$

$$h_{21}^* z_3 = a \left(|h_{21}|^2 s_1 + h_{21}^* h_{22} s_2 \right) + h_{21}^* n_3 \quad (16.c)$$

$$h_{22} z_4^* = c^* \left(|h_{22}|^2 s_1 - h_{21}^* h_{22} s_2 \right) + h_{22} n_4^* . \quad (16.d)$$

Finally, we compute the signal y_1 as:

$$\begin{aligned} y_1 &= \left(h_{11}^* z_1 + h_{21}^* z_3 \right) / a + \left(h_{12} z_2^* + h_{22} z_4^* \right) / c^* \\ &= \left(|h_{11}|^2 + |h_{12}|^2 + |h_{21}|^2 + |h_{22}|^2 \right) s_1 + w_1 \end{aligned} \quad (17)$$

with $w_1 = \left(h_{11}^* n_1 + h_{21}^* n_3 \right) / a + \left(h_{12} n_2^* + h_{22} n_4^* \right) / c^*$.

Here, the signal y_1 has no terms involving symbol s_2 and the coefficient of the term in s_1 clearly indicates that estimation of s_1 benefits from full fourth-order spatial diversity. By sending this signal to a threshold detector, we get the ML estimate of symbol s_1 conditional on s_3. Note that, similar to the full-rate case (see [4] for more detail), the elimination of the terms involving s_2 can be possible if and only if the coefficients of the symbols s_1 and s_2 are identical in each column of the code matrix.

Similar to (16.a) – (16.d), we also compute $h_{12}^* z_1$, $h_{11} z_2^*$, $h_{22}^* z_3$, $h_{21} z_4^*$, and then, we obtain

$$\begin{aligned} y_2 &= \left(h_{12}^* z_1 + h_{22}^* z_3 \right) / a - \left(h_{11} z_2^* + h_{21} z_4^* \right) / c^* \\ &= \left(|h_{11}|^2 + |h_{12}|^2 + |h_{21}|^2 + |h_{22}|^2 \right) s_2 + w_2 \end{aligned} \quad (18)$$

with $w_2 = \left(h_{12}^* n_1 + h_{22}^* n_3 \right) / a - \left(h_{11} n_2^* + h_{21} n_4^* \right) / c^*$.

As previously, signal y_2 has no terms involving symbol s_1, and the coefficient of the term in s_2 shows that estimation of s_2 benefits from full fourth-order spatial diversity. By sending y_2 to a threshold detector, we get the ML estimate of symbol s_2 conditional on s_3. ML estimation of s_1 and s_2 conditional on s_3 is illustrated in Fig. 1.

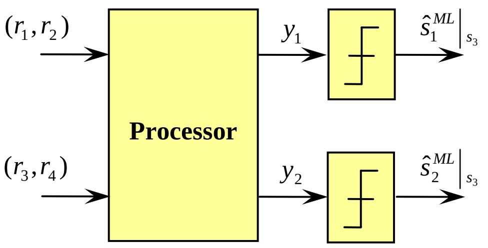

Fig. 1. Processing of the received signals to determine the ML estimate of symbols s_1 and s_2 conditional on a particular value of symbol s_3.

In this way, for a given value of symbol s_3, we get the ML estimate of (s_1, s_2), which we denote by (s_1^{ML}, s_2^{ML}). Now, instead of computing the ML metric $D(s_1, s_2, s_3)$ for all (s_1, s_2, s_3) values, we only need to compute it for $(s_1^{ML}, s_2^{ML}, s_3)$. In other words, the optimum Rx computes the metric $D(s_1, s_2, s_3)$ for $(s_1^{ML}, s_2^{ML}, s_3^k)$, where $k = 1, 2, \ldots, M$. This procedure, which is depicted in Fig. 2, reduces the optimum ML receiver complexity from M^3 to M.

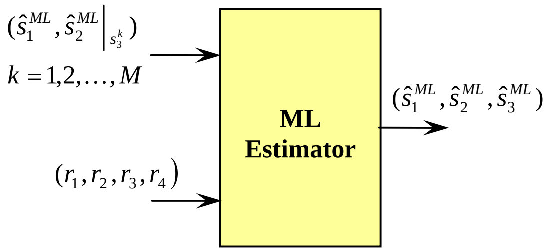

Fig. 2. Second stage of the estimator.

As mentioned above, the *a*, *b*, *c*, *d* parameters in the code matrix are design parameters to be optimized in order to obtain a full-diversity STC with a large coding gain. However, this task is not feasible, especially for higher constellation sizes. Fortunately, average transmit power and received SNR constraints further decrease the number of parameters to be optimized.

First, we point out that the described detector is optimum only when the magnitudes of *a* and *c* are equal. This can be easily seen by looking at the SNR at the receiver input and then at the threshold detector input. Indeed, these two SNR values are the same if and only if $|a| = |c|$. Second, in terms of the average transmitted power, the desired conditions can be expressed as

$$|a|^2 + |b|^2 / 2 = |c|^2 + |d|^2 / 2$$

$$|a|^2 + |c|^2 = |b|^2 + |d|^2 .$$

The first condition ensures an equal average transmit power at each symbol time, while the second condition ensures that equal total power is transmitted for each symbol. These equalities together with the constraint $|a| = |c|$ for optimal detection lead immediately to the fact that all the design parameters should have the same magnitude. Without any loss of generality, we can set the values of the parameters *a* and *c* to a single constant, say $1/\sqrt{2}$. Then, we have

$$|b| = |d| = 1/\sqrt{2} .$$

Now, we can choose the parameters *b* and *d* which lead to a full-diversity scheme with a large coding gain. According to the rank criterion, it is evident that in order to obtain full diversity, the difference matrix $(\mathbf{X}-\hat{\mathbf{X}})$ should be full rank. Thanks to the special form of (11), it can be shown that there exists a set of parameters *b* and *d* for which the resulting STC will have nonzero coding gain for any QAM signal constellation. In light of the design criteria given in Section II.C, this directly ensures that the obtained code will have full diversity for all QAM signal constellations. However, a generalized optimization procedure of the parameter set is beyond the scope of this paper. Instead, we provide an example for the proposed scheme leading to a large coding gain that is independent of the constellation size.

IV. PERFORMANCE RESULTS

We now provide a performance comparison between the proposed rate-3/4 STC and the two MIMO schemes included in the WiMAX system specifications, namely Alamouti's STC (Matrix A) and the SM (Matrix B). Two receive antennas were used in all of these schemes. The simulations were carried out for the QPSK, 16-QAM and 64-QAM signal constellations, and no channel coding was used. Fig. 3 shows the BER performance as a function of E_s / N_0 (E_s = average transmitted signal energy per antenna use) on an uncorrelated Rayleigh fading channel. The *b* and *d* parameters in our code were set as $b = d = (1 + j\sqrt{7})/4$. With these values, the proposed STC maximizes the diversity gain and, therefore, it achieves the same BER curve slope as Alamouti's STC with a coding gain of 2 which is independent of the constellation size. This is a crucial property which is satisfied by the best known codes since we do not want vanishing determinants.

The results of Fig. 3 indicate that the Alamouti scheme achieves a BER of 10^{-3} with an SNR of 10 dB for QPSK, 16.6 dB for 16-QAM, and 22.4 dB for 64-QAM. Next, we can observe that the SM scheme achieves this BER with an SNR of 18.6 dB for QPSK and 26.6 dB for 16-QAM. With 64-QAM, this MIMO scheme requires an SNR well in excess of 30 dB to reach this BER performance level. Finally, our new STC achieves a BER of 10^{-3} with an SNR of 13.8 dB for QPSK, 21.7 dB for 16-QAM, and 28.6 dB for 64-QAM.

Clearly, the Alamouti scheme has the best BER performance, but also the lowest bit rate on a given channel bandwidth. The SM scheme doubles the bit rate, but it involves a strong SNR loss, which increases at lower BER values. (This is due to the lower BER curve slope.) As evidenced from these results, the proposed rate-3/4 scheme is an interesting alternative to these two MIMO schemes, as it substantially improves BER performance compared to SM at the price of a 25% decrease in bit rate, and it increases the bit rate by 50% compared to the Alamouti scheme at the price of some SNR loss.

A closer examination of the results shows that at the spectral efficiency of 3 bits per antenna use, the proposed technique outperforms Alamouti's STC. Indeed, the new STC with 16-QAM and Alamouti's STC with 64-QAM have a spectral efficiency of 3 bits per antenna use, and the results indicate that at the BER of 10^{-3}, the former outperforms the latter by 0.7 dB.

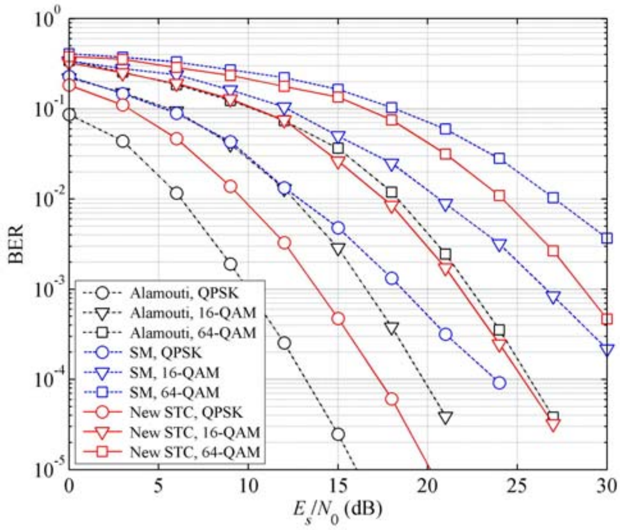

Fig. 3. Performance comparison of the new STC with Alamouti's STC and Spatial Multiplexing (SM).

V. CONCLUSIONS

In this paper, we have presented a full-diversity rate-3/4 2×2 STC whose optimum decoder complexity grows only linearly with the number of constellation points. We have compared its performance to the two MIMO schemes included in the IEEE 802.16e-2005 specifications, and the results indicated that it makes an interesting alternative providing further tradeoffs between performance and spectral efficiency. At the same spectral efficiency of 3 bits per antenna use, the proposed scheme was actually found to outperform Alamouti's STC.

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