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Research Article

A Simple Differential Modulation Scheme for Quasi-Orthogonal Space-Time Block Codes with Partial Transmit Diversity

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We report a simple differential modulation scheme for quasi-orthogonal space-time block codes. A new class of quasi-orthogonal coding structures that can provide partial transmit diversity is presented for various numbers of transmit antennas. Differential encoding and decoding can be simplified for differential Alamouti-like codes by grouping the signals in the transmitted matrix and decoupling the detection of data symbols, respectively. The new scheme can achieve constant amplitude of transmitted signals, and avoid signal constellation expansion; in addition it has a linear signal detector with very low complexity. Simulation results show that these partial-diversity codes can provide very useful results at low SNR for current communication systems. Extension to more than four transmit antennas is also considered.

1. INTRODUCTION

Transmit diversity techniques that can provide effective robustness over fading channels have been extensively investigated in recent years [1–13]. Orthogonal space-time block codes (O-STBCs) were reported in [1, 2], aiming at achieving maximum diversity gain. Later, in order to satisfy the high data rate requirement, a family of quasi-orthogonal space-time block codes (QO-STBCs) has been proposed in [3], which can obtain full rate but partial diversity by mapping the input data to one fixed constellation, and simulation results suggest that these codes can provide very useful results at low SNR. At high SNR, they perform worse than O-STBC due to the reduced diversity. Recently, improved quasi-orthogonal space-time block codes for four transmit antennas were reported in [4, 5], which can provide both full rate and full diversity. However, this is achieved at the cost of significant signal constellation expansion and thus further increase in the computational complexity.

All the above work assumes that the channel can be readily tracked at the transmitter or receiver. In order to combat the environment with poor channel information, [6, 7] differential orthogonal space-time block codes (D-O-STBCs) and [8–10] differential space-time modulation (DSTM) were developed based on the orthogonal properties of the transmission matrices. However, the transmission rate is still low for more than two transmit antennas. Recently, several full-rate, full-diversity differential QO-STBC (D-QO-STBC) schemes have been investigated in [11–13], yielding good performance at very high SNR. However, all these schemes involve the rotation of signal constellations and result in significant constellation expansion in spite of the promising performance. Moreover, it is also worth pointing out that the approaches for D-QO-STBC cannot be extended to the partial-diversity codes in [3] to obtain differential partial-diversity QO-STBC, otherwise all zero transmission matrices might be generated. To the best of our knowledge, there is no corresponding differential scheme so far proposed in the literature.

In this paper, we propose a simple quasi-orthogonal coding structure, which can be used to build up a differential partial-diversity QO-STBC scheme. Encoding and decoding can be carried out by grouping signals in the transmission matrix and decoupling the detection of data symbols. As a result, our method is very general and robust and has very low computational complexity: the decoder complexity is linear on the constellation size, as for O-STBC. It can provide half diversity and full rate without constellation expansion, using only one constellation. Note that we would not expect that our scheme could outperform those in [7–13] since only half diversity can be obtained by QO-STBC. However, just as QO-STBC [3] gives better results at relatively low SNR, such as often occurs in practice, our scheme can provide promising results in a useful range of SNR compared to the work in [7–13]. Note that [8] employs exhaustive search decoding, which has a prohibitive complexity, and [11–13] also have high complexity in both encoding and decoding and cause significant constellation expansion. Hence, from the encoding and decoding complexity point of view, our differential partial-diversity QO-STBC scheme is very promising.

2. PRELIMINARIES

2.1. System model

For simplicity and without loss of generality, we first consider a system with four transmit antennas and one receive antenna operating in a Rayleigh fading environment. At time t, symbols $s_{i,t}$, $i = 1, \dots, 4$, are transmitted from the four antennas simultaneously and r_t is the received signal. The system is modelled by

$$r_t = \sum_{i=1}^{4} h_i s_{i,t} + n_t, \tag{1}$$

where h_i is the path gain from transmitter i to the receive antenna. Here we assume that the channel is constant during a frame period and varies from one frame to another. The noise n_t consists of independent samples of a zero-mean complex Gaussian random variable with variance $E/(2SNR)$. E denotes the total power of transmitted signals.

2.2. Partial-diversity quasi-orthogonal space-time block codes

In this part, we first consider the following new quasi-orthogonal space-time block codes based on a Hadamard transformation for four transmit antennas at time $4t$:

$$\mathbf{S}_{4t} = \begin{pmatrix} \mathbf{S}_{4t}^{12} & \mathbf{S}_{4t}^{34} \\ \mathbf{S}_{4t}^{12} & -\mathbf{S}_{4t}^{34} \end{pmatrix}, \tag{2}$$

where $\mathbf{S}_{4t}^{ij} = \begin{pmatrix} s_{i,4t} & s_{j,4t} \\ -s_{j,4t}^{*} & s_{i,4t}^{*} \end{pmatrix}$. Note that $\mathbf{S}_{4t}^{ij}$ has a form similar to the Alamouti scheme. The differential encoding described later is based on these blocks. This code has rate one, but diversity order two, since each symbol passes through only two of the four transmit antennas. Note that the codes in (2) differ from the QO-STBC used in [3–5]: the Alamouti subgroup in (2) appears on the same group of transmitter antennas, which is a very useful property since it results in very low complexity decoding, as we will see in the next section. But the Alamouti group in [3–5] is distributed in the different column of the matrix.

3. DIFFERENTIAL ENCODING AND DECODING

3.1. Differential encoding process

In this section, we discuss how to obtain each subblock, $\mathbf{S}_{4t+4}^{ij}$, by our simple encoding method. At time interval $4t + 4$, a block of $4b$ bits at the encoder, denoted by d_{4t+4}^{1}, d_{4t+4}^{2}, d_{4t+4}^{3}, and d_{4t+4}^{4} (where each d_{4t+4}^{i}, $i = 1, \dots, 4$, represents a binary b-tuple), is modulated onto four symbols. For convenience, let M_{4t+4}^{1}, M_{4t+4}^{2}, M_{4t+4}^{3}, and M_{4t+4}^{4} refer to the resulting modulated signals from the constellation M. The differential encoder then produces the transmission matrix $\mathbf{S}_{4t+4}$ using the following subblock encoding:

$$\mathbf{s}_{4t+4}^{ij1} = \frac{\mathbf{m}_{4t+4}^{ij}\mathbf{S}_{4t}^{ij}}{C_{4t}^{ij}}, \tag{3}$$

where $\mathbf{s}_{4t+4}^{ijk}$ represents the kth row of $\mathbf{S}_{4t+4}^{ij}$, vector $\mathbf{m}_{4t+4}^{ij}$ consists of the ith and jth outputs from the "mapper" at time $4t+4$, $\mathbf{m}_{4t+4}^{ij} = [M_{4t+4}^{i} \;\; M_{4t+4}^{j}]$, $C_{4t}^{ij} = \sqrt{\mathrm{trace}(\mathbf{S}_{4t}^{ij}\mathbf{S}_{4t}^{ijH})}$, and H denotes complex conjugate transpose. Note that normalization by a factor of C_{4t}^{ij} is required in order to avoid large peak power variations in the transmitted signals. The rest of $\mathbf{S}_{4t+4}$ can be built up according to the structure of (2).

Note that the simple differential encoding process is based on each Alamouti block in (2). If the input is modulated onto four symbols taken from the *PSK* constellation and then the power of each constellation symbol is normalized to 0.5, the differentially encoded signals can maintain constant amplitude. In this case, $C_{4t}^{ij} = 1$ and thus the normalization is clearly not required. Other than this, the transmitted signals in our scheme, like those in [11–13], have nonconstant matrix norm.

3.2. Differential decoding process

The received signals for time $4t + 4$ can be written as

$$\mathbf{r}_{4t+4} = \mathbf{S}_{4t+4}\mathbf{h} + \mathbf{n}_{4t+4}, \tag{4}$$

where $\mathbf{r}_{4t+4} = [r_{4t+1} \;\; \cdots \;\; r_{4t+4}]^T$ and the channel state matrix $\mathbf{h} = [h_1 \;\; \cdots \;\; h_4]^T$, where T denotes transpose, $\mathbf{n}_{4t+4}$ consists of the noise terms. By further transformation, we can obtain

$$\mathbf{R}_{4t+4}^{12} = 2\mathbf{S}_{4t+4}^{12}\mathbf{H}_{12} + \mathbf{N}_{4t+4}^{12}, \tag{5}$$

$$\mathbf{R}_{4t+4}^{34} = 2\mathbf{S}_{4t+4}^{34}\mathbf{H}_{34} + \mathbf{N}_{4t+4}^{34}, \tag{6}$$

$$\mathbf{R}_{4t+4}^{12} = \begin{pmatrix} r_{4t+1} + r_{4t+3} & (r_{4t+2} + r_{4t+4})^{*} \\ (r_{4t+2} + r_{4t+4})^{*} & -(r_{4t+1} + r_{4t+3}) \end{pmatrix}, \tag{7}$$

$$\mathbf{R}_{4t+4}^{34} = \begin{pmatrix} r_{4t+1} - r_{4t+3} & (r_{4t+2} - r_{4t+4})^{*} \\ (r_{4t+2} - r_{4t+4})^{*} & -(r_{4t+1} - r_{4t+3}) \end{pmatrix}, \tag{8}$$

where $\mathbf{H}_{ij} = \begin{pmatrix} h_i & h_j^{*} \\ h_j & -h_i^{*} \end{pmatrix}$.

Recalling the encoding process in (3), we can reach

$$\mathbf{s}_{4t+4}^{121}(\mathbf{S}_{4t}^{12})^H = \mathbf{m}_{4t+4}^{12}\mathbf{S}_{4t}^{12}(\mathbf{S}_{4t}^{12})^H = C_{4t}^{12}\mathbf{m}_{4t+4}^{12}I_2, \quad (9)$$

$$\mathbf{s}_{4t+4}^{341}(\mathbf{S}_{4t}^{34})^H = \mathbf{m}_{4t+4}^{34}\mathbf{S}_{4t}^{34}(\mathbf{S}_{4t}^{34})^H = C_{4t}^{34}\mathbf{m}_{4t+4}^{34}I_2, \quad (10)$$

where I_2 is a 2×2 identity matrix. By differentially combining received signals from the previous time slots and then using (7)–(10) we have

$$\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{121H} = C_{4t}^{12}C_3M_{4t+4}^1 + n_{4t,4t+4}^1, \quad (11)$$

$$\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{122H} = C_{4t}^{12}C_3M_{4t+4}^2 + n_{4t,4t+4}^2, \quad (12)$$

$$\mathbf{r}_{4t+4}^{341}\mathbf{r}_{4t}^{341H} = C_{4t}^{34}C_4M_{4t+4}^3 + n_{4t,4t+4}^3, \quad (13)$$

$$\mathbf{r}_{4t+4}^{341}\mathbf{r}_{4t}^{342H} = C_{4t}^{34}C_4M_{4t+4}^4 + n_{4t,4t+4}^4, \quad (14)$$

where $\mathbf{r}_{4t}^{ijk}$ denotes the values in the kth row of $\mathbf{R}_{4t}^{ij}$, $C_3 = 4\sum_{i=1}^{2}|h_i|^2$, and $C_4 = 4\sum_{i=3}^{4}|h_i|^2$.

For convenience, let $T = C_{4t}^{12}C_3$ and $Q = C_{4t}^{34}C_4$. Obviously, if T and Q are available at the receiver, an exhaustive search over all combinations of M^1 to M^4 can be carried out to recover the most likely mapping signals in (11)–(14). The received signals can be rewritten in a matrix form as

$$\begin{aligned}\mathbf{r}_1 &= h_1\mathbf{s}_1 + h_2\mathbf{s}_2 + h_3\mathbf{s}_3 + h_4\mathbf{s}_4 + \mathbf{n}_1,\\ \mathbf{r}_3 &= h_1\mathbf{s}_1 + h_2\mathbf{s}_2 - h_3\mathbf{s}_3 - h_4\mathbf{s}_4 + \mathbf{n}_3,\end{aligned} \quad (15)$$

where vector $\mathbf{s}_i$ contains all the signals transmitted by antenna i in each trial and the length of those signals at each antenna is equal to L. We can derive the average channel power, neglecting the noise, by the following transformation:

$$C_3 = \frac{(\mathbf{r}_1^H\mathbf{r}_1 + \mathbf{r}_3^H\mathbf{r}_3)}{L}, \qquad C_4 = \frac{(\mathbf{r}_1^H\mathbf{r}_1 - \mathbf{r}_3^H\mathbf{r}_3)}{L}. \quad (16)$$

We can then multiply the received signal vector by its Hermitian transpose:

$$\mathbf{r}_{4t}^{121}\mathbf{r}_{4t}^{121H} = (C_{4t}^{12})^2C_3 + n_1'. \quad (17)$$

Similarly, we have

$$\mathbf{r}_{4t}^{341}\mathbf{r}_{4t}^{341H} = (C_{4t}^{34})^2C_4 + n_3', \quad (18)$$

where n_1' and n_3' denote the corresponding noise terms. So, the estimate of combined channel power and signal power can then be written as

$$T \approx \sqrt{\mathbf{r}_{4t}^{121}\mathbf{r}_{4t}^{121H}C_3}, \qquad Q \approx \sqrt{\mathbf{r}_{4t}^{341}\mathbf{r}_{4t}^{341H}C_4}. \quad (19)$$

Note that the additional complexity in the above detector comes only from the channel power and amplitude power estimation, which can be neglected. Next, we discuss the final decoding algorithm for differential partial-diversity QO-STBC according to the differential encoding schemes.

3.3. Decoding algorithms

3.3.1. Differential partial-diversity QO-STBC with QAM constellations

Now, we have all the results needed for differential decoding. In (11), for example, the decision signal $\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{121H}$ is a function only of input signals M_{4t+4}^1. Then by using a corresponding least square decoder, we can recover the signals from these constellations:

$$\hat{m} = \arg\min_{M_{4t+4}^1\in M} \left|\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{121H} - TM_{4t+4}^1\right|^2. \quad (20)$$

The detector described above can be further simplified to

$$\hat{m}_1 = \arg\min_{M_{4t+4}^1\in M} \{T\left|M_{4t+4}^1\right|^2 - 2\,\mathrm{Re}\,((\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{121H})^*M_{4t+4}^1)\}. \quad (21)$$

We can also use a similar method to decode other inputs:

$$\begin{aligned}\hat{m}_2 &= \arg\min_{M_{4t+4}^2\in M} \{T\left|M_{4t+4}^2\right|^2 - 2\,\mathrm{Re}\,((\mathbf{r}_{4t+4}^{121}\mathbf{r}_{4t}^{122H})^*M_{4t+4}^2)\},\\ \hat{m}_3 &= \arg\min_{M_{4t+4}^3\in M} \{Q\left|M_{4t+4}^3\right|^2 - 2\,\mathrm{Re}\,((\mathbf{r}_{4t+4}^{221}\mathbf{r}_{4t}^{221H})^*M_{4t+4}^3)\},\\ \hat{m}_4 &= \arg\min_{M_{4t+4}^4\in M} \{Q\left|M_{4t+4}^4\right|^2 - 2\,\mathrm{Re}\,((\mathbf{r}_{4t+4}^{221}\mathbf{r}_{4t}^{222H})^*M_{4t+4}^4)\}.\end{aligned} \quad (22)$$

The complexity of this process is linear and proportional to 2^b, since this is the number of combinations of constellation points to be examined. In practice it could be replaced by a slicing operation with even less complexity. The decoder in [8] has computational complexity 2^{4b}, and [11–13] have complexity about 2^{2b+1}. Note that the QAM constellation has better Euclidean distance than PSK, such that it can give a relatively better performance.

3.3.2. Differential partial-diversity QO-STBC with PSK constellations

If M is a PSK constellation, which has constant amplitude, the distribution of the combined received signals in (21) and

(22) will not be affected by the real constant values T and Q, which can be removed in the final detection. The major advantage of the use of the *PSK* constellation is that it allows the use of a very low complexity and can also obtain a reasonable system performance.

4. EXTENSIONS

4.1. Four transmit antennas

There are other possible structures that can provide behaviour similar to that of (2). A couple of examples is given below

$$\mathbf{S}_{4t} = \begin{pmatrix} \mathbf{S}_{4t}^{12} & -\mathbf{S}_{4t}^{34} \\ \mathbf{S}_{4t}^{12} & \mathbf{S}_{4t}^{34} \end{pmatrix},$$

$$\mathbf{S}_{4t} = \begin{pmatrix} \mathbf{S}_{4t}^{12} & -\mathbf{S}_{4t}^{34} \\ \mathbf{S}_{4t}^{12*} & \mathbf{S}_{4t}^{34*} \end{pmatrix}, \qquad (23)$$

$$\mathbf{S}_{4t} = \begin{pmatrix} \mathbf{S}_{4t}^{12} & \mathbf{S}_{4t}^{34} \\ -\mathbf{S}_{4t}^{12*} & \mathbf{S}_{4t}^{34*} \end{pmatrix}.$$

The principle here is to ensure that a given Alamouti block $\mathbf{S}_{4t}^{ij}$ appears on the same group of transmitter antennas (i.e., in the same column of the matrix), such that they can provide similar performance as the codes defined in (2).

4.2. Eight transmit antennas

While coherent quasi-orthogonal schemes exist for eight transmit antennas, it is not trivial to derive differential techniques directly from the existing literature, and very few schemes have so far been devised. In this section, following the ideas introduced before, we derive the differential scheme for partial-diversity QO-STBC for eight transmit antennas. Structures similar to that in (2) can be used to build up a rate 3/4 transmission matrix based on the rate 3/4 orthogonal space-time block code. An example is given below

$$\mathbf{S}_{8t} = \begin{pmatrix} \mathbf{S}_{8t}^{123} & \mathbf{S}_{8t}^{456} \\ \mathbf{S}_{8t}^{123} & -\mathbf{S}_{8t}^{456} \end{pmatrix}, \qquad (24)$$

where

$$\mathbf{S}_{8t}^{ijk} = \begin{pmatrix} s_{i,8t} & 0 & s_{j,8t} & -s_{k,8t} \\ 0 & s_{i,8t} & s_{k,8t}^* & s_{j,8t}^* \\ -s_{j,8t}^* & -s_{k,8t} & s_{i,8t}^* & 0 \\ s_{k,8t}^* & -s_{j,8t} & 0 & s_{i,8t}^* \end{pmatrix}, \qquad (25)$$

which can be encoded as a whole at the transmitter end. At time $8t+8$, $\mathbf{S}_{8t+8}^{ijk}$ is differentially encoded as

$$\mathbf{S}_{8t+8}^{ijk} = \frac{\mathbf{M}_{8t+8}^{ijk}\mathbf{S}_{8t}^{ijk}}{C_{8t}^{ijk}}, \qquad (26)$$

$$\mathbf{M}_{8t+8}^{ijk} = \begin{pmatrix} M_{8t+8}^{i} & 0 & -M_{8t+8}^{j*} & -M_{8t+8}^{k*} \\ 0 & M_{8t+8}^{i} & M_{8t+8}^{k} & -M_{8t+8}^{j} \\ M_{8t+8}^{j} & -M_{8t+8}^{k*} & M_{8t+8}^{i*} & 0 \\ M_{8t+8}^{k} & M_{8t+8}^{j*} & 0 & M_{8t+8}^{i*} \end{pmatrix}, \qquad (27)$$

$$C_{8t}^{ijk} = \sqrt{\mathbf{S}_{8t}^{ijk}\mathbf{S}_{8t}^{ijkH}}. \qquad (28)$$

Then, $\mathbf{S}_{8t+8}$ can be generated according to (24). We now discuss how to derive the corresponding decoding algorithm. The received signals for time $8t+8$ can be written as

$$\mathbf{r}_{8t+8} = \mathbf{S}_{8t+8}\mathbf{h} + \mathbf{n}_{8t+8}, \qquad (29)$$

where $\mathbf{r}_{8t+8} = [r_{8t+1} \ \cdots \ r_{8t+8}]^T$ and $\mathbf{h} = [h_1 \ \cdots \ h_8]^T$.

Similar to the four transmit antenna cases, we can further transform (29) as

$$\begin{aligned} \mathbf{r}_{8t+8}^1 &= 2\mathbf{S}_{8t+8}^{123}\mathbf{h}_1 + \mathbf{n}_{8t+8}^1, \\ \mathbf{r}_{8t+8}^2 &= 2\mathbf{S}_{8t+8}^{456}\mathbf{h}_2 + \mathbf{n}_{8t+8}^2, \\ \mathbf{r}_{8t+8}^1 &= \left[r_{8t+1} + r_{8t+5} \ \ r_{8t+2} + r_{8t+6} \ \ r_{8t+3} + r_{8t+7} \ \ r_{8t+4} + r_{8t+8} \right], \\ \mathbf{r}_{8t+8}^2 &= \left[r_{8t+1} - r_{8t+5} \ \ r_{8t+2} - r_{8t+6} \ \ r_{8t+3} - r_{8t+7} \ \ r_{8t+4} - r_{8t+8} \right], \end{aligned} \qquad (30)$$

where $\mathbf{h}_1 = [h_1 \ h_2 \ h_3 \ h_4]$ and $\mathbf{h}_2 = [h_5 \ h_6 \ h_7 \ h_8]$.

Recalling (26)–(30), the received signals at time $8t+8$ can be combined as

$$\begin{aligned} \mathbf{r}_{8t+8}^1\mathbf{r}_{8t}^{1H} &= C_{8t+8}^{123}C_1\mathbf{M}_{8t+8}^{123} + \mathbf{n}_{8t+8}^1\mathbf{r}_{8t}^{1H}, \\ \mathbf{r}_{8t+8}^2\mathbf{r}_{8t}^{2H} &= C_{8t+8}^{456}C_2\mathbf{M}_{8t+8}^{456} + \mathbf{n}_{8t+8}^2\mathbf{r}_{8t}^{2H}, \end{aligned} \qquad (31)$$

where $C_1 = 4\sum_{i=1}^{4}|h_i|^2$ and $C_2 = 4\sum_{i=5}^{8}|h_i|^2$. Also, for convenience, let $T = C_{8t+8}^{123}C_1$ and $Q = C_{8t+8}^{456}C_2$.

So far, we can clearly see that the differential encoding and decoding process for eight transmit antennas is almost identical to the process for four transmit antennas. Therefore, by following the power estimation and detection procedure in Section 3.2, we can obtain the signal decoder for partial-diversity QO-STBC.

(1) Differential partial-diversity QO-STBC decoder

$$
\begin{aligned}
\hat{m}_1 = \arg \min_{M^1_{8t+8}\in M} \Big\{ & T^2 \left| M^1_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(T (r^{11*}_{8t+8} r^{11}_{8t} + r^{12*}_{8t+8} r^{12}_{8t} \\
& \quad + r^{13}_{8t+8} r^{13*}_{8t} + r^{14}_{8t+8} r^{14*}_{8t}) M^1_{8t+8} \big) \Big\}, \\
\hat{m}_2 = \arg \min_{M^2_{8t+8}\in M} \Big\{ & T^2 \left| M^2_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(T (r^{11*}_{8t+8} r^{13}_{8t} + r^{12}_{8t+8} r^{14*}_{8t} \\
& \quad - r^{13}_{8t+8} r^{11*}_{8t} - r^{14*}_{8t+8} r^{12}_{8t}) M^2_{8t+8} \big) \Big\}, \\
\hat{m}_3 = \arg \min_{M^3_{8t+8}\in M} \Big\{ & T^2 \left| M^3_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(T (r^{11*}_{8t+8} r^{14}_{8t} - r^{12}_{8t+8} r^{13*}_{8t} \\
& \quad + r^{13*}_{8t+8} r^{12}_{8t} - r^{14}_{8t+8} r^{11*}_{8t}) M^3_{8t+8} \big) \Big\}, \\
\hat{m}_4 = \arg \min_{M^4_{8t+8}\in M} \Big\{ & Q^2 \left| M^4_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(Q (r^{21*}_{8t+8} r^{21}_{8t} + r^{22*}_{8t+8} r^{22}_{8t} \\
& \quad + r^{23}_{8t+8} r^{23*}_{8t} + r^{24}_{8t+8} r^{24*}_{8t}) M^4_{8t+8} \big) \Big\}, \\
\hat{m}_5 = \arg \min_{M^5_{8t+8}\in M} \Big\{ & Q^2 \left| M^5_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(Q (r^{21*}_{8t+8} r^{23}_{8t} + r^{22}_{8t+8} r^{24*}_{8t} \\
& \quad - r^{23}_{8t+8} r^{21*}_{8t} - r^{24*}_{8t+8} r^{22}_{8t}) M^5_{8t+8} \big) \Big\}, \\
\hat{m}_6 = \arg \min_{M^6_{8t+8}\in M} \Big\{ & Q^2 \left| M^6_{8t+8} \right|^2 \\
& - 2\,\mathrm{Re}\big(Q (r^{21*}_{8t+8} r^{24}_{8t} - r^{22}_{8t+8} r^{23*}_{8t} \\
& \quad + r^{23*}_{8t+8} r^{22}_{8t} - r^{24}_{8t+8} r^{21*}_{8t}) M^6_{8t+8} \big) \Big\},
\end{aligned} \tag{32}
$$

where r^{ij}_{8t} is the jth element of $\mathbf{r}^i_{8t}$. When a *PSK* constellation is applied, the above detectors can be further simplified without the need of power estimation like those in Section 3.3.2.

(2) Sixteen transmit antennas

For sixteen transmitter antennas, rate 1/2 O-STBC with partial diversity is given by

$$
\mathbf{S} = \begin{pmatrix} \mathbf{S}_{1234} & \mathbf{S}_{5678} \\ -\mathbf{S}_{1234} & \mathbf{S}_{5678} \end{pmatrix}, \qquad \mathbf{S} = \begin{pmatrix} \mathbf{S}_{1234} & \mathbf{S}_{5678} \\ -\mathbf{S}^*_{1234} & \mathbf{S}^*_{5678} \end{pmatrix}, \tag{33}
$$

where

$$
S_{ijkl} = \begin{pmatrix}
s_i & s_j & s_k & 0 & s_l & 0 & 0 & 0 \\
-s_j^* & s_i^* & 0 & s_k & 0 & s_l & 0 & 0 \\
-s_k^* & 0 & s_i^* & -s_j & 0 & 0 & s_l & 0 \\
0 & -s_k^* & s_j^* & s_i & 0 & 0 & 0 & s_l \\
-s_l^* & 0 & 0 & 0 & s_i^* & -s_j & -s_k & 0 \\
0 & -s_l^* & 0 & 0 & s_j^* & s_i & 0 & -s_k \\
0 & 0 & -s_l^* & 0 & s_k^* & 0 & s_i & s_j \\
0 & 0 & 0 & -s_l^* & 0 & s_k^* & -s_j^* & s_i^*
\end{pmatrix} \tag{34}
$$

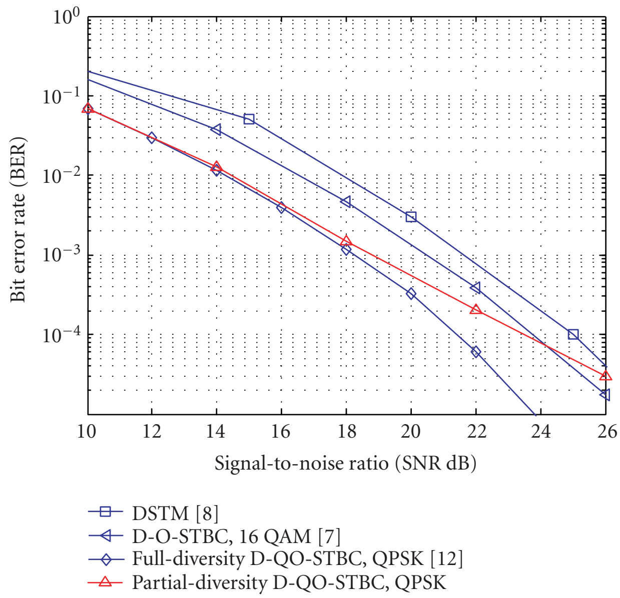

Figure 1: Differential QO-STBC schemes at rate 2 bps/Hz, four transmit antennas.

and the resulting codes have partial diversity. Similar methods can be used to establish the differential encoding and decoding process for partial-diversity QO-STBC.

5. SIMULATION RESULTS

Simulation results have been obtained assuming a wireless system with one receive antenna in a Rayleigh slow and flat fading channel. Performance is studied in comparison with the corresponding coherent detection results and the work in [7, 8, 12], whose results are independently simulated in our environment, and hence the same simulation parameters are used. Note that although [10] proposed low complexity decoding algorithm for DSTM, the corresponding performance cannot match that in [8]. Hence, only the results in [8] will be adopted for comparison. A block of symbols in the format of (2) or (24) is sent first as the reference, which carries no information and is unknown to the receiver. Note that these schemes are simulated with a relatively slowly time-varying channel, but since the decoding algorithm does not rely on channel coherence over more than two time slots, the performance will not be significantly affected by much more rapid time variance.

5.1. *Differential partial-diversity QO-STBC at rate 2 bps/Hz, four transmit antennas*

There is no purpose in applying BPSK with partial-diversity QO-STBC for transmission rate 1 bps/Hz, since differential O-STBC with a real constellation has been reported in [7]. In this part, QPSK constellations are used to generate a full-rate (two bits per symbol) transmission, half-diversity code. In this case, as shown in Figure 1, our scheme outperforms

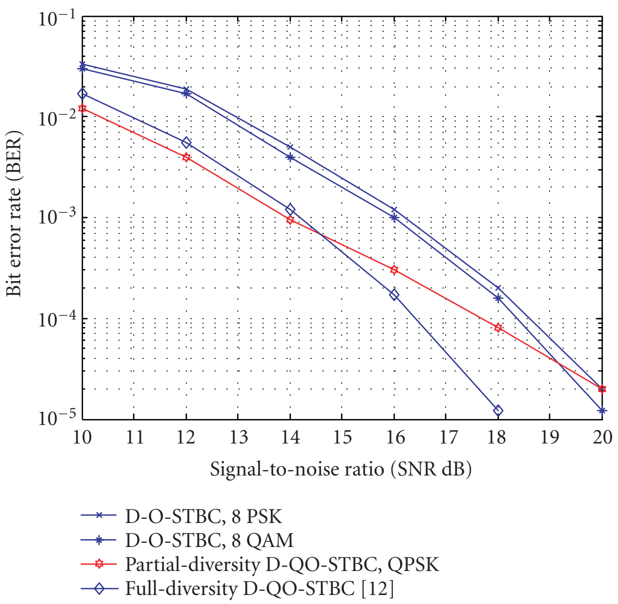

Figure 2: Differential QO-STBC schemes at rate 1.5 bps/Hz, eight transmit antennas.

DSTM [8] in the SNR region below 26 dB and D-O-STBC [7] for SNR below 24 dB. At higher SNR, [7, 8] are better, since at very high SNR, performance largely depends on the diversity of the system. Note that in many communication systems, the lower SNR range, below 20 dB, is more practically useful, assuming that an outer FEC code is used to ensure low enough BER for useful services. Comparing with full-diversity QO-STBC [12], we can observe that for SNR below 18 dB, our scheme provides almost the same performance. At high SNR, full-diversity QO-STBC begins to give better performance since it can obtain full diversity and full rate; however its complexity is much higher than the scheme considered here. Moreover, our scheme can avoid signal constellation expansion.

5.2. *Differential partial-diversity QO-STBC at rate 1.5 bps/Hz, eight transmit antennas*

Figure 2 gives the simulation results of rate 3/4 differential partial-diversity QO-STBC with eight transmit antennas at transmission rate 1.5 bps/Hz employing QPSK constellation. A similar conclusion can be also drawn that in the low SNR region, below 19 dB and 20 dB, it can provide better performance than the corresponding D-O-STBC with 8QAM and 8PSK constellations, respectively. But at high SNR, D-O-STBC begins to perform better. In comparison to full-diversity D-QO-STBC in [12], at SNR below 15 dB, partial-diversity D-QO-STBC can obtain a little better performance because of the comparatively robust coding structure in (24). But at high SNR, full-diversity D-QO-STBC obtains lower BER. Again, the major advantage of our scheme is that it has low complexity and avoids signal constellation expansion.

6. CONCLUSIONS

In this paper we present a QO-STBC-based differential modulation scheme for multiple antenna systems. The major contributions of the method are that the transmission signals can maintain constant amplitude, and avoid signal constellation expansion. They also have a linear signal detector with very low complexity. Simulation results show that these codes can provide very useful results in the practical range of SNR for current systems.

Note that it would be impossible to derive full-rate, full-diversity complex orthogonal space-time codes for more than two transmit antennas by our proposed coding structure using one constellation. Although recent work in [14, 15] describes the construction of full-rate, full-diversity complex orthogonal space-time codes for four transmit antennas, the code design is complicated, and exhaustive search decoding is required at the receiver end and the resulting performance cannot outperform those in [4, 5]. Hence, in this paper, we do not use the codes in [14, 15] for comparison.

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