DESIGN OF ORTHOGONAL SPACE-TIME BLOCK CODES FOR MIMO-OFDM SYSTEMS WITH FULL DIVERSITY AND FAST ML DECODING

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ABSTRACT

In this paper, a design of orthogonal space-time block codes (OSTBC) is proposed for MIMO-OFDM systems operating in frequency-selective fading channels. By stacking the repeated symbols as entries of an OSTBC matrix, the proposed OSTBC can achieve full diversity in frequency-selective fading channels. Moreover, the ML decoding can be performed separately on each symbol, i.e., single-symbol decoding. Simulation results show that for two transmit antennas the proposed OSTBC can achieve a higher diversity gain than the Alamouti code yet maintaining a single-symbol decoding simplicity.

1. INTRODUCTION

The orthogonal space-time block code (OSTBC) design has recently received recognition because the special structure of orthogonality achieves full-diversity in flat fading channels and simple maximum-likelihood (ML) decoding. The first OSTBC design was proposed by Alamouti in [1] for two transmit antennas and was then extended by Tarokh *et. al.* [2] for any number of transmit antennas. Later, a systematic construction of complex OSTBC of rates $(k+1)/(2k)$ for $M_t = 2k-1$ or $M_t = 2k$ transmit antennas for any positive integer k was proposed in [3].

Focusing on frequency-selective MIMO channels, the Alamouti code was applied to each subcarrier of OFDM for two transmit antennas [4]. It can only exploit space diversity. However, the maximum diversity gain is the product of the number of transmit antennas M_t, the number of receive antennas M_r and the number of channel paths L [5]. To exploit the potential multipath diversity, a variety of coding schemes have been proposed, such as space-time trellis coded (STTC) OFDM in [6, 7], space-frequency coding (SFC) in [8, 9], and space-time-frequency coding (STFC) in [10, 11]. However, most of the existing coding methods have to sacrifice low decoding complexity in order to achieve full diversity. Recently, a low-complexity full-diversity STFC was constructed in [12], but it can only achieve the low complexity over the paired real symbols, i.e., single complex symbol, in the case of two channel paths.

In this paper, a design of OSTBC is proposed for MIMO-OFDM systems. Using a special structure of stacking and repeating, the proposed OSTBC can achieve the full diversity gain $M_t M_r L$ in frequency-selective fading channels. Furthermore, the proposed codes design admits a fast ML decoding, i.e., single-symbol decoding, for any number of transmit antennas and any number of channel paths. The rest of the paper is organized as follows. In Section 2, a typical MIMO-OFDM system is described and the corresponding signal model is given. In Section 3, a design of full-diversity OSTBC over frequency-selective fading channels is proposed. The fast ML decoding of the proposed OSTBC is shown and the maximum diversity gain is proved. In Section 4, simulation results are presented and discussed.

2. CODED MIMO-OFDM SYSTEM

Focusing on MIMO-OFDM systems over quasi-static frequency-selective Rayleigh fading channels, we provide in this section a brief review on the STF coded signal model which covers the STC and the SFC.

Suppose that a block of N_s information symbols which are from discrete alphabet $\mathcal{A}$ such as PSK or QAM are encoded into $\mathbf{C}$ via space-time-frequency (STF) coding. The STF codeword $\mathbf{C}$ of size $NT \times M_t$ can be expressed as

$$\mathbf{C} = \begin{pmatrix} \mathbf{C}^1 \\ \vdots \\ \mathbf{C}^T \end{pmatrix} = \begin{pmatrix} \mathbf{c}_1^1 & \cdots & \mathbf{c}_{M_t}^1 \\ \vdots & \ddots & \vdots \\ \mathbf{c}_1^T & \cdots & \mathbf{c}_{M_t}^T \end{pmatrix}, \tag{1}$$

where $\mathbf{C}^u \in \mathbb{C}^{N \times M_t} (u = 1, \cdots, T)$ denotes the signals transmitted during the uth OFDM block. The jth column $\mathbf{c}_j^u$ of $\mathbf{C}^u$ will be fed to the N-point IFFT at the jth transmit antenna during the uth OFDM block.

At the ith receive antenna, the frequency domain signals during the uth OFDM block can be given by

$$\mathbf{Y}_i^u = \sqrt{\frac{\rho}{M_t}} \sum_{j=1}^{M_t} \operatorname{diag}(\mathbf{c}_j^u) \mathbf{H}_{i,j} + \mathbf{n}_i^u, \tag{2}$$

where $\mathbf{H}_{i,j} \in \mathbb{C}^{N \times 1}$ denotes the channel frequency response from the jth transmit antenna to the ith receive antenna, $\mathbf{n}_i^u \in \mathbb{C}^{N \times 1}$

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denotes noise and its kth entry $n_i^u(k)$ is a circular symmetric, complex Gaussian with zero mean and unit variance and assumed to be statistically independent with respect to k, i and u. The normalization factor ρ ensures that ρ is the signal-to-noise ratio (SNR) at each receive antenna, independently of M_t.

With some matrix permutations, (2) can be rewritten as [11]

$$\mathbf{Y}_i^u = \sqrt{\frac{\rho}{M_t}}\mathbf{X}_u \mathbf{h}_i^u + \mathbf{n}_i^u, \tag{3}$$

for $u = 1, \cdots, T$ and $i = 1, \cdots, M_r$, where

$$\begin{aligned} \mathbf{X}_u &= \begin{bmatrix} \mathbf{D}_0\mathbf{C}^u & \cdots & \mathbf{D}_{L-1}\mathbf{C}^u \end{bmatrix}, \\ \mathbf{h}_i^u &= \begin{bmatrix} (\tilde{\mathbf{h}}_{i,0}^u)^{\mathcal{T}} & \cdots & (\tilde{\mathbf{h}}_{i,L-1}^u)^{\mathcal{T}} \end{bmatrix}^{\mathcal{T}}, \end{aligned}$$

where the superscript $^{\mathcal{T}}$ denotes the transpose. The matrix $\mathbf{D}_l = \mathrm{diag}\begin{pmatrix} 1 & e^{-\mathbf{j}2\pi\frac{\tau_l}{T_s}} & \ldots & e^{-\mathbf{j}2\pi(N-1)\frac{\tau_l}{T_s}} \end{pmatrix}$. T_s denotes the effect duration of an OFDM block and τ_l is the delay of the lth path. $\tilde{\mathbf{h}}_{i,l}^u = \begin{pmatrix} h_{i,1}^u(l) & \cdots & h_{i,M_t}^u(l) \end{pmatrix}^{\mathcal{T}}$, for $l = 0, \cdots, L-1$. The entry $h_{i,j}^u(l)$ denotes the lth tap of the channel between the jth transmit antenna and the ith receive antenna during the uth block and $\mathbf{E}[|h_{i,j}^u(l)|^2] = \delta_l^2$ for any l, i, j and u. As the channel is assumed to be quasi-static, it has $\mathbf{h}_i^1 = \cdots = \mathbf{h}_i^T$. In the following, the superscript u in $\mathbf{h}_i^u$ will be omitted.

Stacking all the received signals in (3) into a vector for $u = 1, \cdots, T$ and $i = 1, \cdots, M_r$, we can obtain

$$\mathbf{Y} = \sqrt{\frac{\rho}{M_t}}\mathbf{X}\mathbf{h} + \mathbf{n}$$

where

$$\begin{aligned} \mathbf{Y} &= \begin{bmatrix} (\mathbf{Y}_1^1)^{\mathcal{T}} & \cdots & (\mathbf{Y}_1^T)^{\mathcal{T}} & \cdots & (\mathbf{Y}_{M_r}^T)^{\mathcal{T}} \end{bmatrix}^{\mathcal{T}}, \quad (4) \\ \mathbf{X} &= \mathbf{I}_{M_r} \otimes \mathbf{G}, \\ \mathbf{h} &= \begin{bmatrix} \mathbf{h}_1^{\mathcal{T}} & \cdots & \mathbf{h}_{M_r}^{\mathcal{T}} \end{bmatrix}^{\mathcal{T}}, \\ \mathbf{n} &= \begin{bmatrix} (\mathbf{n}_1^1)^{\mathcal{T}} & \cdots & (\mathbf{n}_1^T)^{\mathcal{T}} & \cdots & (\mathbf{n}_{M_r}^T)^{\mathcal{T}} \end{bmatrix}^{\mathcal{T}}, \quad (5) \end{aligned}$$

and the $NT \times M_tL$ matrix $\mathbf{G}$ is given by

$$\mathbf{G} = \begin{bmatrix} (\mathbf{I}_T \otimes \mathbf{D}_0)\mathbf{C} & \cdots & (\mathbf{I}_T \otimes \mathbf{D}_{L-1})\mathbf{C} \end{bmatrix}. \tag{6}$$

The symbol $\otimes$ represents the Kronecker product.

In the code design of $\mathbf{C}$, diversity gain is always mostly concerned and it can be given by [11]

$$\begin{aligned} r &= M_r \cdot \min_{\forall \mathbf{C} \neq \hat{\mathbf{C}}} \left[\mathrm{rank}\left((\mathbf{G} - \hat{\mathbf{G}})\mathbf{Q}^{\frac{1}{2}}\right)\right] \\ &\leq M_t M_r L, \end{aligned} \tag{7}$$

where $\mathbf{Q} = \boldsymbol{\Upsilon} \otimes \mathbf{I}_{M_t}$ and $\boldsymbol{\Upsilon} = \mathrm{diag}\begin{pmatrix} \delta_0^2 & \delta_1^2 & \cdots & \delta_{L-1}^2 \end{pmatrix}$. It shows that the diversity gain is $M_t M_r L$ in frequency-selective fading channels.

3. AN OSTBC FOR MIMO-OFDM ACHIEVING FULL DIVERSITY AND FAST ML DECODING

In this section, we propose a design of OSTBC and show that they can both achieve full-diversity and fast ML detection in frequency-selective fading channels.

Let $\bar{\mathcal{G}}_{M_t}$ be a $Q \times M_t$ matrix with entries of complex linear combinations of elements from the set

$$\{\pm s_1, \pm s_1^*, \pm s_2, \pm s_2^*, \cdots, \pm s_{N_u}, \pm s_{N_u}^*\}.$$

It is an OSTBC design if the following orthogonality holds for any complex values of $s_{j'}$, $(j' = 1, 2, \cdots, N_u)$:

$$\bar{\mathcal{G}}_{M_t}^{\mathcal{H}}\bar{\mathcal{G}}_{M_t} = (|s_1|^2 + |s_2|^2 + \cdots + |s_{N_u}|^2)\mathbf{I}_{M_t},$$

where $^{\mathcal{H}}$ denotes the conjugate transpose. The N_u information symbols $s_{j'}, j' = 1, 2, \cdots, N_u$ coming from a signal constellation will be transmitted through M_t transmit antennas in Q time slots. The rate of the general OSTBC $\bar{\mathcal{G}}_{M_t}$ is $\bar{\mathcal{R}}_{OSTBC} = N_u/Q$. In [3], a closed form design of complex general OSTBC of rate $\bar{\mathcal{R}}_{OSTBC} = \frac{\lceil M_t/2 \rceil + 1}{2 \cdot \lceil M_t/2 \rceil}$ is given. $\lceil x \rceil$ denotes the smallest integer larger than x.

Based on the general OSTBC design, for example, in [3], we develop an OSTBC tailored for MIMO-OFDM system. It is structured as the following $QN \times M_t$ matrix:

$$\mathbf{C} = \sqrt{\gamma}\mathcal{G}_{M_t} \otimes \mathbf{1}_{\Gamma \times 1}, \tag{8}$$

where the scalar $\sqrt{\gamma}$ ensures that the OSTBC obeys the energy constraint, i.e., $||\mathbf{C}||_F^2 = QNM_t$, and $\mathbf{1}_{\Gamma \times 1}$ is a length-Γ column vector with all 1's entries. The matrix $\mathcal{G}_{M_t}$ of size $\frac{NQ}{\Gamma} \times M_t$ has the same structure as the general OSTBC matrix $\bar{\mathcal{G}}_{M_t}$ but an information symbol, $s_{j'}$, in $\bar{\mathcal{G}}_{M_t}$ is replaced by a vector of symbols, $\mathbf{S}_{j'}$, where $\mathbf{S}_{j'} = \begin{bmatrix} s_{j'}(1) & \cdots & s_{j'}(\frac{N}{\Gamma}) \end{bmatrix}^{\mathcal{T}} \in \mathcal{A}^{\frac{N}{\Gamma} \times 1}, j' = 1, \cdots, N_u$. $\mathcal{A}$ is the discrete alphabet such as PSK or QAM constellations normalized into the unity power. $\Gamma = 2^{\lceil \log_2 L' \rceil}$ for $1 \leq L' \leq L$. Clearly, $\Gamma \geq L'$ and $\frac{N}{\Gamma}$ is a positive integer. The orthogonality of (8) can be verified by

$$\mathbf{C}^{\mathcal{H}}\mathbf{C} = \gamma\Gamma\left(\sum_{j'=1}^{N_u}\sum_{n=1}^{N/\Gamma}|s_{j'}(n)|^2\right)\mathbf{I}_{M_t},$$

where $\mathbf{I}_{M_t}$ denotes the identity matrix of size $M_t \times M_t$.

An example of the proposed OSTBC for two transmit antennas is given by the following $2N \times 2$ matrix as

$$\mathbf{C} = \begin{pmatrix} \mathbf{S}_1 & \mathbf{S}_2 \\ -\mathbf{S}_2^* & \mathbf{S}_1^* \end{pmatrix} \otimes \mathbf{1}_{\Gamma \times 1}.$$

Compared with the general STFC form (1), the proposed OSTBC for two transmit antennas can be described as follows: in the first OFDM block, $\mathbf{S}_1 \otimes \mathbf{1}_{\Gamma \times 1}$ and $\mathbf{S}_2 \otimes \mathbf{1}_{\Gamma \times 1}$ are sent from the first and the second transmit antenna, respectively; in the second OFDM block, $-\mathbf{S}_2^* \otimes \mathbf{1}_{\Gamma \times 1}$ and $\mathbf{S}_1^* \otimes \mathbf{1}_{\Gamma \times 1}$ are sent from the two transmit antennas, respectively.

Here, we would like to give the following theorem.

Theorem 1: Suppose that an MIMO-OFDM system with M_t transmit antennas and M_r receive antennas has N OFDM tones and that the MIMO channels are spatially uncorrelated and each channel experiences the frequency-selective fading characterized by L independent paths, in which the maximum path delay is less than the length of the cyclic prefix of OFDM and the path gains are constant over T OFDM blocks. The STBC as described in (8), for any $1 \leq L' \leq L$,

1. achieve a symbol transmission rate as $\frac{1}{\Gamma}$ of that of general OSTBC, where $\Gamma = 2^{\lceil \log_2 L' \rceil}$;
2. have a fast ML decoding algorithm with single-symbol decoding; and
3. can achieve a diversity gain $M_t M_r L'$. If $L' = L$, then the full-diversity can be achieved.

Proof of Theorem 1.1

Using (1), we can observe that the total number of information symbols embedded in one codeword of (8) transmitted through N subcarriers in M_t transmit antennas during $T(T = Q)$ OFDM blocks is $N_s = N_u N/\Gamma$. Thus the symbol rate of the proposed OSTBC in (8) is

$$\mathcal{R} = \frac{N/\Gamma \cdot N_u}{NQ} = \frac{1}{\Gamma}\frac{N_u}{Q} = \frac{1}{\Gamma}\bar{\mathcal{R}}_{\mathrm{OSTBC}}.$$

Proof of Theorem 1.2

Combining (8) with (2), the received signals at the ith receive antenna during T ($T = Q$) OFDM blocks can be given by

$$\begin{bmatrix} \mathbf{Y}_i^1 \\ \vdots \\ \mathbf{Y}_i^T \end{bmatrix} = \sqrt{\frac{\rho}{M_t}}\mathbf{\Phi} \begin{bmatrix} \mathbf{H}_{i,1} \\ \vdots \\ \mathbf{H}_{i,M_t} \end{bmatrix} + \begin{bmatrix} \mathbf{n}_i^1 \\ \vdots \\ \mathbf{n}_i^T \end{bmatrix}$$

where $\mathbf{\Phi} \in \mathbb{C}^{NQ\times NM_t}$ has the same form as the general OSTBC $\bar{\mathcal{G}} \in \mathbb{C}^{Q\times M_t}$ but with the $N \times N$ block matrix entries $\mathbf{\Lambda}_{j'} = \sqrt{\gamma}\mathrm{diag}(\mathbf{S}_{j'} \otimes \mathbf{1}_{\Gamma\times 1})$ instead of the scalar entry $s_{j'}$ in $\bar{\mathcal{G}}$, for $j' = 1, \cdots, N_u$. Let $\mathbf{H} = \begin{bmatrix} \mathbf{H}_{1,1}^{\mathcal{T}} & \cdots & \mathbf{H}_{M_r,M_t}^{\mathcal{T}} \end{bmatrix}^{\mathcal{T}}$. With the notations (4) and (5), we can have

$$\mathbf{Y} = \sqrt{\frac{\rho}{M_t}}\mathbf{\Phi}'\mathbf{H} + \mathbf{n},$$

where $\mathbf{\Phi}' = \mathbf{I}_{M_r} \otimes \mathbf{\Phi}$. Note

$$(\mathbf{\Phi}')^{\mathcal{H}}\mathbf{\Phi}' \quad = \quad \gamma\mathbf{I}_{M_rM_t} \otimes \mathrm{diag}\left(\sum_{j'=1}^{N_u} |\mathbf{S}_{j'}|^2 \otimes \mathbf{1}_{\Gamma\times 1}\right)$$

the ML decoding can be reduced to

$$\hat{\mathbf{S}} = \arg\min_{s\in\mathcal{A}} \left\{ \sum_{j'=1}^{N_u}\sum_{n=1}^{N/\Gamma} f_{j',n}(s_{j'}(n)) \right\},$$

where $f_{j',n}$ is a quadratic form of $s_{j'}(n)$. It can be seen that the ML detection can be done separately on each $s_{j'}(n)$. This leads to the fast ML decoding.

Proof of Theorem 1.3

From (7) it is known that the achievable diversity gain of the proposed OSTBC depends on the minimum rank of $(\mathbf{G} - \hat{\mathbf{G}})\mathbf{Q}^{\frac{1}{2}}$ for any two different codewords $\mathbf{C}$ and $\hat{\mathbf{C}}$ given by (8), where $\mathbf{G}$ (and $\hat{\mathbf{G}}$ accordingly) is defined in (6).

For any $\breve{\mathbf{C}} = \mathbf{C} - \hat{\mathbf{C}} \neq \mathbf{0}$, there exists at least one nonzero $\breve{s}_{j'}(n_0) = s_{j'}(n_0) - \hat{s}_{j'}(n_0)$ in $\breve{\mathbf{C}}$ for $1 \le n_0 \le \frac{N}{\Gamma}$ and $1 \le j' \le N_u$. Let $\sigma(n_0)$ denote the smallest row index of $s_{j'}(n_0)$ in $\mathbf{C}$. Let

$$\begin{aligned} b_{m,u} = (u-1)N + \sigma(n_0) + (m-1)\nu, \\ m = 1, \cdots, \Gamma;\ u = 1, \cdots, T, \end{aligned}$$

where ν is the separation factor representing the distance between any two repeated symbols during one OFDM block. For the proposed OSTBC in (8), it has $\nu = 1$ and $\sigma(n_0) = (n_0 - 1)\Gamma + 1$, for $1 \le n_0 \le \frac{N}{\Gamma}$. Further define a set

$$\mathcal{I} = \{\ b_{1,1},\ \ \cdots,\ \ b_{1,T},\ \ \cdots,\ \ b_{\Gamma,1},\ \ \cdots,\ \ b_{\Gamma,T}\ \}.$$

For $1 \le L' \le L$ and $\Gamma = 2^{\lceil \log_2 L' \rceil}$, let

$$\mathbf{G}' = \begin{bmatrix} (\mathbf{I}_T \otimes \mathbf{D}_0)\mathbf{C} & \cdots & (\mathbf{I}_T \otimes \mathbf{D}_{L'-1})\mathbf{C} \end{bmatrix}, \qquad (9)$$

which are in fact the first M_tL' columns of the matrix $\mathbf{G}$ as shown in (6). Then we construct a matrix $\mathbf{B} \in \mathbb{C}^{T\Gamma\times M_tL'}$ by extracting the $\mathcal{I}(p)$th row of $\mathbf{G}' \in \mathbb{C}^{TN\times M_tL'}$ as the pth row of $\mathbf{B}$ for all p, $1 \le p \le T\Gamma$. In fact, $\mathbf{B}$ is a sub-matrix of $\mathbf{G}'$ after some rows permutation. Consequently, the $T\Gamma \times M_tL'$ matrix $\breve{\mathbf{B}} = \mathbf{B} - \hat{\mathbf{B}}$ can be given by

$$\breve{\mathbf{B}} = \sqrt{\gamma}(\mathbf{V} \otimes \mathbf{I}_T)\left(\mathbf{I}_{L'} \otimes \bar{\mathcal{G}}\right), \qquad (10)$$

where $\bar{\mathcal{G}}$ is the OSTBC matrix of size $T \times M_t$ with variables $\breve{s}_1(n_0), \breve{s}_2(n_0), \cdots, \breve{s}_{N_u}(n_0)$ and the (m,l)th entry of the $\Gamma \times L'$ $(\Gamma \ge L')$ matrix $\mathbf{V}$ is $[\mathbf{V}]_{m,l} = \xi^{(b_{m,1}-1)\tau_l}$, $m = 0, 1, \cdots, \Gamma$ and $l = 0, 1, \cdots, L' - 1$. $\xi = e^{-\mathrm{j}2\pi/T_s}$. It can be proved that $\mathrm{rank}(\mathbf{V}) = L'$. Using (10) we can have

$$\mathbf{Q}^{\frac{\mathcal{H}}{2}}\breve{\mathbf{B}}^{\mathcal{H}}\breve{\mathbf{B}}\mathbf{Q}^{\frac{1}{2}} = \gamma(\mathbf{\Upsilon}^{\frac{\mathcal{H}}{2}}\mathbf{V}^{\mathcal{H}}\mathbf{V}\mathbf{\Upsilon}^{\frac{1}{2}}) \otimes (\bar{\mathcal{G}}^{\mathcal{H}}\bar{\mathcal{G}}). \qquad (11)$$

Its rank is $L'M_t$. Thus $(\mathbf{B} - \hat{\mathbf{B}})\mathbf{Q}^{\frac{1}{2}} \in \mathbb{C}^{T\Gamma\times M_tL'}$ is of full rank. Considering that $\mathbf{B} - \hat{\mathbf{B}}$ is a sub-matrix of $\mathbf{G}' - \hat{\mathbf{G}}'$ with some selected rows, $(\mathbf{G}' - \hat{\mathbf{G}}')\mathbf{Q}^{\frac{1}{2}} \in \mathbb{C}^{TN\times M_tL'}$ is also of full rank. Recall (9), the matrix $\mathbf{G}'$ is a sub-matrix of $\mathbf{G}$ with the first ML' columns of $\mathbf{G}$. Then the minimum rank of $(\mathbf{G}' - \hat{\mathbf{G}}')\mathbf{Q}^{\frac{1}{2}} \in \mathbb{C}^{TN\times M_tL}$ is ML'. Therefore, from (7) we obtain that the proposed OSTBC achieves the diversity gain M_tM_rL' over frequency-selective fading channels. If L' is chosen to be equal to L, the OSTBC achieves the full-diversity M_tM_rL.

4. SIMULATION RESULTS

A 2×1 system with 64 OFDM tones is simulated. The bandwidth is 20 MHz and the cyclic prefix of the OFDM is fixed to 16. A two-ray channel model is considered with equal power and the second path delay is 0.5 μs. Four coding schemes combined with OFDM are considered at 1 bit/s/Hz. They are space-time coding (STC) without repeating, i.e., Alamouti coded OFDM [4], STC via repeating [8], space-frequency coding (SFC) via algebraic rotation [9] and the proposed OSTBC in (8) with $L' = L = 2$. If the channel path delay profile is known at transmitter, then an optimal permutation can be applied to achieve a better coding gain [9].

Table 1. Performance comparison in diversity gain and decoding complexity at 1 bit/s/Hz.

	Diversity Gain	Number of symbols jointly in decoding
STC [4]	2	1
STC [8]	4	2
SFC [9]	4	4
Proposed OSTBC	4	1

It can be seen from Fig. 1 that the STC without repeating marked by " $*$ " gives the worst BER performance. This is due to the fact that the space diversity gain 2 is achieved only [4]. To achieve the maximum diversity gain 4 in the simulated system, the STC was suggested to be repeated two times as shown in [8]. It can also be observed that the STC via mapping [8], the SFC ($\nu = 1$) [9], and the proposed OSTBC ($\nu = 1$) can achieve the comparable performance with a deeper slope than the STC without repeating. With optimum permutation ($\nu = 16$), the SFC and the proposed OSTBC can both achieve a better BER performance than the ones without permutation ($\nu = 1$).

The ML decoding complexity can be evaluated by the number of symbols involved in the detection. In Table 1, it is shown

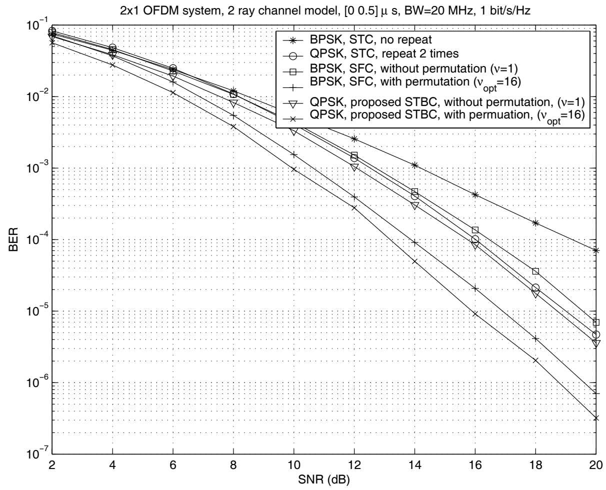

Fig. 1. Performance comparison among the Alamouti coded OFDM [4], STC via repeating [8], space-frequency coding (SFC) via algebraic rotation [9], and the proposed STBC for 2×1 OFDM system in 2-ray channel model ($N = 64$, BW=20 MHz, $\tau = [0, 0.5]\mu s$) at 1 bit/s/Hz.

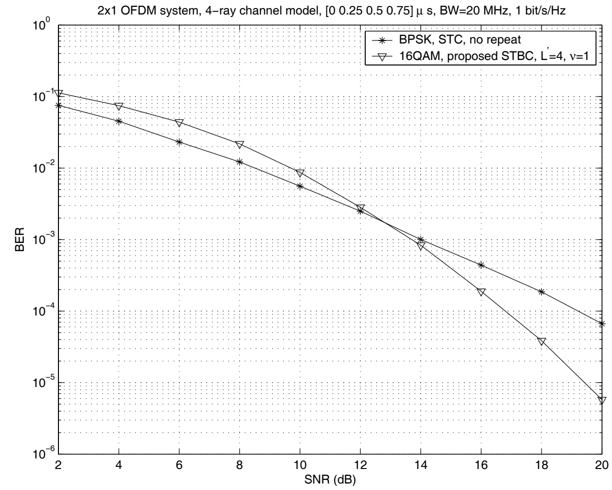

Fig. 2. Performance comparison between the Alamouti coded OFDM [4] and the proposed full-diversity STBC ($L' = L$) for 2×1 OFDM system in 4-ray channel model ($L = 4$, $N = 64$, BW=20 MHz, $\tau = [0, 0.25, 0.5, 0.75]\mu s$) at 1 bit/s/Hz.

that only the Alamouti coded OFDM in [4] and the proposed OSTBC have single-symbol decoding complexity. But the the proposed OSTBC can achieve the maximum diversity gain 4 while the Alamouti coded OFDM can only achieve the space diversity 2.

To further investigate the BER performance comparison between the two low-complexity coded OFDM, viz. the Alamouti coded OFDM [4] and the proposed OSTBC, we considered a 4-ray channel model ($L = 4$). The path delays are $[0, 0.25, 0.5, 0.75]$ μs. To keep at 1 bit/s/Hz bandwidth efficiency, BPSK is used for Alamouti coded OFDM [4] and 16QAM is for our proposed OSTBC in (8) with $L' = L = 4$. The simulation results in Fig. 2 show that in the low SNR region less than 12 dB the rate-1 Alamouti coded OFDM has a better BER performance than our proposed OSTBC. This is due to the fact that the Alamouti code with BPSK has a large coding gain than our proposed OSTBC with 16QAM. But when SNR is larger than 12 dB, the proposed OSTBC produces a BER curve with deeper slope than the Alamouti code. It is resulted from the multipath diversity gain exploited by our code design.

5. CONCLUSION

In this paper, an OSTBC design is proposed for MIMO-OFDM systems. By replacing the scalar entry of the general OSTBC with a vector of stacked and repeated symbols, the proposed OSTBC can achieve the full-diversity and the fast ML decoding in frequency-selective fading channels.

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