Construction of Low Delay Maximal Rate Single-Symbol Decodable Distributed STBC with Channel Phase Information

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Abstract: Exploiting the source-to-relay channel phase information at the relays can increase the rate upper-bound of distributed orthogonal space-time block codes (STBC) from $2/K$ to $1/2$, where K is the number of relays. This technique is known as distributed orthogonal space-time block codes with channel phase information (DOSTBC-CPI). However, the decoding delay of existing DOSTBC-CPIs is not optimal. Therefore, based on the rate of $1/2$ balanced complex orthogonal design (COD), an algorithm is provided to construct a maximal rate DOSTBC-CPI with only half the decoding delay of existing DOSTBC-CPI. Simulation results show that the proposed method exhibits lower symbol error rate than the existing DOSTBC-CPIs.

Key words: distributed STBC; channel phase information; decoding delay; single-symbol decoding

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0 Introduction

Distributed space-time block coding is an effective way of implementing cooperative communication systems[1-4]. To find distributed space-time block codes (DSTBC) in the set of complex orthogonal designs (COD) is natural and reasonable. However, the decoding of DSTBC depends on the covariance matrix $\boldsymbol{\Gamma}$ of the equivalent noise at the destination, which will destroy the orthogonality of COD. Therefore, when the COD for point-to-point multiple input multiple output (MIMO) systems are directly applied to cooperative wireless networks, they are usually no longer single-symbol decodable and the decoding complexity grows exponentially with the size of the symbol constellation used[5].

To construct single symbol decodable DSTBCs, Yi et al.[5] proposed distributed row-monomial orthogonal space-time block codes (DOSTBC). The rate of row-monomial DOSTBC is upper-bounded by $2/K$, where K is the number of relays. Although this is twice that of the repetition-based schemes, it is still inversely proportional to the number of relays. To address this issue, Sreedhar et al.[6] and Yi et al.[7] have independently illustrated that exploiting the source-to-relay channel phase information (CPI) at the relays can increase the rate of DOSTBC dramatically from $2/K$ to $1/2$, and this rate is achievable by the well-known rate-1/2 COD given by Tarokh et al.[8] (denoted as TJC code). The DOSTBCs that use CPI are known DOSTBC-CPIs. However, Refs. [9—10] show that the decoding delay of the TJC code is not optimal.

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Therefore, based on the balanced COD (BCOD) given in Ref. [10], we propose an algorithm to construct a class of DOSTBC-CPIs that not only retain all the advantages of the TJC code such as maximal rate, full diversity and single-symbol decodable, but also hold only half the decoding delay of the latter. Simulation results show that the proposed design also outperforms TJC code in terms of symbol error rate (SER).

1 System Model

Consider a cooperative wireless network as shown in Fig. 1, with one source node S, one destination node D, and K relay nodes denoted as R_i for $i=1, 2, \cdots, K$.

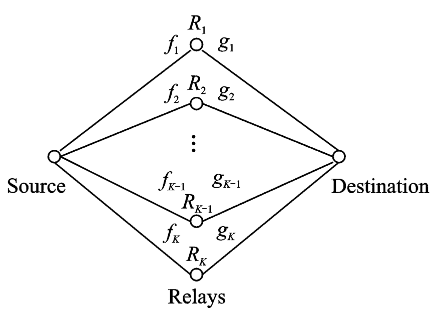

Fig. 1 Cooperative wireless network model

Each node has a single antenna, which can be used for transmitting and receiving, but not simultaneously. The relays are placed randomly and independently. The channel between S and R_i is denoted as $f_i=|f_i|e^{j\theta_i}$, where θ_i is the phase of f_i. The channel between R_i and D is denoted as g_i. Assume that f_i and g_i are independent complex Gaussian random variables with zero mean and unit variance. As described in Refs. [6, 7], the destination is assumed to have full channel state information (CSI) and the relays are assumed to have the source-to-relay CPI by using pilot signals, while the source is assumed to have no CSI.

Transmission from source to destination is carried out in two stages. In the first stage, the source transmits N complex information-bearing symbols $x_1, x_2, \cdots, x_N$, chosen in a normalized signal set in N consecutive time slots. The symbols can be denoted as an N-dimensional symbol vector $\boldsymbol{x}=(x_1 \quad x_2 \quad \cdots \quad x_N)$, normalized as $E(\boldsymbol{x}\boldsymbol{x}^{\mathrm{H}})=1$. The power per channel used for the source to transmit the symbols is P_s. Thus the symbol vector transmitted by the source in the first stage can be denoted as $\sqrt{P_sN}\,\boldsymbol{x}$. As described in Refs. [1, 5—7], we assume that the coherence time of the source-to-relay channel is larger than N time slots, so that the source-to-relay channels remain static during the transmission. Then the signal received by relay R_i is given by

$$\boldsymbol{r}_i=\sqrt{P_sN}f_i\boldsymbol{x}+\boldsymbol{n}_i=\sqrt{P_sN}\,|f_i|\,e^{j\theta_i}\boldsymbol{x}+\boldsymbol{n}_i \tag{1}$$

where $\boldsymbol{r}_i=(r_{i,1} \quad r_{i,2} \quad \cdots \quad r_{i,N})$ is the received symbol vector, and $\boldsymbol{n}_i=(n_{i,1} \quad n_{i,2} \quad \cdots \quad n_{i,N})$ is the complex additive white Gaussian noise (AWGN) at R_i with zero mean and identity covariance matrix. We assume the symbol vector from the source cannot be directly received by the destination. The same assumption has been made in many previous publications such as Refs. [1, 5—7]. Consider amplify-and-forward transmission at the relays. In the second stage, relay R_i first compensates for θ_i and obtains

$$\hat{\boldsymbol{r}}_i=\boldsymbol{r}_i e^{-j\theta_i}=\sqrt{P_sN}\,|f_i|\,\boldsymbol{x}+\boldsymbol{n}_i e^{-j\theta_i} \tag{2}$$

Then $\hat{\boldsymbol{r}}_i$ is amplified and encoded to get the symbol vector to be transmitted

$$\boldsymbol{t}_i=\sqrt{\frac{P_rT}{(1+P_s)N}}(\hat{\boldsymbol{r}}_i\boldsymbol{A}_i+\hat{\boldsymbol{r}}_i^*\boldsymbol{B}_i)=$$
$$\alpha\,|f_i|\,(\boldsymbol{x}\boldsymbol{A}_i+\boldsymbol{x}^*\boldsymbol{B}_i)+\beta(e^{-j\theta_i}\boldsymbol{n}_i\boldsymbol{A}_i+e^{j\theta_i}\boldsymbol{n}_i^*\boldsymbol{B}_i) \tag{3}$$

where $\boldsymbol{A}_i$ and $\boldsymbol{B}_i$ are $N\times T$ matrices known as the relay matrices or the associated matrices[8], $\alpha=\sqrt{P_rP_sT/(1+P_s)}$, and $\beta=\sqrt{P_rT/(1+P_s)N}$ is the amplifying coefficient at each relay, with P_r of the power per channel used at each relay in the second stage. As in many previous publications such as Refs. [1, 2, 11], β is used to ensure the power at each relay for the transmission in the second stage to be P_rT. Similar to the source-to-relay channels, the relay-to-destination channels are assumed to keep unchanged over T time slots. Thus the signal vector received at the destination is given by

$$\boldsymbol{r}_{\mathrm{D}}=\sum_{i=1}^{K}g_i\boldsymbol{t}_i+\boldsymbol{n}_{\mathrm{D}}=\alpha\sum_{i=1}^{K}|f_i|g_i(\boldsymbol{x}\boldsymbol{A}_i+\boldsymbol{x}^*\boldsymbol{B}_i)+\beta\sum_{i=1}^{K}g_i(\mathrm{e}^{-\mathrm{j}\theta_i}\boldsymbol{n}_i\boldsymbol{A}_i+\mathrm{e}^{\mathrm{j}\theta_i}\boldsymbol{n}_i^*\boldsymbol{B}_i)+\boldsymbol{n}_{\mathrm{D}}=\alpha(|f_1|g_1\quad|f_2|g_2\quad\cdots\quad|f_K|g_K)\begin{pmatrix}\boldsymbol{x}\boldsymbol{A}_1+\boldsymbol{x}^*\boldsymbol{B}_1\\\boldsymbol{x}\boldsymbol{A}_2+\boldsymbol{x}^*\boldsymbol{B}_2\\\vdots\\\boldsymbol{x}\boldsymbol{A}_K+\boldsymbol{x}^*\boldsymbol{B}_K\end{pmatrix}+\beta\sum_{i=1}^{K}g_i(\mathrm{e}^{-\mathrm{j}\theta_i}\boldsymbol{n}_i\boldsymbol{A}_i+\mathrm{e}^{\mathrm{j}\theta_i}\boldsymbol{n}_i^*\boldsymbol{B}_i)+\boldsymbol{n}_{\mathrm{D}}=\alpha\boldsymbol{h}\boldsymbol{X}+\boldsymbol{n}\tag{4}$$

where $\boldsymbol{X}=\begin{pmatrix}\boldsymbol{x}\boldsymbol{A}_1+\boldsymbol{x}^*\boldsymbol{B}_1\\\boldsymbol{x}\boldsymbol{A}_2+\boldsymbol{x}^*\boldsymbol{B}_2\\\vdots\\\boldsymbol{x}\boldsymbol{A}_K+\boldsymbol{x}^*\boldsymbol{B}_K\end{pmatrix}$ is the DOSTBC-CPI, $\boldsymbol{h}=(|f_1|g_1\quad|f_2|g_2\quad\cdots\quad|f_K|g_K)$, $\boldsymbol{n}_{\mathrm{D}}=(n_{\mathrm{D},1}\quad n_{\mathrm{D},2}\quad\cdots\quad n_{\mathrm{D},T})$ is the complex AWGN with zero mean and identity covariance matrix, and

$$\boldsymbol{n}=\beta\sum_{i=1}^{K}g_i(\mathrm{e}^{-\mathrm{j}\theta_i}\boldsymbol{n}_i\boldsymbol{A}_i+\mathrm{e}^{\mathrm{j}\theta_i}\boldsymbol{n}_i^*\boldsymbol{B}_i)+\boldsymbol{n}_{\mathrm{D}}$$

is the equivalent noise at the destination. Notice that $\boldsymbol{X}$ is a $K\times T$ matrix, where T is the block length of $\boldsymbol{X}$, and stands for the decoding delay (delay for short in the rest of this paper). Since $\boldsymbol{X}$ contains N information-bearing symbols $x_1, x_2, \cdots, x_N$ and lasts for T time slots, the rate of $\boldsymbol{X}$ is defined as $R=N/T$. It is easy to see that the mean of $\boldsymbol{n}$ is zero and the covariance matrix of $\boldsymbol{n}$ is given by

$$\boldsymbol{\Gamma}=E(\boldsymbol{n}^{\mathrm{H}}\boldsymbol{n})=\beta^2\sum_{i=1}^{K}|g_i|^2(\boldsymbol{A}_i^{\mathrm{H}}\boldsymbol{A}_i+\boldsymbol{B}_i^{\mathrm{H}}\boldsymbol{B}_i)+\boldsymbol{I}_T\tag{5}$$

The maximum likelihood (ML) estimate of the symbol vector $\boldsymbol{x}$ is given by

$$\hat{\boldsymbol{x}}=\arg\min_{\boldsymbol{x}\in C}(\boldsymbol{r}_{\mathrm{D}}-\alpha\boldsymbol{h}\boldsymbol{X})\boldsymbol{\Gamma}^{-1}(\boldsymbol{r}_{\mathrm{D}}-\alpha\boldsymbol{h}\boldsymbol{X})^{\mathrm{H}}=\arg\min_{\boldsymbol{x}\in C}(\alpha\boldsymbol{h}\boldsymbol{X}\boldsymbol{\Gamma}^{-1}\boldsymbol{X}^{\mathrm{H}}\boldsymbol{h}^{\mathrm{H}}-2\mathrm{Re}(\boldsymbol{h}\boldsymbol{X}\boldsymbol{\Gamma}^{-1}\boldsymbol{r}_{\mathrm{D}}^{\mathrm{H}}))\tag{6}$$

where C is the set containing all the possible symbol vector $\boldsymbol{x}$.

2 Distributed Orthogonal Space-Time Block Code with Channel Phase Information

Definition 1 (Yi and Kim[7]) A $K\times T$ complex matrix $\boldsymbol{X}$ on complex variables $x_1, x_2, \cdots, x_N$ is called a DOSTBC-CPI if the following conditions are satisfied:

(1) The entries of $\boldsymbol{X}$ are 0, $\pm x_i$, $\pm x_i^*$ or multiples of these indeterminates by $j=\sqrt{-1}$;

2) $\boldsymbol{X}$ satisfies the following equality

$$\boldsymbol{X}\boldsymbol{\Gamma}^{-1}\boldsymbol{X}^{\mathrm{H}}=\sum_{i=1}^{N}|x_i|^2\boldsymbol{D}_i\tag{7}$$

where $\boldsymbol{D}_i=\mathrm{diag}(D_{i,1}, D_{i,2}, \cdots, D_{i,K})$ and $D_{i,n}\neq 0$ for $1\leqslant i\leqslant N, 1\leqslant n\leqslant K$.

It is shown in Ref. [7] that the upper-bound on the rate of DOSTBC-CPI is exactly 1/2, and this upper-bound is achievable by the well-known rate-1/2 TJC code. Two examples of rate-1/2 DOSTBC-CPI are as follows[7]

$$\boldsymbol{X}=\begin{pmatrix}x_1&-x_2&0&0\\x_2^*&x_1^*&0&0\\0&0&x_1&x_2\end{pmatrix}\tag{8}$$

and

$$\boldsymbol{X}=\begin{pmatrix}x_1&-x_2&-x_3&-x_4&x_1^*&-x_2^*&-x_3^*&-x_4^*\\x_2&x_1&x_4&-x_3&x_2^*&x_1^*&x_4^*&-x_3^*\\x_3&-x_4&x_1&x_2&x_3^*&-x_4^*&x_1^*&x_2^*\\x_4&x_3&-x_2&x_1&x_4^*&x_3^*&-x_2^*&x_1^*\end{pmatrix}\tag{9}$$

The latter is a TJC code for four antennas[8]. However, as shown in Refs. [9, 10], the delay of TJC code is not optimal. In the rest of this paper, we will propose an algorithm to construct maximal rate DOSTBC-CPI with lower delay.

3 Construction of Maximal Rate DOSTBC- CPI with Lower Delay

Refs. [9, 10] independently proposed the rate-1/2, zero-entry-free and minimal delay CODs. However, the COD in Ref. [9] depends on a complicated iterative algorithm, while the COD in Ref. [10] is not suitable to be used as DOSTBC-CPI since it is easy to verify that $\boldsymbol{\Gamma}$ in this case is not diagonal, which makes the decoding much harder.

Therefore, we propose an algorithm to construct a class of rate 1/2 zero-entry-free and low delay CODs so that when the COD is used as a DOSTBC-CPI and the noise covariance matrix $\boldsymbol{\Gamma}$ is a scaled identity matrix. Due to space limitations, the detailed definition of balanced complex orthogonal design is illustrated in Definition 3.1 in Ref. [10]. The definition of column companion matrix is obtained by simply changing the "row" in Definition 4.1 in Ref. [10] into "column", and the definition of zero-masking column companion matrix is obtained by changing the "row" and "column" in Definition 4.2 in Ref. [10] into "column" and "row", respectively. The proofs of the following lemmas, theorem and corollary are omitted for the same reason.

Definition 2 In this paper, for any two different row (or column) vectors $\boldsymbol{x}$ and $\boldsymbol{y}$ of equal length L, if x_i and y_i satisfy $\begin{cases} y_i=0 & x_i=0 \\ y_i\neq 0 & x_i\neq 0 \end{cases}$, for $i=1, 2, \cdots, L$, then $\boldsymbol{x}$ and $\boldsymbol{y}$ are said to have identical zero patterns. On the contrary, if x_i and y_i satisfy $\begin{cases} y_i\neq 0 & x_i=0 \\ y_i=0 & x_i\neq 0 \end{cases}$, then $\boldsymbol{x}$ and $\boldsymbol{y}$ are said to have complementary zero patterns.

Lemma 1 Every column $\boldsymbol{l}$ of a $(2k, 2n, k)$ BCOD $\boldsymbol{P}$ has precisely one other partner column $\boldsymbol{l}'$ in $\boldsymbol{P}$ so that

(1) $\boldsymbol{l}$ and $\boldsymbol{l}'$ have complementary zero patterns;

(2) $\boldsymbol{l}$ and $\boldsymbol{l}'$ include the same variables;

(3) if any variable χ appears in $\boldsymbol{l}$, then it must appear as χ^* in $\boldsymbol{l}'$, and vice versa.

Such columns $\boldsymbol{l}$ and $\boldsymbol{l}'$ are called zero-masking partners.

Lemma 2 Any $(2k, 2n, k)$ BCOD $\boldsymbol{P}$ has a $(2k, 2n, k)$ zero-masking column companion matrix $\boldsymbol{Q}$ so that

(1) the ith column of $\boldsymbol{Q}$ is the zero-masking partner of the ith column of $\boldsymbol{P}$;

(2) the ith column of $\boldsymbol{Q}$ has the same variables as the ith column of $\boldsymbol{P}$;

(3) if any variable χ appears in the ith column of $\boldsymbol{P}$, then it must appear as χ^* in the ith column of $\boldsymbol{Q}$, and vice versa, for all $1\leqslant i\leqslant 2n$.

The idea of the following Theorem is similar to Theorem 5.2 of Ref. [10].

Theorem 1 Let $\boldsymbol{P}$ be a $(2k, 2n, k)$ BCOD with entries from $\{0, \pm x_1, \cdots, \pm x_k, \pm x_1^*, \cdots, \pm x_k^*\}$, and $\boldsymbol{Q}$ be the zero-masking column companion matrix of $\boldsymbol{P}$ as described in Lemma 2. Define a $2n\times 2n$ diagonal matrix $\boldsymbol{\Phi}$ whose (i, i)th entry is 1 if $(\boldsymbol{Q})_{1,i}$ is nonzero, and -1 if $(\boldsymbol{Q})_{1,i}$ is zero. Then $\boldsymbol{P}+\boldsymbol{Q}\boldsymbol{\Phi}$ is a rate 1/2 zero-entry-free $(2k, 2n, k)$ COD .

Like Theorem 5.2 of Ref. [10], Theorem 1 also provides an algorithm for constructing a class of rate-1/2, zero-entry-free CODs for any number of columns when followed by column deletion as necessary. Similar to the rate 1/2, zero-entry-free and minimal delay COD constructed in section V of Ref. [10], our design is also derived from BCODs that are generated via the modified-Liang algorithm in Ref. [10]. Hence following the method used in Ref. [10], it is not difficult to prove that the new rate-1/2 zero-entry-free CODs also have only half the delay of the TJC code and can achieve the lower bound on delay for most numbers of columns.

In addition, the property of not containing any zero entry not only avoids the need to switch on and off the antennas frequently, but also leads to low peak-to-average power ratio (PAPR)[10,12].

By transposing the rate-1/2 zero-entry-free CODs given above, we obtain a class of maximal rate DOSTBC-CPIs for cooperative wireless networks, which can achieve the lower bound on delay for most numbers of relays.

Corollary 1 When the rate-1/2 zero-entry-free COD constructed in Theorem 1 is used as a DOSTBC-CPI, the noise covariance matrix $\boldsymbol{\Gamma}$ is a scaled identity matrix.

Example Consider the BCOD for six antennas in Ref. [10] that is generated via the modified Liang algorithm

$$\boldsymbol{P}=\begin{pmatrix} x_1 & 0 & 0 & 0 & x_2 & x_3 \\ 0 & x_1 & 0 & -x_2 & 0 & x_4 \\ 0 & 0 & x_1 & -x_3 & -x_4 & 0 \\ 0 & x_2^* & x_3^* & x_1^* & 0 & 0 \\ -x_2^* & 0 & x_4^* & 0 & x_1^* & 0 \\ -x_3^* & -x_4^* & 0 & 0 & 0 & x_1^* \\ x_4 & -x_3 & x_2 & 0 & 0 & 0 \\ 0 & 0 & 0 & x_4^* & -x_3^* & x_2^* \end{pmatrix} \tag{10}$$

Its zero masking column companion matrix is

$$\boldsymbol{Q}=\begin{pmatrix} 0 & x_2 & x_3 & x_1 & 0 & 0 \\ -x_2 & 0 & x_4 & 0 & x_1 & 0 \\ -x_3 & -x_4 & 0 & 0 & 0 & x_1 \\ x_1^* & 0 & 0 & 0 & x_2^* & x_3^* \\ 0 & x_1^* & 0 & -x_2^* & 0 & x_4^* \\ 0 & 0 & x_1^* & -x_3^* & -x_4^* & 0 \\ 0 & 0 & 0 & x_4 & -x_3 & x_2 \\ x_4^* & -x_3^* & x_2^* & 0 & 0 & 0 \end{pmatrix} \tag{11}$$

$\boldsymbol{\Phi}=\mathrm{diag}(-1 \quad 1 \quad 1 \quad 1 \quad -1 \quad -1)$, thus we obtain a maximal rate DOSTBC-CPI for six relays as follows

$$\boldsymbol{X}=(\boldsymbol{P}+\boldsymbol{Q\Phi})^{\mathrm{T}}=$$

$$\begin{pmatrix} x_1 & x_2 & x_3 & -x_1^* & -x_2^* & -x_3^* & x_4 & -x_4^* \\ x_2 & x_1 & -x_4 & x_2^* & x_1^* & -x_4^* & -x_3 & -x_3^* \\ x_3 & x_4 & x_1 & x_3^* & x_4^* & x_1^* & x_2 & x_2^* \\ x_1 & -x_2 & -x_3 & x_1^* & -x_2^* & -x_3^* & x_4 & x_4^* \\ x_2 & -x_1 & -x_4 & -x_2^* & x_1^* & x_4^* & x_3 & -x_3^* \\ x_3 & x_4 & -x_1 & -x_3^* & -x_4^* & x_1^* & -x_2 & x_2^* \end{pmatrix}$$

It is easy to verify that the noise covariance matrix $\boldsymbol{\Gamma}$ is

$$\boldsymbol{\Gamma}=E(\boldsymbol{n}^{\mathrm{H}}\boldsymbol{n})=$$

$$\beta^2\sum_{i=1}^{6}|g_i|^2(\boldsymbol{A}_i^{\mathrm{H}}\boldsymbol{A}_i+\boldsymbol{B}_i^{\mathrm{H}}\boldsymbol{B}_i)+\boldsymbol{I}_8=$$

$$\left(1+\beta^2\sum_{i=1}^{6}|g_i|^2\right)\boldsymbol{I}_8 \tag{12}$$

The rate is 1/2 and the delay is $T=8$, which is the lower bound on delay for six relays (for the lower bound on decoding delay of rate-1/2 CODs, see Refs. [9, 10]).

For comparison, we list the TJC code for six relays as follows

$$\boldsymbol{X}=\begin{pmatrix} x_1 & -x_2 & -x_3 & -x_4 & -x_5 & -x_6 & -x_7 & -x_8 & x_1^* & -x_2^* & -x_3^* & -x_4^* & -x_5^* & -x_6^* & -x_7^* & -x_8^* \\ x_2 & x_1 & -x_4 & x_3 & -x_6 & x_5 & x_8 & -x_7 & x_2^* & x_1^* & -x_4^* & x_3^* & -x_6^* & x_5^* & x_8^* & -x_7^* \\ x_3 & x_4 & x_1 & -x_2 & -x_7 & -x_8 & x_5 & x_6 & x_3^* & x_4^* & x_1^* & -x_2^* & -x_7^* & -x_8^* & x_5^* & x_6^* \\ x_4 & -x_3 & x_2 & x_1 & -x_8 & x_7 & -x_6 & x_5 & x_4^* & -x_3^* & x_2^* & x_1^* & -x_8^* & x_7^* & -x_6^* & x_5^* \\ x_5 & x_6 & x_7 & x_8 & x_1 & -x_2 & -x_3 & -x_4 & x_5^* & x_6^* & x_7^* & x_8^* & x_1^* & -x_2^* & -x_3^* & -x_4^* \\ x_6 & -x_5 & x_8 & -x_7 & x_2 & x_1 & x_4 & -x_3 & x_6^* & -x_5^* & x_8^* & -x_7^* & x_2^* & x_1^* & x_4^* & -x_3^* \end{pmatrix} \tag{13}$$

The delay is $T=16$, which is twice that of the new code given above.

The DOSTBCs-CPI proposed in this paper have many desirable properties:

(1) They can achieve full diversity order and are single-symbol decodable since they are also CODs.

(2) They also achieve the rate 1/2, which is the upper-bound on the rate of DOSTBC-CPI.

(3) They have only half the delay of the TJC code, and can achieve the lower bound on delay for most numbers of relays like the rate 1/2 CODs in Refs. [9, 10].

(4) They have low PAPR since they do not contain any zero entry.

(5) They are obtained through a simple algebraic approach avoiding iterative algorithms.

4 Simulation Results

We consider a cooperative wireless network scenario in which one source node communicates with one destination node through the help of K relay nodes. The fading coefficients between the source and the relays f_i, and between the relays and the destination g_i, are modeled as independent complex Gaussian random variables with zero-mean and unit-variance. f_i and g_i are assumed to keep constant for the entire transmission of one code block. The noises at the relays and the destination are modeled as independent zero-mean unit-variance Gaussian additive noise. We compare the performance of the proposed code and the TJC code in terms of SER versus P_s, the power used by the source node per channel use. The optimum power allocation proposed in Ref. [1] is adopted, i. e. $P_r = P_s / K$. Since the proposed code and the TJC code both have rate 1/2, they employ the same signal sets to achieve the same bandwidth efficiency. Hence QPSK, 8PSK and 16QAM are employed for both codes to achieve bandwidth efficiency of 1 b/(s · Hz), 3/2 b/(s · Hz) and 2 b/(s · Hz), respectively. The symbol vectors are decoded symbol by symbol via ML estimation algorithm given in Eq. (6). The SER performance is obtained by source sending 10^7 symbols at a time.

Fig. 2 compares the SER performance of the proposed code and the TJC code for six relays given in Eq. (13). Fig. 3 shows the SER performance comparison of the proposed code and the TJC code for four relays. As we have expected, the proposed code provides full diversity, for the same bandwidth efficiency, the SER curve of the proposed code moves parallel to that of the TJC code. Also as expected, the proposed code achieves the same diversity order for different modulation schemes, since at large values of P_s, the SER curves of the proposed code for different modulation schemes are parallel to each other. The reason for these is that the diversity order that a DOSTBC can provide in high SNR region is determined by the number of relays. It is also

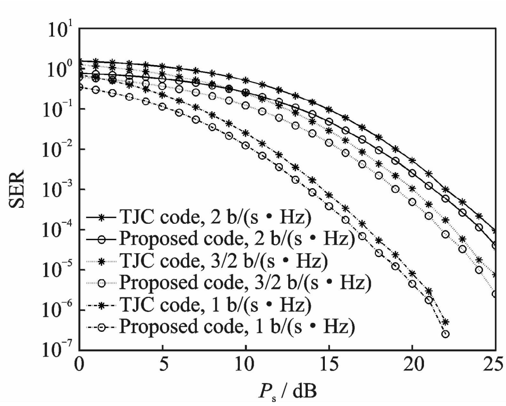

Fig. 2 SER performance comparison of the proposed code and TJC code for $K=6$

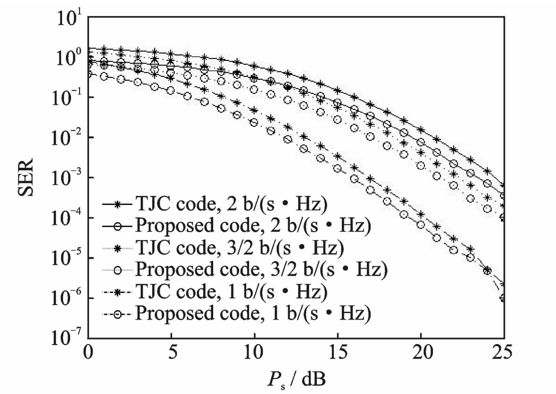

Fig. 3 SER performance comparison of the proposed code and the TJC code for $K=4$

observed that for same bandwidth efficiency, the proposed code outperforms the TJC code by about 1—2 dB. Therefore, the proposed code outperforms the TJC code in SER performance as well as in decoding delay.

5 Conclusions

We propose an algorithm to construct a class of maximal rate DOSTBC-CPIs with half the decoding delay of the TJC code. The proposed method not only hold all the desirable properties of the TJC code such as full diversity order, maximal rate, single-symbol decodable, low PAPR, noise covariance matrix being a scaled identity matrix and obtained through a simple algebraic approach, but also outperform the latter in SER performance as well as in decoding delay.

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