PAPER

Space-Time Block Coded Joint Transmit/Receive Diversity in a Frequency-Nonselective Rayleigh Fading Channel

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SUMMARY Antenna diversity is an effective technique for improving the transmission performance in a multi-path fading channel. Recently, transmit diversity has been attracting much attention since it can alleviate the complexity problem of the mobile terminals. Joint transmit diversity/receive diversity achieves a much improved transmission performance. In this paper, we propose a new space-time block coding algorithm for joint transmit/receive diversity, which requires the channel state information (CSI) only at the transmitter side. Unlike the conventional space-time transmit diversity (STTD), the space-time block coded joint transmit/receive diversity (STBC-JTRD) can use arbitrary number of transmit antennas, while the number of receive antennas is limited to 4. STBC-JTRD achieves a larger diversity gain than joint STTD/receive antenna diversity. The bit error rate (BER) analysis in a frequency-nonselective Rayleigh fading channel is presented. The BER performance is evaluated and is confirmed by the computer simulation.
key words: *antenna diversity, space-time block coding, time-division duplex*

1. Introduction

In mobile radio, the BER performance seriously degrades due to multi-path fading [1], [2]. Antenna diversity is a well-known technique for improving the transmission performance in a multi-path fading channel [1], [2]. Recently, transmit antenna diversity has been attracting much attention because the complexity problem of a mobile terminal can be alleviated [3]–[10]. Transmit diversity techniques are roughly classified into two types: the first type requires the channel state information (CSI) while the 2nd type requires no CSI. Space-time block coded transmit diversity (STTD) [3]–[7] belongs to the 2nd type. STTD can be jointly used with the receive antenna diversity of an arbitrary number of antennas; but, when more than 3 transmit antennas are used, the STTD coding rate is reduced to less than 3/4. For example, the STTD with 3/4 (2/3) coding rate was presented for 3 or 4 (5 or 6) transmit antennas [4]–[7]. Maximal ratio transmit (MRT) diversity technique [9] belongs to the first type. In MRT, each transmit signal from a different antenna is multiplied by a complex transmit weight so that the maximal ratio combining (MRC) diversity gain is obtained at a receiver side. It is shown in Ref. [9] that for the same transmit bit energy-to-noise power spectrum density ratio (E_b/N_0), MRT using N transmit antennas achieves the same bit error rate (BER) performance as MRC receive diversity using N receive antennas. To further improve the BER performance, joint use of MRT and receive antenna diversity can be used [10], [11]. However, the CSI is required at both transmitter and receiver sides.

In this paper, we propose a new space-time block coding algorithm for joint transmit/receive diversity, called STBC-JTRD, which requires the CSI only at the transmitter side (in the case of time division duplex (TDD), since the same carrier frequency is used for both transmit/receive channels [12], the transmit channel CSI can be relatively easily estimated using the received signal). In STBC-JTRD, an arbitrary number of transmit antennas is allowed, while the number of receive antennas is limited to 4. STBC-JTRD is suitable for the downlink (base-to-mobile) applications since more antennas can be provided at a base station than at a mobile station for the given total number of antennas.

The remainder of this paper is organized as follows. Section 2 describes the principle of STBC-JTRD. The BER analysis in a frequency-nonselective Rayleigh fading channel is presented in Sect. 3. In Sect. 4, the theoretical BER performance is evaluated and is confirmed by the computer simulation. Section 5 offers some conclusions.

2. STBC-JTRD

Figure 1 illustrates the transmitter and receiver structure of the proposed (N_t, N_r) STBC-JTRD with N_t transmit antennas and N_r receive antennas. An information symbol sequence to be transmitted is grouped into a sequence of blocks of K symbols each. Each block is encoded into N_t parallel codewords; each codewords consists of I symbols

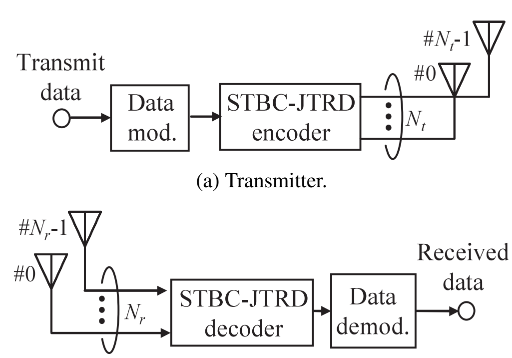

Fig. 1 Transmitter and receiver structure of STBC-JTRD.

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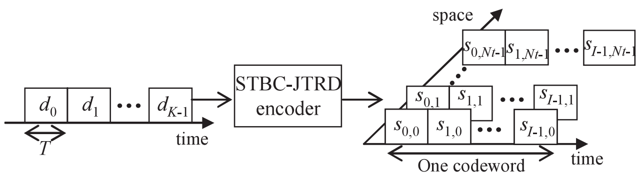


Fig. 2 STBC-JTRD encoding.

Table 1 K, I and R for N_r = 2–4.

No. of receive antennas, N_r	No. of information symbols in a codeword, K	No. of coded symbols in a codeword, I	Coding rate, R
2	2	2	1
3	3	4	3/4
4	3	4	3/4

(see Fig. 2), and is transmitted from one of N_t transmit antennas. Table 1 shows the number K of information symbols in a codeword, the number I of coded symbols in a codeword, and coding rate R for N_r=2, 3 and 4.

2.1 Encoding

The k-th information symbol in a codeword is denoted by d_k and the i-th coded symbol to be transmitted from the n-th transmit antenna is denoted by $s_{i,n}$. STBC-JTRD encoding is expressed as

$$\begin{pmatrix} \boldsymbol{s}_0 \\ \boldsymbol{s}_1 \end{pmatrix} = \sqrt{\frac{2S}{C_2}} \begin{pmatrix} d_0\boldsymbol{h}_0^* + d_1\boldsymbol{h}_1^* \\ d_0^*\boldsymbol{h}_1^* - d_1^*\boldsymbol{h}_0^* \end{pmatrix} \text{ for } N_r = 2, \quad (1a)$$

$$\begin{pmatrix} \boldsymbol{s}_0 \\ \boldsymbol{s}_1 \\ \boldsymbol{s}_2 \\ \boldsymbol{s}_3 \end{pmatrix} = \sqrt{\frac{2S}{C_3}} \begin{pmatrix} d_0\boldsymbol{h}_0^* + d_1\boldsymbol{h}_1^* + d_2\boldsymbol{h}_2^* \\ d_0^*\boldsymbol{h}_1^* - d_1^*\boldsymbol{h}_0^* \\ d_0^*\boldsymbol{h}_2^* - d_2^*\boldsymbol{h}_0^* \\ d_1^*\boldsymbol{h}_2^* - d_2^*\boldsymbol{h}_1^* \end{pmatrix} \text{ for } N_r = 3, \quad (1b)$$

$$\begin{pmatrix} \boldsymbol{s}_0 \\ \boldsymbol{s}_1 \\ \boldsymbol{s}_2 \\ \boldsymbol{s}_3 \end{pmatrix} = \sqrt{\frac{2S}{C_4}} \begin{pmatrix} d_0\boldsymbol{h}_0^* + d_1\boldsymbol{h}_1^* + d_2\boldsymbol{h}_2^* \\ d_0^*\boldsymbol{h}_1^* - d_1^*\boldsymbol{h}_0^* + d_2\boldsymbol{h}_3^* \\ d_0^*\boldsymbol{h}_2^* - d_1\boldsymbol{h}_3^* - d_2^*\boldsymbol{h}_0^* \\ d_0\boldsymbol{h}_3^* + d_1^*\boldsymbol{h}_2^* - d_2^*\boldsymbol{h}_1^* \end{pmatrix} \text{ for } N_r = 4, \quad (1c)$$

where $\boldsymbol{s}_i = [s_{i,0}, s_{i,1}, \ldots, s_{i,Nt-1}]^T$ is the signal vector to be transmitted at the i-th time slot, $\boldsymbol{h}_m = [h_{m,0}, h_{m,1}, \ldots, h_{m,Nt-1}]^T$ is the channel gain vector with $h_{m,n}$ representing the complexed-value channel gain between the n-th transmit antenna ($n = 0 \sim (N_t - 1)$) and the m-th receive antenna ($m = 0 \sim (N_r - 1)$), S denotes the average total transmit power, and C_{Nr} is the power normalization factor, given by

$$C_{N_r} = \sum_{n=0}^{N_t-1} \sum_{m=0}^{N_r-1} |h_{m,n}|^2. \quad (2)$$

2.2 Decoding

The transmitted signals go through different fading channels and are received by N_r receive antennas. In this paper, the block fading is assumed. The i-th symbol in a codeword received by the m-th receive antenna is denoted by $r_{i,m}$, which can be expressed as

$$\boldsymbol{r}_i = \boldsymbol{H}\boldsymbol{s}_i + \boldsymbol{\eta}_i, \quad i = 0 \sim I - 1, \quad (3)$$

where $\boldsymbol{r}_i = [r_{i,0}, r_{i,1}, \ldots, r_{i,Nr-1}]^T$ and $\boldsymbol{H} = [\boldsymbol{h}_0, \boldsymbol{h}_1, \ldots, \boldsymbol{h}_{Nr-1}]^T$. $\boldsymbol{\eta}_i = [\eta_{i,0}, \eta_{i,1}, \ldots, \eta_{i,Nr-1}]^T$ is the noise vector, where $\eta_{i,m}$ denotes the noise component due to the additive white Gaussian noise (AWGN) with zero mean and variance $2N_0/T$ with N_0 being the single-sided power spectrum density and T being the transmit symbol period.

STBC-JTRD decoding is carried out on $\{r_{i,m};\ i = 0 \sim (I-1),\ m = 0 \sim (N_r - 1)\}$ to obtain the decision variables $\{\hat{d}_k;\ k = 0 \sim (K-1)\}$ as follows:

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \end{pmatrix} = \begin{pmatrix} r_{0,0} + r_{1,1}^* \\ r_{0,1} - r_{1,0}^* \end{pmatrix} \text{ for } N_r = 2, \quad (4a)$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \begin{pmatrix} r_{0,0} + r_{1,1}^* + r_{2,2}^* \\ r_{0,1} - r_{1,0}^* + r_{3,2}^* \\ r_{0,2} - r_{2,0}^* - r_{3,1}^* \end{pmatrix} \text{ for } N_r = 3, \quad (4b)$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \begin{pmatrix} r_{0,0} + r_{1,1}^* + r_{2,2}^* + r_{3,3} \\ r_{0,1} - r_{1,0}^* - r_{2,3} + r_{3,2}^* \\ r_{0,2} + r_{1,3} - r_{2,0}^* - r_{3,1}^* \end{pmatrix} \text{ for } N_r = 4. \quad (4c)$$

Substituting Eqs. (1) and (3) into Eq. (4), we obtain

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \end{pmatrix} = \sqrt{2SC_2} \begin{pmatrix} d_0 \\ d_1 \end{pmatrix} + \begin{pmatrix} \eta_{0,0} + \eta_{1,1}^* \\ \eta_{0,1} - \eta_{1,0}^* \end{pmatrix} \text{ for } N_r = 2, \quad (5a)$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \sqrt{2SC_3} \begin{pmatrix} d_0 \\ d_1 \\ d_2 \end{pmatrix} + \begin{pmatrix} \eta_{0,0} + \eta_{1,1}^* + \eta_{2,2}^* \\ \eta_{0,1} - \eta_{1,0}^* + \eta_{3,2}^* \\ \eta_{0,2} - \eta_{2,0}^* - \eta_{3,1}^* \end{pmatrix} \text{ for } N_r = 3, \quad (5b)$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \sqrt{2SC_4} \begin{pmatrix} d_0 \\ d_1 \\ d_2 \end{pmatrix} + \begin{pmatrix} \eta_{0,0} + \eta_{1,1}^* + \eta_{2,2}^* + \eta_{3,3} \\ \eta_{0,1} - \eta_{1,0}^* - \eta_{2,3} + \eta_{3,2}^* \\ \eta_{0,2} + \eta_{1,3} - \eta_{2,0}^* - \eta_{3,1}^* \end{pmatrix} \text{ for } N_r = 4. \quad (5c)$$

Equation (5) indicates that the maximal-ratio combining (MRC)-receive diversity [1], with equivalent number of antennas equal to $N_t \times N_r$, can be achieved, but the noise power is increased by N_r times.

2.3 Comparison of STBC-JTRD and STTD

Comparison of STBC-JTRD and STTD is summarized in Table 2. An advantage of STBC-JTRD is that an arbitrary number of transmit antennas can be used while the number N_r of receive antennas is limited to 4. This is suitable for the downlink (base-to-mobile) applications since most of the antennas can be provided at the base station for the given total number of antennas. When N_r=2, the coding rate is R=1. However, when N_r=3 and 4, the coding rate reduces to 3/4. On the other hand, if STTD is used, only N_t=2 antennas can be equipped at the base station for coding rate R=1 and other antennas must be implemented at a mobile station (note that

Table 2 Comparison of STBC-JTRD and STTD.

Diversity scheme	No. of transmit antennas, N_t	No. of receive antennas, N_r	CSI required at	Coding rate, R
STBC -JTRD	Arbitrary	2	Transmitter side	1
		3		3/4
		4		3/4
STTD [3-7]	2	Arbitrary	Receiver side	1
	3			3/4
	4			3/4
	5			2/3
	6			2/3
	7			5/8

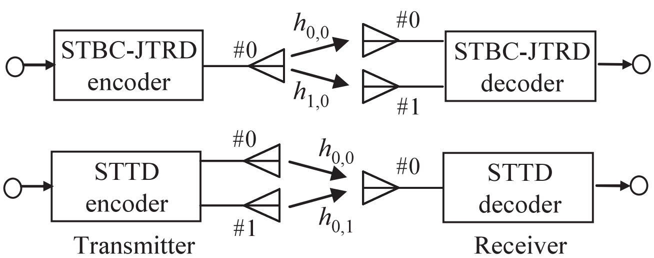


Fig. 3 (1, 2) STBC-JTRD and (2, 1) STTD.

although N_t=3 or 4, 5 or 6 and 7 transmit antennas can be used at the base station, the coding rate reduces to R=3/4, 2/3 and 5/8, respectively [5]–[7]).

In STBC-JTRD, the CSI is required for encoding, but it is not for decoding. On the contrary, STTD requires no CSI for encoding, but requires it for decoding. It is quite difficult to mathematically show the relationship in the encoding/decoding algorithms between STBC-JTRD and STTD. However, we can show the relationship between (1,2) STBC-JTRD and (2,1) STTD (see Fig. 3). For simplicity, the noise is omitted. The STTD encoding/decoding algorithm is presented in Appendix. The (2,1) STTD transforms the fading channel into the equivalent channel characterized by $\tilde{\boldsymbol{H}} = \begin{pmatrix} h_{0,0} & h_{0,1} \\ h_{0,1}^* & -h_{0,0}^* \end{pmatrix}$ (which can be obtained from STTD encoding without CSI given by Eq. (A· 1a), and the conjugate operation on $r_{1,0}$ as shown in Eq. (6)). The i-th received STTD encoded symbol $r_{i,0}$ (i=0, 1) is given, from Eqs. (A· 1) and (A· 2), as

$$\begin{pmatrix} r_{0,0} \\ r_{1,0}^* \end{pmatrix} = \sqrt{S}\tilde{\boldsymbol{H}}\boldsymbol{d}, \tag{6}$$

where $\boldsymbol{d} = [d_0, d_1]^T$. At the receiver, the (2,1) STTD decoding is carried out, by using the decoding matrix $\boldsymbol{D} = \tilde{\boldsymbol{H}}^H$, as

$$\hat{\boldsymbol{d}} = \boldsymbol{D} \cdot \left(\sqrt{S}\tilde{\boldsymbol{H}}\boldsymbol{d}\right) = \sqrt{S}\boldsymbol{\Lambda}\boldsymbol{d}, \tag{7}$$

where $\hat{\boldsymbol{d}} = [\hat{d}_0, \hat{d}_1]^T$, $(\cdot)^H$ denotes the Hermitian transposition and

$$\boldsymbol{\Lambda} = diag\{|h_{0,0}|^2 + |h_{0,1}|^2, |h_{0,0}|^2 + |h_{0,1}|^2\}. \tag{8}$$

On the other hand, the (1,2) STBC-JTRD decoding without CSI given as Eq. (4a) transforms the fading channel into the equivalent channel characterized by $\bar{\boldsymbol{H}} = \begin{pmatrix} h_{0,0} & h_{1,0}^* \\ h_{1,0} & -h_{0,0}^* \end{pmatrix}$. The (1,2) STBC-JTRD encoding is performed using the encoding matrix $\bar{\boldsymbol{D}}^T = \bar{\boldsymbol{H}}^H$, from Eq. (1a), as

$$\begin{pmatrix} s_{0,0} \\ s_{1,0}^* \end{pmatrix} = \sqrt{\frac{2S}{C_2}}\bar{\boldsymbol{D}}^T\boldsymbol{d}. \tag{9}$$

The encoded signal goes through the equivalent channel $\bar{\boldsymbol{H}}$ and is received by two antennas at the receiver. The decision variable $\hat{\boldsymbol{d}} = [\hat{d}_0, \hat{d}_1]^T$ after decoding can be expressed as

$$\hat{\boldsymbol{d}} = \bar{\boldsymbol{H}} \cdot \left(\sqrt{\frac{2S}{C_2}}\bar{\boldsymbol{D}}^T\boldsymbol{d}\right) = \sqrt{\frac{2S}{C_2}}\bar{\boldsymbol{\Lambda}}\boldsymbol{d}, \tag{10}$$

where

$$\bar{\boldsymbol{\Lambda}} = diag\{|h_{0,0}|^2 + |h_{1,0}|^2, |h_{0,0}|^2 + |h_{1,0}|^2\}. \tag{11}$$

Comparison of Eqs. (7) and (10) shows that (1, 2) STBC-JTRD uses the STTD decoding matrix $\boldsymbol{D}$ but with $h_{0,n}$ replaced by $h_{m,0}$ as the encoding matrix $\bar{\boldsymbol{D}}$.

3. BER Analysis

A theoretical BER expression for (N_t, N_r) STBC-JTRD is derived in a frequency-nonselective Rayleigh fading channel. Quaternary phase shift keying (QPSK) data-modulation and the data transmission of "1" are assumed without loss of generality.

The average BER for the given set of $\boldsymbol{H} = [\boldsymbol{h}_0, \boldsymbol{h}_1, \ldots, \boldsymbol{h}_{Nr-1}]^T$ is given by [2]

$$P_b\left(\frac{E_s}{N_0}, \boldsymbol{H}\right) = \frac{1}{2}erfc\left[\sqrt{\frac{1}{4}\gamma\left(\frac{E_s}{N_0}, \boldsymbol{H}\right)}\right], \tag{12}$$

where $erfc[x] = (2/\sqrt{\pi})\int_x^\infty \exp(-t^2)dt$ is the complementary error function, $E_s = ST$ is the transmit symbol energy and $\gamma(E_s/N_0, \boldsymbol{H})$ is the conditional signal-to-noise power ratio (SNR) after decoding. Since $\{\eta_{i,m}; m = 0 \sim (N_r - 1)\}$ are independent zero-mean complex Gaussian random variables and from Eqs. (2) and (5), $\gamma(E_s/N_0, \boldsymbol{H})$ is given by

$$\gamma\left(\frac{E_s}{N_0}, \boldsymbol{H}\right) = 2\left(\frac{1}{N_r}\frac{E_s}{N_0}\right)\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}|h_{m,n}|^2. \tag{13}$$

For the BER analysis, we rewrite Eq. (13) as

$$\gamma\left(\frac{E_s}{N_0}, \boldsymbol{H}\right) = 2\frac{\Gamma}{N_r}\sum_{l=0}^{L-1}|h_l|^2, \tag{14}$$

where $L = N_t \times N_r$, $\Gamma = E_s/N_0$ is the average transmit SNR, and $\{h_l; l = 0 \sim (L-1)\}$ are independent zero-mean complex-valued Gaussian variables with unity variance. The average BER is obtained from

$$P_b\left(\frac{E_s}{N_0}\right) = \int_0^\infty \frac{1}{2}erfc\left(\sqrt{\frac{1}{4}\gamma}\right)p(\gamma)d\gamma, \tag{15}$$

where $p(\gamma)$ is the pdf of $\gamma(E_s/N_0, \boldsymbol{H})$ and is given by [2]

$$p(\gamma) = \frac{N_r^L}{(L-1)!\Gamma^L}\left(\frac{\gamma}{2}\right)^{L-1}\exp\left(-\frac{\gamma}{2}\frac{N_r}{\Gamma}\right). \tag{16}$$

Thus, the average BER is given as [2]

$$P_b\left(\frac{E_s}{N_0}\right) = \left[\frac{1}{2}\left(1-\sqrt{\frac{\Gamma/N_r}{2+\Gamma/N_r}}\right)\right]^L \times\sum_{k=0}^{L-1}\begin{pmatrix} L-1+k \\ k \end{pmatrix}\left[\frac{1}{2}\left(1+\sqrt{\frac{\Gamma/N_r}{2+\Gamma/N_r}}\right)\right]^k, \tag{17}$$

where $\begin{pmatrix} a \\ b \end{pmatrix}$ is the binomial coefficient. An approximate BER expression for $\Gamma \gg 1$ can be obtained, from Eq. (17), as [2]

$$P_b\left(\frac{E_s}{N_0}\right) \approx \frac{1}{2^L}\left(\frac{\Gamma}{N_r}\right)^{-L}\begin{pmatrix} 2L-1 \\ L \end{pmatrix}, \tag{18}$$

which indicates that (N_t, N_r) STBC-JTRD can achieve the $N_t \times N_r$-branch MRC receive antenna diversity effect, but with the reduced SNR by a factor of $1/N_r$.

In Appendix, the average BER for STTD is derived and is given by

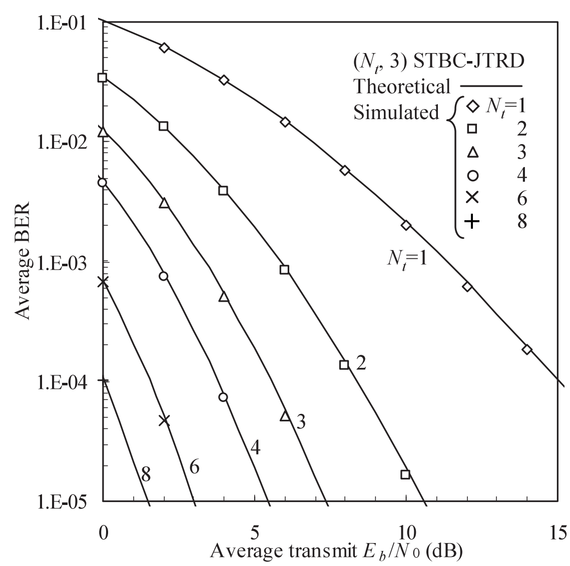
$$\begin{aligned} P_b\left(\frac{E_s}{N_0}\right) &= \left[\frac{1}{2}\left(1-\sqrt{\frac{\Gamma/N_t}{2+\Gamma/N_t}}\right)\right]^L \times\sum_{k=0}^{L-1}\begin{pmatrix} L-1+k \\ k \end{pmatrix}\left[\frac{1}{2}\left(1+\sqrt{\frac{\Gamma/N_t}{2+\Gamma/N_t}}\right)\right]^k \\ &\approx \frac{1}{2^L}\left(\frac{\Gamma}{N_t}\right)^{-L}\begin{pmatrix} 2L-1 \\ L \end{pmatrix} \text{ for } \Gamma \gg 1. \end{aligned} \tag{19}$$

4. Numerical and Computer Simulation Results

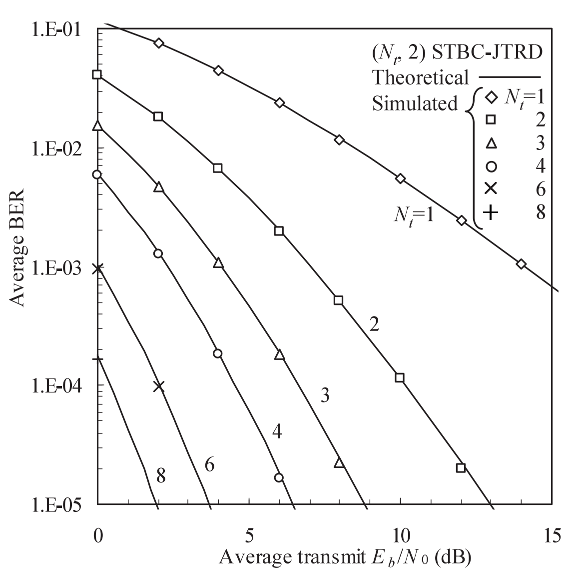
The theoretical BER performance of STBC-JTRD is evaluated and is confirmed by computer simulation. Table 3 summarizes the numerical and simulation conditions. QPSK data-modulation and ideal channel estimation are assumed.

Table 3 Numerical and simulation conditions.

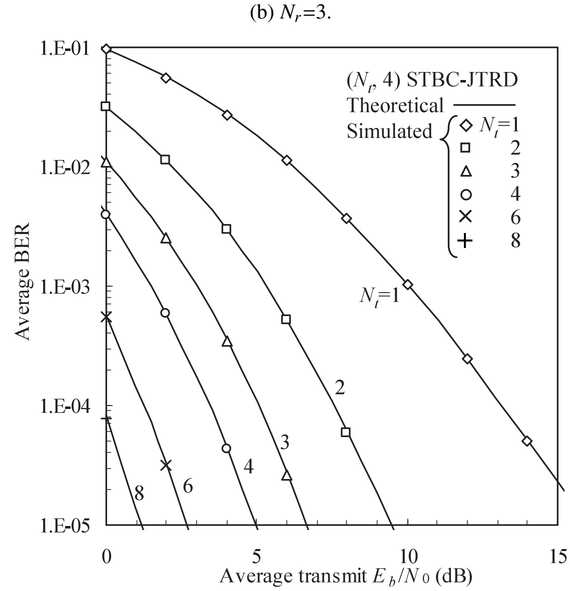
Data modulation	QPSK
No. of transmit antennas	N_t=1~8
Channel model	Frequency-nonselective Rayleigh fading channel
No. of received antennas	N_r=1~4
Channel estimation	Ideal



(b) N_r=3.



(a) N_r=2.



(c) N_r=4.

Fig. 4 Average BER performance of (N_t, N_r) STBC-JTRD.

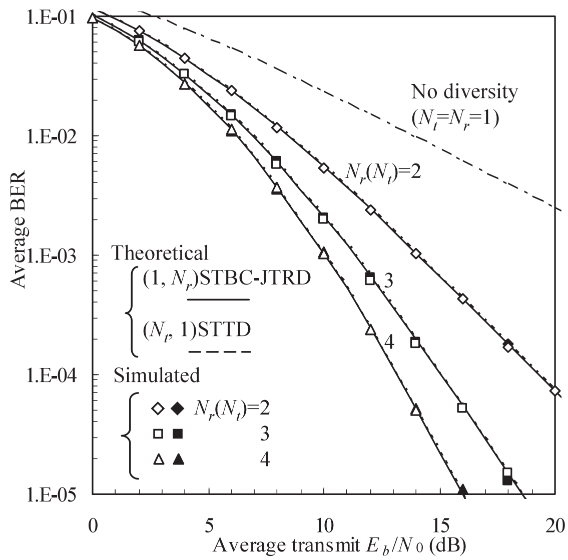


Fig. 5 Performance comparison between (1, N_r) STBC-JTRD and (N_t, 1) STTD.

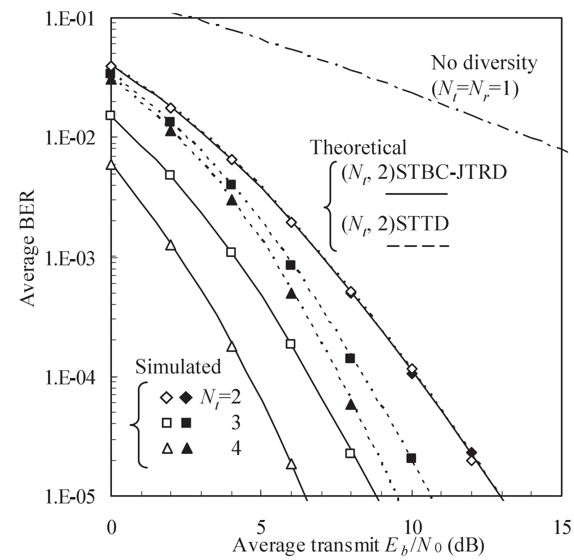


Fig. 6 Performance comparison of (N_t, N_r=2) STBC-JTRD and (N_t, N_r=2) STTD.

For comparison, we also evaluate the BER performance of STTD [3], [4], which requires the CSI at the receiver side only, and also those of MRT [10], the improved MRT (IMRT) [11] and the maximal ratio transmit and combining (MRTC) [11]. MRT, IMRT and MRTC require the CSI at both the transmitter and receiver sides.

Figure 4 plots the theoretical and simulated average BER performances of (N_t, N_r) STBC-JTRD as a function of the average transmit signal energy per bit-to-AWGN power spectrum density ratio E_b/N_0 (=0.5 (E_s/N_0)). A fairly good agreement between the theoretical and simulated results is seen. As the number of N_t of transmit antennas increase, the average BER performance consistently improves for the given number of N_r of receive antennas. When N_r=2, the required transmit E_b/N_0 for the BER=10^{-3} can be reduced by about 7, 3 and 2 dB by increasing N_t to 2, 3 and 4, respectively. Let's consider the case of the total number of transmit and receive antennas is kept the same. An interesting question is which provides better BER performance, increasing the number N_t transmit antennas or increasing the number N_r of receive antennas? It can be seen from Figs. 4(a)–(c) that, as N_r increases for the given N_t, the BER performance consistently improves; but when N_t=2, the reduction in the required E_b/N_0 for the BER=10^{-3} is about 4, 1.2 and 0.4 dB for N_r=2, 3 and 4, respectively. This clearly shows the advantage of our proposed STBC-JTRD which allows the use of an arbitrary number of transmit antennas. Figure 5 plots the theoretical and simulated average BER performances achievable with (1, N_r) STBC-JTRD and (N_t, 1) STTD as a function of the average transmit E_b/N_0. It is seen that STBC-JTRD can achieve the same BER performance as STTD since the instantaneous received SNR is the same for both schemes (see Eqs. (13) and (A·5)).

Figure 6 shows the theoretical and simulated average BER performances as a function of the average transmit E_b/N_0 with N_t as a parameter when N_r=2. When N_t=2, STBC-JTRD is equivalent to STTD. However, when N_t=3 and 4, STBC-JTRD achieves better BER performance than STTD. Note that when N_r=2, although STTD reduces the transmission data-rate to 3/4, STBC-JTRD does not at all. Again a good agreement between the theoretical and the simulation results is observed in Figs. 5 and 6.

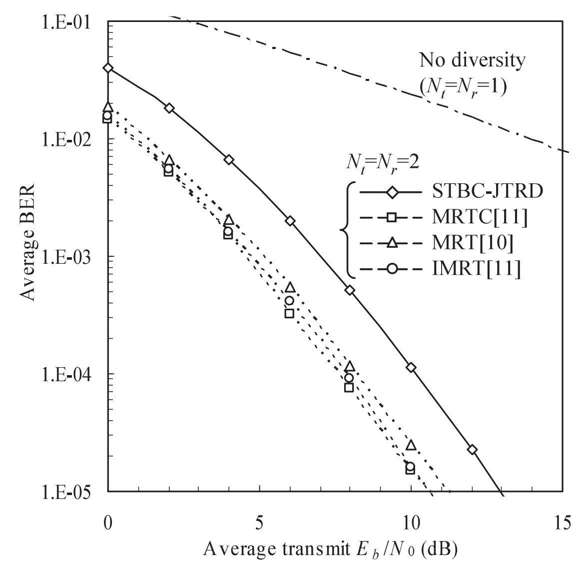


Fig. 7 Performance comparison between (2,2) STBC-JTRD and various (2, 2) transmit/receive diversity schemes.

Figure 7 compares the simulated average BER performances of STBC-JTRD, MRT [10], IMRT and MRTC [11]. It is seen from Fig. 7 that MRTC can provide the best BER performance among four diversity schemes. STBC-JTRD requires about 2.5 dB larger transmit E_b/N_0 for BER=10^{-3} than MRTC. This is because MRTC can coherently combine the received signals using CSI and achieve the $N_t \times N_r$-

branch MRC receive diversity gain, while STBC-JTRD combines the received signals without CSI and therefore, the SNR after decoding is smaller by a factor of $1/N_r$ compared to $N_t \times N_r$-branch MRC receive antenna diversity. However, it should be noted that the receiver complexity of STBC-JTRD is less than MRTC since STBC-JTRD requires the CSI at the transmitter side only.

5. Conclusion

In this paper, we proposed a new space-time block coding algorithm for joint transmit/receive antenna diversity, called STBC-JTRD, which requires channel state information (CSI) only at a transmitter side. An arbitrary number of transmit antennas can be used without reducing the coding rate when 2 receive antennas can be used; however the coding rate reduces to 3/4 when 3 or 4 receive antennas. On the other hand, the well-known STTD reduces the transmission data-rate if more than 3 transmit antennas are employed although an arbitrary number of receive antennas can be used. STBC-JTRD requires the channel state information (CSI) at the transmitter side, while STTD requires the CSI at the receiver side.

The theoretical average BER expression was theoretically derived in a frequency-nonselective Rayleigh fading. A fairly good agreement between the theoretical and simulated results is seen. It was shown that (1, N_r) STBC-JTRD and (N_t, 1) STTD are equivalent since the instantaneous received SNR is the same for both schemes. However, when N_t=3 and 4, (N_t, 2) STBC-JTRD achieves better BER performance than (N_t, 2) STTD, without sacrificing the transmission data-rate at all.

In this paper, we presented the STBC-JTRD coding schemes of up to 4 receive antennas. STTD using $N_t > 5$ transmit antennas was presented in [6], [7]. It is an open problem to find STBC-JTRD with $N_r > 5$ receive antennas.

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Appendix: BER Analysis of STTD

STTD encoding is expressed as [3]–[6]

$$\begin{pmatrix} \bar{\boldsymbol{s}}_0^T \\ \bar{\boldsymbol{s}}_1^T \end{pmatrix} = \sqrt{S} \begin{pmatrix} d_0 & d_1 \\ -d_1^* & d_0^* \end{pmatrix} \text{ for } N_t = 2, \tag{A· 1a}$$

$$\begin{pmatrix} \bar{\boldsymbol{s}}_0^T \\ \bar{\boldsymbol{s}}_1^T \\ \bar{\boldsymbol{s}}_2^T \\ \bar{\boldsymbol{s}}_3^T \end{pmatrix} = \sqrt{\frac{2S}{3}} \begin{pmatrix} d_0 & d_1 & d_2 \\ -d_1^* & d_0^* & 0 \\ -d_2^* & 0 & d_0^* \\ 0 & -d_2^* & d_1^* \end{pmatrix} \text{ for } N_t = 3, \tag{A· 1b}$$

$$\begin{pmatrix} \bar{\boldsymbol{s}}_0^T \\ \bar{\boldsymbol{s}}_1^T \\ \bar{\boldsymbol{s}}_2^T \\ \bar{\boldsymbol{s}}_3^T \end{pmatrix} = \sqrt{\frac{S}{2}} \begin{pmatrix} d_0 & d_1 & d_2 & 0 \\ -d_1^* & d_0^* & 0 & d_2 \\ -d_2^* & 0 & d_0^* & -d_1 \\ 0 & -d_2^* & d_1^* & d_0 \end{pmatrix} \text{ for } N_t = 4, \tag{A· 1c}$$

where $\bar{\boldsymbol{s}}_i = [\bar{s}_{i,0}, \bar{s}_{i,1}, \ldots, \bar{s}_{i,N_t-1}]^T$. The n-th column and i-th row element of Eq. (A· 1) represents the transmit signal from the n-th transmit antenna during the i-th time slot.

N_r received signals associated with the i-th transmitted symbol are represented by $\boldsymbol{r}_i = [r_{i,0}, r_{i,1}, \ldots, r_{i,Nr-1}]^T$. $\boldsymbol{r}_i$ can be expressed as

$$\boldsymbol{r}_i = \bar{\boldsymbol{H}}^T \bar{\boldsymbol{s}}_i + \boldsymbol{\eta}_i, \quad i = 0 \sim I - 1, \tag{A· 2}$$

where $\bar{\boldsymbol{H}} = [\bar{\boldsymbol{h}}_0, \bar{\boldsymbol{h}}_1, \ldots, \bar{\boldsymbol{h}}_{N_t-1}]^T$ and $\bar{\boldsymbol{h}}_n = [h_{0,n}, h_{1,n}, \ldots, h_{N_r-1,n}]^T$. $\boldsymbol{\eta}_i = [\eta_{i,0}, \eta_{i,1}, \ldots, \eta_{i,Nr-1}]^T$ denotes the noise vector.

STTD decoding is expressed as

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \end{pmatrix} = \begin{pmatrix} \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_0^* + \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_1 \\ \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_1^* - \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_0 \end{pmatrix} \text{ for } N_t = 2, \tag{A· 3a}$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \begin{pmatrix} \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_0^* + \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_1 + \boldsymbol{r}_2^H \bar{\boldsymbol{h}}_2 \\ \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_1^* - \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_0 + \boldsymbol{r}_3^H \bar{\boldsymbol{h}}_2 \\ \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_2^* - \boldsymbol{r}_2^H \bar{\boldsymbol{h}}_0 - \boldsymbol{r}_3^H \bar{\boldsymbol{h}}_1 \end{pmatrix} \text{ for } N_t = 3, \tag{A· 3b}$$

$$\begin{pmatrix} \hat{d}_0 \\ \hat{d}_1 \\ \hat{d}_2 \end{pmatrix} = \begin{pmatrix} \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_0^* + \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_1 + \boldsymbol{r}_2^H \bar{\boldsymbol{h}}_2 + \boldsymbol{r}_3^T \bar{\boldsymbol{h}}_3^* \\ \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_1^* - \boldsymbol{r}_1^H \bar{\boldsymbol{h}}_0 - \boldsymbol{r}_2^T \bar{\boldsymbol{h}}_3^* + \boldsymbol{r}_3^H \bar{\boldsymbol{h}}_2 \\ \boldsymbol{r}_0^T \bar{\boldsymbol{h}}_2^* + \boldsymbol{r}_1^T \bar{\boldsymbol{h}}_3^* - \boldsymbol{r}_2^H \bar{\boldsymbol{h}}_0 - \boldsymbol{r}_3^H \bar{\boldsymbol{h}}_1 \end{pmatrix} \text{ for } N_t = 4. \tag{A· 3c}$$

Substituting Eqs. (A· 1) and (A· 2) into (A· 3), we obtain

$$\begin{pmatrix}\hat{d}_0\\ \hat{d}_1\end{pmatrix} = \sqrt{S}\sum_{n=0}^{1}\sum_{m=0}^{N_r-1}|h_{m,n}|^2\begin{pmatrix}d_0\\ d_1\end{pmatrix} + \begin{pmatrix}\boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_0^* + \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_1\\ \boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_1^* - \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_0\end{pmatrix} \text{ for } N_t = 2, \quad (\text{A}\cdot 4\text{a})$$

$$\begin{pmatrix}\hat{d}_0\\ \hat{d}_1\\ \hat{d}_2\end{pmatrix} = \sqrt{\frac{2S}{3}}\sum_{n=0}^{2}\sum_{m=0}^{N_r-1}|h_{m,n}|^2\begin{pmatrix}d_0\\ d_1\\ d_2\end{pmatrix} + \begin{pmatrix}\boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_0^* + \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_1 + \boldsymbol{\eta}_2^H\bar{\boldsymbol{h}}_2\\ \boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_1^* - \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_0 + \boldsymbol{\eta}_3^H\bar{\boldsymbol{h}}_2\\ \boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_2^* - \boldsymbol{\eta}_2^H\bar{\boldsymbol{h}}_0 - \boldsymbol{\eta}_3^H\bar{\boldsymbol{h}}_1\end{pmatrix} \text{ for } N_t = 3, \quad (\text{A}\cdot 4\text{b})$$

$$\begin{pmatrix}\hat{d}_0\\ \hat{d}_1\\ \hat{d}_2\end{pmatrix} = \sqrt{\frac{S}{2}}\sum_{n=0}^{3}\sum_{m=0}^{N_r-1}|h_{m,n}|^2\begin{pmatrix}d_0\\ d_1\\ d_2\end{pmatrix} + \begin{pmatrix}\boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_0^* + \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_1 + \boldsymbol{\eta}_2^H\bar{\boldsymbol{h}}_2 + \boldsymbol{\eta}_3^T\bar{\boldsymbol{h}}_3^*\\ \boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_1^* - \boldsymbol{\eta}_1^H\bar{\boldsymbol{h}}_0 - \boldsymbol{\eta}_2^T\bar{\boldsymbol{h}}_3^* + \boldsymbol{\eta}_3^H\bar{\boldsymbol{h}}_2\\ \boldsymbol{\eta}_0^T\bar{\boldsymbol{h}}_2^* + \boldsymbol{\eta}_1^T\bar{\boldsymbol{h}}_3^* - \boldsymbol{\eta}_2^H\bar{\boldsymbol{h}}_0 - \boldsymbol{\eta}_3^H\bar{\boldsymbol{h}}_1\end{pmatrix} \text{ for } N_t = 4. \quad (\text{A}\cdot 4\text{c})$$

From Eq. (A$\cdot$4), the conditional SNR $\gamma(E_s/N_0, \bar{\boldsymbol{H}})$ is given by

$$\gamma\left(\frac{E_s}{N_0}, \bar{\boldsymbol{H}}\right) = 2\left(\frac{1}{N_t}\frac{E_s}{N_0}\right)\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}|h_{m,n}|^2, \quad (\text{A}\cdot 5)$$

which is equal to Eq. (13), but with N_r being replaced by N_t. Thus, the average BER is given by

$$P_b\left(\frac{E_s}{N_0}\right) = \left[\frac{1}{2}\left(1 - \sqrt{\frac{\Gamma/N_t}{2+\Gamma/N_t}}\right)\right]^L \times\sum_{k=0}^{L-1}\begin{pmatrix}L-1+k\\ k\end{pmatrix}\left[\frac{1}{2}\left(1 + \sqrt{\frac{\Gamma/N_t}{2+\Gamma/N_t}}\right)\right]^k, \approx \frac{1}{2^L}\left(\frac{\Gamma}{N_t}\right)^{-L}\begin{pmatrix}2L-1\\ L\end{pmatrix} \text{ for } \Gamma \gg 1, \quad (\text{A}\cdot 6)$$

where $L = N_t \times N_r$.

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