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Rate 1 space-time and space-frequency spreading diversity technique

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Abstract

In this study, a rate 1 transmitter diversity technique is proposed. The proposed technique uses spreading transform and space-time block coding (STBC) together. Space-time, space-frequency, and frequency diversity applications of the proposed technique are shown. The code matrix of the proposed technique can be designed systematically. The proposed technique needs the channel coefficients to stay constant over transmission of 2 rows of the coding matrix regardless of the size of the coding matrix or the diversity order. However, joint detection of the symbols is required for the proposed technique. We used computer simulations to compare our technique with the quasi-orthogonal space-time block coding (QOSTBC), orthogonal space-time block coding (OSTBC), and spreading transform diversity methods. The results showed that the proposed technique provides higher SNR-BER gain than OSTBC and spreading transform diversity, and can provide a higher gain than QOSTBC for time-varying channels. The proposed technique is less sensitive to the time selectivity of the channel than QOSTBC or OSTBC. The detection complexity of the proposed technique is lower than that of spreading diversity.

Key Words: *Diversity, space-time coding, space-frequency coding, carrier spreading, OFDM*

1. Introduction

Multiple-input multiple-output (MIMO) systems are capable of large diversity and capacity gains. The large capacity gain of MIMO systems over single antenna systems was shown in [1, 2]. However, utilizing multiple antennas at the transmitter is more desirable for mobile applications. It is possible to obtain diversity gain for multitransmit antennas by the use of space-time coding [3]. The basic performance criterion for space-time codes was shown in [3, 4]. A simple space-time diversity technique was proposed by Alamouti, which

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obtains diversity gain by using 2 transmit and any number of receive antennas [5]. The Alamouti technique also allows the use of linear maximum likelihood detection. The Alamouti technique was generalized for any number of transmit antennas by using the theory of orthogonal design by Tarokh, Jafarkhani, and Calderbank, referred to as space-time block coding (STBC) [6]. STBC is also referred to as orthogonal space-time block coding (OSTBC) because of its orthogonal code matrices. Unfortunately, if the number of transmit antennas is greater than 2 and complex constellations are used, OSTBC causes transmission rate loss [6]. Therefore, quasi-orthogonal space-time block coding (QOSTBC) is proposed even if low complexity linear decoding cannot be used [7-9]. Both OSTBC and QOSTBC require the channel coefficients to be constant over several symbol periods, and this duration increases by the number of transmit antennas. Assuming that the channel coefficients stay constant over several symbol periods is not suitable for time-selective fading channels [10].

Orthogonal frequency division multiplexing (OFDM) is used in frequency-selective fading channels. OFDM converts frequency-selective channels to many parallel flat fading channels. OFDM transmits N symbols simultaneously by orthogonal subcarriers in NT_s symbol durations, where T_s is the symbol duration. Therefore, the OFDM symbol duration becomes NT_s, and this duration may be very large. The first space-frequency codes proposed for OFDM systems made use of previously existing space-time codes [11]. Lee and Williams proposed methods to use STBC in OFDM systems [12, 13]. In the case of frequency-selective channels, OFDM has also a potential to obtain frequency diversity. Carrier spreading transform diversity (CSTD) is another diversity technique that utilizes frequency diversity for OFDM systems [14, 15].

In this study, the Alamouti code and the spreading transforms were used together to obtain diversity gain. The proposed diversity technique did not cause any loss in transmission rate. Unfortunately, the proposed technique has more detection complexity than OSTBC. However, the detection complexities of the proposed technique and QOSTBC are the same. The new technique can employ any number of transmit antennas as a space-time coding technique without causing any loss in transmission rate. The proposed technique is also suitable to be used in OFDM systems and can obtain a higher order diversity gain with a smaller spreading matrix, a lower decoding complexity, and a higher minimum product distance when compared to carrier spreading diversity techniques [14, 15].

This paper is organized as follows. Section 2 summarizes OSTBC, QOSTBC, OFDM, space-frequency block coding (SFBC), STBC-OFDM, CSTD, and a system model for these systems. Section 3 defines the proposed diversity technique. Section 4 presents the simulation model and analyzes simulation results of the proposed technique, and Section 5 presents some conclusions.

Notation: Underscores and double underscores are used to denote vectors and matrices, respectively; ^{T} and ^{H} represent transpose and complex transpose, respectively; * denotes complex conjugates; $\|.\|_F$ denotes the Frobenius norm; $\underline{\underline{I_N}}$ denotes an $N \times N$ identity matrix; $\underline{S}(r)$ denotes the $(r)^{th}$ entry of $\underline{S}$; and $Diag(\underline{d})$ denotes a diagonal matrix in which the diagonal entries are elements of $\underline{d}$.

2. STBC and QOSTBC

In a STBC system, the channel is assumed to be quasi-static. Therefore, the channel coefficients are assumed to stay constant over several symbol periods. A STBC system can utilize multiple transmit antennas and any number of receive antennas. However, we assumed that the receiver has one receive antenna in this study. The receive vector of such a STBC system is given below.

$$\underline{Y} = \underline{\underline{C}}\underline{H} + \underline{N} \tag{1}$$

We denoted the channel coefficient between the j^{th} transmit antenna and the receive antenna at the t^{th} symbol duration by α_j^t. The assumed noise is additive white Gaussian (AWGN) and has a variance of σ^2 per dimension. $\underline{H}$ and $\underline{N}$ are the *px1* channel and noise vectors, respectively. $\underline{C}$ is the *pxn* STBC code matrix. Each row of $\underline{C}$ is transmitted simultaneously and each column of $\underline{C}$ is transmitted from one of the transmit antennas. The number of transmit antennas is denoted by n and the total duration of transmitting $\underline{C}$ is denoted by p. The channel is a quasi-static Rayleigh fading channel, so the channel coefficients or the elements of the channel vector are assumed to stay constant while transmitting the entire code matrix $\underline{\underline{C}}$. If OSTBC or QOSTBC is employed and the transmit symbols are complex-valued, $\mp s_1, \mp s_2, \ldots, \mp s_k, \mp s_1^*, \mp s_2^*, \ldots, \mp s_k^*$ are the elements of $\underline{\underline{C}}$.

As mentioned previously, OSTBC cannot be applied without reducing the transmission rate for complex constellations with $n > 2$ [6]. The transmission rate is defined as $R = k/p$, where k is the number of transmitted symbols. The maximum transmission rate for $n > 2$ is 3/4 [16]. The transmission rate of 1/2 is always reachable for any number of transmit antennas [6]. The codes for the 3/4 transmission rate for $n = 3$ and $n = 4$ are given in [6, 17, 18], and R =3/5 codes for $n = 5$ and $n = 6$ are shown in [19]. The only rate 1 STBC code for complex constellations is the Alamouti code. The receive vector can be written as follows for the Alamouti code.

$$\underbrace{\begin{bmatrix} y_1 \\ y_2 \end{bmatrix}}_{\underline{Y}} = \begin{bmatrix} s_1 & s_2 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} \alpha_1^t \\ \alpha_2^t \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ -s_2^* & s_1^* \end{bmatrix} \begin{bmatrix} \alpha_1^{t+1} \\ \alpha_2^{t+1} \end{bmatrix} + \underbrace{\begin{bmatrix} n_1 \\ n_2 \end{bmatrix}}_{\underline{N}} \tag{2}$$

Because of the quasi-static assumption of the channel, $\alpha_j^t = \alpha_j^{t+1} = \alpha_j \, ; j = 1, 2$ is satisfied and Eq. (3) can be written as follows.

$$\underbrace{\begin{bmatrix} y_1 \\ y_2 \end{bmatrix}}_{\underline{Y}} = \underbrace{\begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix}}_{\underline{\underline{C}}} \underbrace{\begin{bmatrix} \alpha_1 \\ \alpha_2 \end{bmatrix}}_{\underline{H}} + \underbrace{\begin{bmatrix} n_1 \\ n_2 \end{bmatrix}}_{\underline{N}} \tag{3}$$

Using Eq. (3), we can write the equation below:

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \underbrace{\begin{bmatrix} \alpha_1 & \alpha_2 \\ \alpha_2^* & -\alpha_1^* \end{bmatrix}}_{\underline{\underline{A_1}}} \begin{bmatrix} s_1 \\ s_{2,} \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \end{bmatrix}, \tag{4}$$

where $\underline{\underline{A_1}}$ is the virtual code matrix. Using Eq. (4), we can write the following equation:

$$\underline{\underline{A_1^H}} \begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \begin{bmatrix} |\alpha_1|^2 + |\alpha_2|^2 & 0 \\ 0 & |\alpha_1|^2 + |\alpha_2|^2 \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \underline{\underline{A_1^H}} \begin{bmatrix} n_1 \\ n_2^* \end{bmatrix}. \tag{5}$$

The symbols in Eq. (5) can be detected by linear maximum likelihood detection. The transmitted symbols can be detected separately when OSTBC is used [6]. Unfortunately, OSTBC causes transmission rate loss when the transmit antenna number is greater than 2 and complex constellations are used [6].

The quasi-orthogonal space-time block coding (QOSTBC) does not cause transmission rate loss for $n=4$ [7-9] or $n = 8$ [8-20]. However, linear maximum likelihood detection cannot be used for QOSTBC. The virtual code matrix $\underline{\underline{B}}$ for QOSTBC is given below.

$$\begin{aligned} \underline{\underline{A_v^t}} &= \begin{bmatrix} \alpha_{2v-1}^t & \alpha_{2v}^t \\ (\alpha_{2v}^{t+1})^* & -(\alpha_{2v-1}^{t+1})^* \end{bmatrix} ; v = 1, 2, \ldots \\ \underline{\underline{B}} &= \begin{bmatrix} \underline{\underline{A_1^t}} & \underline{\underline{A_2^t}} \\ \underline{\underline{(A_2^{t+2})^*}} & \underline{\underline{-(A_1^{t+2})^*}} \end{bmatrix} \end{aligned} \tag{6}$$

We can assume that $\underline{\underline{A_v^t}} = \underline{\underline{A_v^{t+2}}} = \underline{\underline{A_v}}$ or $\alpha_j^t = \alpha_j^{t+1} = \alpha_j^{t+2} = \alpha_j^{t+3} = \alpha_j \;; j = 1, 2, 3, 4$ is satisfied for the quasi-static channels. The number of transmit antennas is 4 for the coding matrix $\underline{\underline{B}}$. If we multiply $\underline{\underline{B}}^H$ by the receive vector, we obtain:

$$\underline{\underline{B}}^H \begin{bmatrix} y_1 \\ y_2^* \\ y_3^* \\ y_4 \end{bmatrix} = \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \underline{\underline{B}}^H \underbrace{\begin{bmatrix} n_1 \\ n_2^* \\ n_3^* \\ n_4 \end{bmatrix}}_{\underline{N'}}, \tag{7}$$

where $a = |\alpha_1|^2 + |\alpha_2|^2 + |\alpha_3|^2 + |\alpha_4|^2$ and $b = \alpha_1\alpha_4^* + \alpha_4\alpha_1^* - \alpha_2\alpha_3^* - \alpha_3\alpha_2^*$. Two separate receive vectors can be obtained at the receiver, which is shown below.

$$\begin{aligned} \begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} &= \begin{bmatrix} a & b \\ b & a \end{bmatrix} \begin{bmatrix} s_1 \\ s_4 \end{bmatrix} + \begin{bmatrix} \tilde{n}_1 \\ \tilde{n}_4 \end{bmatrix} \\ \begin{bmatrix} y_3^* \\ y_4 \end{bmatrix} &= \begin{bmatrix} a & -b \\ -b & a \end{bmatrix} \begin{bmatrix} s_2 \\ s_3 \end{bmatrix} + \begin{bmatrix} \tilde{n}_2 \\ \tilde{n}_3 \end{bmatrix} \end{aligned} \tag{8}$$

$$\begin{bmatrix} \tilde{n}_1 \\ \tilde{n}_2 \\ \tilde{n}_3 \\ \tilde{n}_4 \end{bmatrix} = \underline{\underline{B}}^H \underline{N'} \tag{9}$$

Using the receive vectors in Eq. (8), the transmitted symbols can be detected pair by pair.

The pairwise error probability of space-time codes for quasi-static Rayleigh fading was provided in [3, 4] and is given below.

$$P(\underline{\underline{C}} \to \underline{\underline{\hat{C}}}) \le \frac{1}{2} \left(\prod_{j=1}^{r} \lambda_j \right)^{-m} \left(\frac{SNR}{4} \right)^{-rm} \tag{10}$$

Here, $P(\underline{\underline{C}} \to \underline{\underline{\hat{C}}})$ is the probability of detecting $\underline{\underline{\hat{C}}}$ as $\underline{\underline{C}}$; r is the rank of $\left(\underline{\underline{C}} - \underline{\underline{\hat{C}}}\right)$; $\lambda_1, \lambda_2, \ldots, \lambda_r$ are the nonzero eigenvalues of $\left(\underline{\underline{C}} - \underline{\underline{\hat{C}}}\right)$; and m is the number of receive antennas. To reduce pairwise error probability, the rank r and the product of the eigenvalues of $\left(\underline{\underline{C}} - \underline{\underline{\hat{C}}}\right)$ should be as large as possible. Therefore, the minimum rank of the $n \times n$ QOSTBC matrix should be n to obtain full diversity gain [3]. Symbol rotation is recommended to guarantee full diversity gain under QOSTBC [21-24]. Eq. (8) can be arranged for rotated symbols, as below.

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \begin{bmatrix} a & b \\ b & a \end{bmatrix} \begin{bmatrix} s_1 \\ s_4 e^{j\phi} \end{bmatrix} + \begin{bmatrix} \tilde{n}_1 \\ \tilde{n}_2 \end{bmatrix}$$
$$\begin{bmatrix} y_3^* \\ y_4 \end{bmatrix} = \begin{bmatrix} a & -b \\ -b & a \end{bmatrix} \begin{bmatrix} s_2 \\ s_3 e^{j\phi} \end{bmatrix} + \begin{bmatrix} \tilde{n}_3 \\ \tilde{n}_4 \end{bmatrix} \tag{11}$$

Determination of an optimum rotation angle (ϕ) for different constellations was given in [21]. If the number of transmit antennas is 8 and QOSTBC is used, the virtual code matrix $\underline{\underline{B}}$ may contain rate 3/4 STBC virtual code matrices; therefore, we may obtain $R=3/4$ for 8 transmit antennas [7]. Rate 1 QOSTBC for 8 transmit antennas was explained in [8-20]. The rate 1 virtual code matrix for 8 transmit antennas was given in [8] and is presented below.

$$\underline{\underline{B8a}} = \begin{bmatrix} \underline{\underline{A_1}} & \underline{\underline{A_2}} & \underline{\underline{A_3}} & \underline{\underline{A_4}} \\ \underline{\underline{A_2}} & \underline{\underline{A_1}} & \underline{\underline{A_4}} & \underline{\underline{A_3}} \\ \underline{\underline{A_3}} & \underline{\underline{A_4}} & \underline{\underline{A_1}} & \underline{\underline{A_2}} \\ \underline{\underline{A_4}} & \underline{\underline{A_3}} & \underline{\underline{A_2}} & \underline{\underline{A_1}} \end{bmatrix} \tag{12}$$

Another rate 1 virtual code matrix for 8 transmit antennas was given in [10] and is presented below.

$$\underline{\underline{B8b}} = \begin{bmatrix} \underline{\underline{A_1}} & \underline{\underline{A_2}} & \underline{\underline{A_3}} & \underline{\underline{A_4}} \\ -\underline{\underline{A_2^*}} & \underline{\underline{A_1^*}} & -\underline{\underline{A_4^*}} & \underline{\underline{A_3^*}} \\ -\underline{\underline{A_3^*}} & -\underline{\underline{A_4^*}} & \underline{\underline{A_1^*}} & \underline{\underline{A_2^*}} \\ \underline{\underline{A_4}} & -\underline{\underline{A_3}} & -\underline{\underline{A_2}} & \underline{\underline{A_1}} \end{bmatrix} \tag{13}$$

If code matrices $\underline{\underline{B8a}}$ and $\underline{\underline{B8b}}$ are used, the detection should be made jointly for 4 symbols. Full order diversity gain can be obtained for $\underline{\underline{B8a}}$ or $\underline{\underline{B8b}}$ by the rotation of the symbols.

3. OFDM, space-frequency, and carrier spreading transform diversity

Orthogonal frequency division multiplexing (OFDM) is used for high data rate communications. OFDM divides a frequency-selective fading channel into many parallel flat fading channels. An OFDM receiver requires only a fast Fourier transform processor followed by a single-tap equalizer. Single carrier systems employ multitap equalizers while OFDM systems employ single-tap equalizers. Thus, the complexity of the receiver is reduced by the use of OFDM. An OFDM block is formed by multiplying an $N \times N$ inverse discrete Fourier transform matrix ($\underline{\underline{F^H}}$) and a transmit symbol vector ($\underline{S}$):

$$\underline{Z_k} = \underline{\underline{F^H}} \underbrace{\begin{bmatrix} s_0 \\ \vdots \\ s_{N-1} \end{bmatrix}}_{\underline{S}} = \begin{bmatrix} z_0 \\ \vdots \\ z_{N-1} \end{bmatrix}, \tag{14}$$

where $\underline{Z_k}$ contains OFDM symbols that will be transmitted simultaneously. N symbols are transmitted simultaneously in an NT_s duration, where T_s is the symbol duration. A cyclic prefix should be added to an OFDM transmit vector to convert the linear convolution of the channel into a circular convolution. A cyclic prefix added to an OFDM transmit vector is shown below.

$$\underline{Z_k^g} = \begin{bmatrix} z_{N-1-G+1} \\ \vdots \\ z_{N-1} \\ z_0 \\ \vdots \\ z_{N-1} \end{bmatrix} \tag{15}$$

By adding the last G entries of vector $\underline{Z_k}$ to its own head, the vector $\underline{Z_k^g}$, with an added cyclic prefix, can be obtained, as seen in Eq. (15). Each of the channels between the transmit and the receive antennas are frequency-selective fading channels and have L taps with the same power delay profiles. $G \geq L$ should be satisfied to avoid interblock interference. After adding a cyclic prefix, the OFDM block is transmitted through a frequency-selective channel. After match filtering, the receiver removes the cyclic prefix from the received signal and multiplies it by the discrete Fourier transform matrix $\underline{\underline{F}}$ to obtain the received symbol vector $\underline{X}$.

$$\underline{X} = \begin{bmatrix} \alpha^t(0) & 0 & \dots & 0 \\ 0 & \alpha^t(1) & \dots & 0 \\ 0 & 0 & \ddots & 0 \\ 0 & \dots & \dots & \alpha^t(N-1) \end{bmatrix} \begin{bmatrix} s_0 \\ s_1 \\ \vdots \\ s_{N-1} \end{bmatrix} + \underline{\underline{F}} \begin{bmatrix} n^t(0) \\ n^t(1) \\ \vdots \\ n^t(N-1) \end{bmatrix} \tag{16}$$

$\alpha^t(k)$is the channel coefficient for the k^{th} subcarrier at the t^{th} symbol duration and is given by:

$$\alpha^t(k) = \sum_{l=0}^{L-1} h^t(l) e^{-j2\pi k \Delta f \tau_l} k = 0, 1, \dots N-1, \tag{17}$$

where $h^t(l)$ is the zero-mean complex Gaussian random variable with variances $E\left[\left|h^t(l)\right|^2\right] = \delta_l^2$, τ_l is the delay of the l^{th} path, and Δf is the subcarrier frequency separation. If multiple transmit antennas and one receive antenna were employed, the receive signal for subcarrier $k$at the t^{th} symbol duration would be as follows:

$$x^t(k) = \sum_{j=1}^{n} \alpha_j^t(k) s_j^t(k) + n^t(k), \tag{18}$$

where $\alpha_j^t(k)$ is the channel coefficient between the j^{th} transmit antenna and the receive antenna at the t^{th} symbol duration, $s_j^t(k)$ is the data transmitted by transmit antenna j, and $n^t(k)$ is the zero-mean complex Gaussian noise term corresponding to subcarrier k. The channel coefficients between transmit antenna j and the receive antenna can be denoted as follows.

$$\underline{\alpha_j^t} = \begin{bmatrix} \alpha_j^t(0) & \alpha_j^t(1) & \dots & \alpha_j^t(N-1) \end{bmatrix}^T = \underline{\underline{W}}\underline{A_j} \tag{19}$$

Here,

$$\underline{\underline{W}} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ e^{-j2\pi\Delta f\tau_0} & e^{-j2\pi\Delta f\tau_1} & \cdots & e^{-j2\pi\Delta f\tau_{l-1}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-j(N-1)2\pi\Delta f\tau_0} & e^{-j(N-1)2\pi\Delta f\tau_1} & \cdots & e^{-j(N-1)2\pi\Delta f\tau_{l-1}} \end{bmatrix} \tag{20}$$

is related to the delay distribution and

$$\underline{A_j} = \begin{bmatrix} h_j^t(0) & h_j^t(1) & \cdots & h_j^t(L-1) \end{bmatrix}^T \tag{21}$$

is related to the power distribution of the channel. The correlation matrix of the channel frequency response can be written as follows.

$$\underline{\underline{R_j}} = \underline{\underline{R}} = E\{\underline{\alpha_j^t}(\underline{\alpha_j^t})^H\} = \underline{\underline{W}}Diag([\ \delta_0^2 \quad \delta_1^2 \quad \cdots \quad \delta_{l-1}^2]^T)\ \underline{\underline{W^H}} \tag{22}$$

Lee and Williams proposed techniques to use STBC in OFDM [12, 13]. Assuming that the nearest subcarriers come through the same channel conditions, it is possible to transmit each row of the Alamouti matrix by the nearest subcarriers [12]. However, this method employs only spatial diversity, but OFDM also has a potential to obtain frequency diversity gain.

Carrier spreading techniques can be used to obtain frequency diversity in OFDM [14, 15]. The spreading transform is realized by multiplying the symbol vector $\underline{S}$ by a spreading matrix:

$$\underline{S_{Sp}} = \underline{\underline{G}} \underbrace{\begin{bmatrix} s_0 \\ s_1 \\ \vdots \\ s_{p-1} \end{bmatrix}}_{\underline{S}}. \tag{23}$$

The elements of $\underline{S_{Sp}}$ are transmitted by different subcarriers to obtain frequency diversity. The spreading matrix $\underline{\underline{G}}$ should be square and orthogonal. The absolute values of the elements of $\underline{\underline{G}}$ should also be equal [14]. If a Hadamard matrix is used as a spreading matrix, it shows asymptotically bad error performance. The rotation of the constellation was recommended in [14] to overcome this problem. Therefore, we rotated the Hadamard spreading matrix in this study. The rotation of the Hadamard spreading transform is shown below, where $\underline{\underline{G}}$ is the Hadamard matrix.

$$\underline{\underline{G_{rot}}} = \underline{\underline{G}}Diag(\underline{r}) \tag{24}$$

Meanwhile, $\underline{r}$ is the rotation vector and is given below.

$$\underline{r} = [\ r_0 \quad r_1 \quad \ldots \quad r_{p-1}\ ]^T;\ r_i = \exp\left(j\frac{2\pi}{\mu}\frac{i}{p}\right) \tag{25}$$

The value of μ is dependent on the constellation. If MPSK modulation is employed, we may have $\mu = M$ [14].

The Vandermonde matrix was utilized as the spreading matrix in [25-27]. If the Vandermonde matrix is used as the spreading matrix, there is no need for constellation rotation. Larger minimum product distances provide smaller pairwise error probabilities, as can be seen from Eq. (10). The Vandermonde matrix provides a larger minimum product distance [25].

Unfortunately, linear maximum likelihood detection cannot be employed for the spreading transform diversity. Symbols should be detected jointly, and detection complexity increases exponentially with the size of the spreading matrix. Suboptimum detection techniques may be used to reduce detection complexity, but this causes some performance degradation [14].

4. Space-time and space-frequency coded spreading diversity

We used Alamouti coding and the spreading transforms together to exceed the second order of transmit diversity. The transmission rate of the proposed technique is 1. The symbol vectors were multiplied by the spreading matrix $\underline{\underline{G}}$ and the spread symbols were coded by Alamouti coding. We used rotated Hadamard and Vandermonde matrices to spread the symbols. The Vandermonde matrix is given below.

$$\underline{\underline{V_{p/2}}} = \begin{bmatrix} 1 & \theta_1 & \cdots & (\theta_1)^{(\frac{p}{2}-1)} \\ 1 & \theta_2 & \cdots & (\theta_2)^{(\frac{p}{2}-1)} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & \theta_{(p/2)} & \cdots & \left(\theta_{(p/2)}\right)^{(\frac{p}{2}-1)} \end{bmatrix} \tag{26}$$

Based on the constellation and the size of $\underline{\underline{V_{p/2}}}$, the values of $\theta_1, \theta_2, \ldots, \theta_{(p/2)}$ were calculated in [25-27]. The values of $\theta_v,\ v = 1, 2, \ldots, p/2$ for the Vandermonde matrix are calculated as follows, as described in [26].

$$\theta_v = \exp\left(j\frac{4v-3}{p}\pi\right) \tag{27}$$

We assumed that $\underline{S}$ and $\underline{K}$ contain the transmitted symbols.

$$\underline{S} = \begin{bmatrix} s_0 \\ s_1 \\ \vdots \\ s_{\frac{p}{2}-1} \end{bmatrix} \underline{K} = \begin{bmatrix} k_0 \\ k_1 \\ \vdots \\ k_{\frac{p}{2}-1} \end{bmatrix} \tag{28}$$

First, we spread the transmit vectors with a spreading matrix $\underline{\underline{G}}$, like below.

$$\begin{gathered} \underline{S_{sp}} = \underline{\underline{G}}\underline{S} \\ \underline{K_{sp}} = \underline{\underline{G}}\underline{K} \end{gathered} \tag{29}$$

After the spreading of the symbols, we transmitted the spread symbols by Alamouti coding. The energy of the symbols was normalized; thus, $\left\|\underline{\underline{G}}\underline{S}\right\|_F^2 + \left\|\underline{\underline{G}}\underline{K}\right\|_F^2 = p/2$ was satisfied. The 2 $\times$ 1 receive vectors can be written as follows:

$$\underline{R_v} = \underline{\underline{A_v}} \begin{bmatrix} \underline{S_{sp}(v)} \\ \underline{K_{sp}(v)} \end{bmatrix} + \underline{N}\ ; v = 1, 2, \ldots p/2. \tag{30}$$

The proposed method can be used as a space-time, a space-frequency, and a frequency diversity method. Therefore, we divided Section 3 into 2 parts. The application of the proposed technique as a space-time diversity method is shown in Section 3.A. The application of the proposed technique as a space-frequency and frequency diversity method is shown in Section 3.B.

4.1. Space-time coded spreading diversity

Transmit antennas are grouped in pairs. The receive vectors can be written as follows:

$$\underline{R_{ST}} = \underbrace{\begin{bmatrix} \underline{\underline{A_1}} & \underline{\underline{0}} & \cdots & \underline{\underline{0}} \\ \underline{\underline{0}} & \underline{\underline{A_2}} & \ddots & \vdots \\ \vdots & & \ddots & \\ \underline{\underline{0}} & \cdots & & \underline{\underline{A_{p/2}}} \end{bmatrix}}_{\underline{\underline{H_{ST}}}} \begin{bmatrix} S_{sp}(1) \\ K_{sp}(1) \\ S_{sp}(2) \\ K_{sp}(2) \\ \vdots \\ K_{sp}(p/2) \end{bmatrix} + \underline{N_{ST}}, \tag{31}$$

where $\underline{N_{ST}}$ denotes the noise vector and $\underline{\underline{0}}$ denotes a 2×2 matrix such that all of the elements are 0. The channel coefficient between transmit antenna j and the receive antenna at the t^{th} symbol period can be denoted by $\alpha_j^t(j = 1,2,\ldots,p)$. We may assume that the channels stay constant over 2 symbol periods, such that $\alpha_j^t = \alpha_j^{t+1}$ is satisfied, if the coherence time of the channel is larger than the OFDM symbol period. The transmitted symbols should be detected jointly, as can be seen from Eq. (32).

$$\min_{\underline{\tilde{S}}} \left\| \underline{R_{ST}} - \underline{\underline{H_{ST}}} \begin{bmatrix} \underline{\tilde{S}_{sp}}(1) \\ 0 \\ \underline{\tilde{S}_{sp}}(2) \\ 0 \\ \vdots \\ 0 \end{bmatrix} \right\|_F^2, \quad \min_{\underline{\tilde{K}}} \left\| \underline{R_{ST}} - \underline{\underline{H_{ST}}} \begin{bmatrix} 0 \\ \underline{\tilde{K}_{sp}}(1) \\ 0 \\ \underline{\tilde{K}_{sp}}(2) \\ \vdots \\ \underline{\tilde{K}_{sp}}(p/2) \end{bmatrix} \right\|_F^2 \tag{32}$$

The $p/2$ of p of the transmitted symbols is detected jointly. However, if we were to use the spreading transform, we would detect p of p of the transmitted symbols jointly to reach the same order of transmit diversity. On the other hand, if we were to use QOSTBC, the channel coefficients would be accepted as staying constant over p symbol periods.

Because the OFDM symbols may have large durations, the channel coefficients may not be accepted as staying constant while transmitting the entire code matrix for the OFDM systems. The correlation between $\alpha_j^t(k)$ and $\alpha_j^{t+1}(k)$ is defined as [31]:

$$R(T_s) = J_0(2\pi f_m T_s), \tag{33}$$

where f_m is the maximum Doppler frequency and J_0 is the first-kind Bessel function of the zeroth-order.

4.2. Space-frequency coded spreading diversity

We can use the frequency domain to obtain different channel conditions. If the symbol duration is larger than the coherence time, it is possible to use the frequency domain to obtain the Alamouti code [12]. By changing the time domain with the frequency domain, the virtual channel matrix of Alamouti coding for subcarriers k and $k+1$ can be written as:

$$\underline{\underline{A_v(k)}} = \begin{bmatrix} \alpha_{2v-1}(k) & \alpha_{2v}(k) \\ \alpha^*_{2v}(k+1) & -\alpha^*_{2v-1}(k+1) \end{bmatrix}, \tag{34}$$

where $\alpha_j(k)$ denotes the channel coefficient between transmit antenna j and the receive antenna for subcarrier k. The nearest subcarriers may be accepted as coming through the same channels. Therefore, we may assume that $\alpha_j(k) = \alpha_j(k+1);\ j = 1,2$ is satisfied. The correlation matrix of the channels related to p consecutive subcarriers can be written as:

$$\underline{\underline{R_{near}}} = \underbrace{\begin{bmatrix} 1 & 1 & \dots & 1 \\ e^{-j2\pi\Delta f\tau_0} & e^{-j2\pi\Delta f\tau_1} & \dots & e^{-j2\pi\Delta f\tau_{l-1}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-j(p-1)2\pi\Delta f\tau_0} & e^{-j(p-1)2\pi\Delta f\tau_1} & \dots & e^{-j(p-1)2\pi\Delta f\tau_{l-1}} \end{bmatrix}}_{\underline{\underline{W_{near}}}} \begin{bmatrix} \delta_0^2 & 0 & \dots & 0 \\ 0 & \delta_1^2 & 0\dots & 0 \\ 0 & 0 & \ddots & \vdots \\ 0 & 0 & \dots & \delta_{l-1}^2 \end{bmatrix} \underline{\underline{W_{near}^H}}. \tag{35}$$

The far subcarriers may have independent channels. The correlation matrix of the channels related to p far subcarriers can be written as:

$$\underline{\underline{R_{far}}} = \underbrace{\begin{bmatrix} 1 & 1 & \dots & 1 \\ e^{-jq2\pi\Delta f\tau_0} & e^{-jq2\pi\Delta f\tau_1} & \dots & e^{-jq2\pi\Delta f\tau_{l-1}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-jq(p-1)2\pi\Delta f\tau_0} & e^{-jq(p-1)2\pi\Delta f\tau_1} & \dots & e^{-jq(p-1)2\pi\Delta f\tau_{l-1}} \end{bmatrix}}_{\underline{\underline{W_{far}}}} \begin{bmatrix} \delta_0^2 & 0 & \dots & 0 \\ 0 & \delta_1^2 & 0\dots & 0 \\ 0 & 0 & \ddots & \vdots \\ 0 & 0 & \dots & \delta_{l-1}^2 \end{bmatrix} \underline{\underline{W_{far}^H}}. \tag{36}$$

The separation factor q guarantees obtaining p channels that have minimum correlation with each other. The finding of the optimum separation factor was shown in [32]. We may not need to use multiple transmit antennas to obtain different channel conditions if we use optimally separated subcarriers. The virtual channel matrix of Alamouti coding for subcarriers k, $k+1$, $k+q$, and $k+q+1$ can be written as below.

$$\underline{\underline{A_(k,q)}} = \begin{bmatrix} \alpha(k) & \alpha(k+q) \\ \alpha^*(k+q+1) & -\alpha^*(k+1) \end{bmatrix} \tag{37}$$

After the matched filtering, the removing of the cyclic prefix, discrete Fourier transform, and the rearranging of the receive vector for frequency and space-frequency application of the proposed technique is as follows.

$$\underline{R_F} = \underbrace{\begin{bmatrix} \underline{\underline{A_(k,q)}} & \underline{\underline{0}} & \cdots & \underline{\underline{0}} \\ \underline{\underline{0}} & \underline{\underline{A_(k+2q,q_)}} & \cdots & \vdots \\ \vdots & & \ddots & \\ \underline{\underline{0}} & \cdots & \cdots & \underline{\underline{A_(k+(p-2)q,q_)}} \end{bmatrix}}_{\underline{\underline{H_F}}} \begin{bmatrix} \underline{S_{sp}}(1) \\ \underline{K_{sp}}(1) \\ \underline{S_{sp}}(2) \\ \underline{K_{sp}}(2) \\ \vdots \\ \underline{K_{sp}}(p/2) \end{bmatrix} + \underline{N_{SF}} \tag{38}$$

If 2 transmit antennas were used, we would need to use $p/2$ far subcarriers to obtain the p^{th} order of transmit diversity. The receive vector of this application is given below.

$$\underline{R_{SF}} = \underbrace{\begin{bmatrix} \underline{\underline{A_1(k)}} & \underline{\underline{0}} & \cdots & \underline{\underline{0}} \\ \underline{\underline{0}} & \underline{\underline{A_1(k+q)}} & \cdots & \vdots \\ \vdots & & \ddots & \\ \underline{\underline{0}} & \cdots & \cdots & \underline{\underline{A_1(k+(\frac{p}{2}-1)q_)}} \end{bmatrix}}_{\underline{\underline{H_{SF}}}} \begin{bmatrix} \underline{S_{sp}}(1) \\ \underline{K_{sp}}(1) \\ \underline{S_{sp}}(2) \\ \underline{K_{sp}}(2) \\ \vdots \\ \underline{K_{sp}}(p/2) \end{bmatrix} + \underline{N_{SF}} \tag{39}$$

All of the elements of the receive vectors are received simultaneously. The decoding of the symbols can be done as in Eq. (32) by substituting $\underline{\underline{H_{ST}}}$ with $\underline{\underline{H_F}}$ or $\underline{\underline{H_{SF}}}$ and substituting $\underline{R_{ST}}$ with $\underline{R_F}$ or $\underline{R_{SF}}$.

5. Simulation model and results

The Jakes' channel simulation model is widely used to simulate Rayleigh fading channels [28]. However, the Jakes' model has difficulty in creating multiple uncorrelated channel coefficients. Therefore, we used a modified Jakes' channel simulation model that creates multiple uncorrelated Rayleigh channels, as shown below [29]:

$$X(t) = X_C(t) + jX_S(t), \tag{40}$$

$$X_C(t) = \sqrt{\frac{2}{M}} \sum_{k=1}^{M} \cos(\psi_k) \cos(2\pi f_m t \cos\alpha_k + \phi), \tag{41}$$

$$X_S(t) = \sqrt{\frac{2}{M}} \sum_{k=1}^{M} \sin(\psi_k) \cos(2\pi f_m t \cos\alpha_k + \phi), \tag{42}$$

$$\alpha_k = \frac{2\pi k - \pi + \theta}{4M}, k = 1, 2, \ldots, M. \tag{43}$$

where $X(t)$ is the complex channel gain, f_m is the maximum Doppler frequency, and M is the number of sinusoids. The θ, ϕ, and ψ_k are statistically independent and uniformly distributed over [-π, π]. We created multiple uncorrelated Rayleigh channel coefficients and inserted them into a vector denoted by $\underline{Q}$.

$$\underline{Q} = \begin{bmatrix} q_0 \\ q_1 \\ \vdots \\ q_{p-1} \end{bmatrix} \tag{44}$$

We can produce channel coefficients that have the desired correlation matrix, as shown below [30].

$$\underline{H} = \frac{\underline{\underline{L}}\underline{Q}}{\sigma_q} \tag{45}$$

Here, the variance of uncorrelated channel coefficients is σ_q. $\underline{H}$ contains the desired channel coefficients. $\underline{\underline{L}}$ is defined as:

$$\underline{\underline{L}}\underline{\underline{L}}^H = \underline{\underline{R}}, \tag{46}$$

where $\underline{\underline{R}}$ is the desired correlation matrix. We can write the following equation by using eigendecomposition:

$$\underline{\underline{R}} = \underline{\underline{V}}\underline{\underline{\Lambda}}\underline{\underline{V}}^H. \tag{47}$$

We can write $\underline{\underline{L}}$ as follows:

$$\underline{\underline{L}} = \underline{\underline{V}}\sqrt{\underline{\underline{\Lambda}}}. \tag{48}$$

$\underline{H}$ has the desired correlation matrix, which can be seen from the following equation.

$$\begin{aligned} &E(\underline{H}\underline{H}^H) = E\left(\frac{\underline{\underline{L}}\underline{Q}\underline{Q}^H\underline{\underline{L}}^H}{\sigma_q^2}\right) = E(\underline{\underline{L}}\underline{\underline{L}}^H) = \underline{\underline{R}} \\ &E(\underline{Q}\underline{Q}^H) = \sigma_q^2 \underline{\underline{I_N}} \end{aligned} \tag{49}$$

Correlation matrices of OFDM channel frequency response for consecutive and far subcarriers are given in Eqs. (35) and (36), respectively. We can create time-varying channels by using the same method if we have the correlation matrix of time-varying channel coefficients. We used Eq. (33) to create the correlation matrix of time-varying channels.

We used computer simulations to compare the SNR-BER performance of the proposed technique with the performances of QOSTBC, OSTBC, and spreading transform diversity. The simulation parameters used were as follows. The channel coefficients were modeled as independent complex Gaussian random variables and had a variance of 0.5 per dimension, and the envelopes of the channels were Rayleigh-distributed. Noise was modeled as zero-mean AWGN. The 4-QAM constellation was used for rate 1 transmit diversity schemes and the 16-QAM constellation was used for rate 1/2 OSTBC; therefore, the spectral efficiency was 2 b/s/Hz. We assumed that the receiver had one receive antenna and perfect knowledge of the channel state information. The number of symbols, which were detected jointly, was the same for all simulations. A rotation angle of $\pi/4$ was used to obtain fourth-order transmit diversity, and the rotation angles $\pi/8$, $\pi/4$, and $3\pi/8$ were used for the eighth-order diversity in the simulations.

The proposed technique is compared with the other techniques for quasi-static channel assumption in Figures 1-4. We assumed the ratio of Doppler frequency to symbol frequency to be 0.02.

Performance results of the proposed technique for time-varying flat channels are shown in Figures 5-8. The maximum Doppler frequency was 30 Hz and the symbol durations were the minimum (280 μs) and the maximum (1120 μs) symbol durations of the DVB-H system for these simulations.

Figures 9 and 10 compare the performances of the space-frequency and frequency diversity applications of the proposed technique with those of the other techniques. The parameters used were as follows. We used the COST 207 6-ray power delay profile for a typical urban channel for these simulations [33]. The bandwidth of the channel was 7 MHz. The number of subcarriers was 8192. The maximum Doppler frequency was 30 Hz, the symbol duration was 1120 μs, and the separation factor (q) to obtain frequency diversity was 2048. We used 2 transmit antennas and 2 far subcarriers for the scenario shown in Figure 9. We used 4 far subcarriers for that of Figure 10. Therefore, the diversity order is 4 in Figures 9 and 10. We transmitted the rows of the coding matrices by consecutive subcarriers for Figures 9 and 10.

The SNR-BER performances of the proposed technique, rotated QOSTBC, and rate 1/2 OSTBC are compared for fourth- and eighth-order diversity in Figures 1 and 2, respectively. The proposed technique and spreading transform diversity are compared in Figures 3 and 4. No correlation between the different channel coefficients was assumed. The performance difference of the proposed method increased for time-varying channels. However, the channel is quasi-static in Figures 1 and 2, and there were no clear performance differences, as expected. The proposed technique and rotated QOSTBC had similar SNR-BER performances, as seen in Figures 1 and 2, but the proposed technique had a slightly higher performance than rotated QOSTBC for eighth-order diversity. Both rotated QOSTBC and the proposed technique could obtain 3 dB more gain than the rate 1/2 OSTBC, as seen from Figures 1 and 2.

The proposed technique can provide almost 1.5 dB more gain than rotated Hadamard spreading diversity for fourth-order diversity, as can be seen from Figure 3. If we take care of the number of the symbols that should be detected jointly, the proposed technique can provide 5 dB more gain than the spreading transform, as seen from Figure 4.

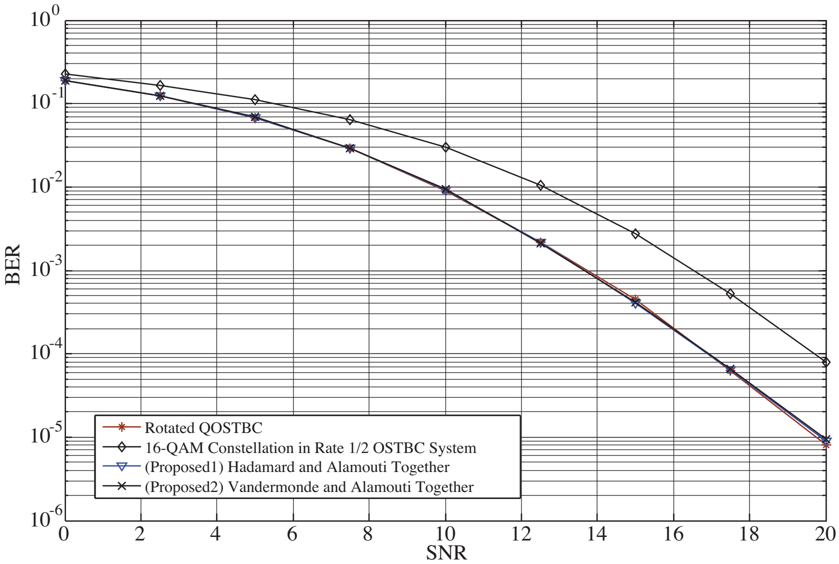

Figure 1. The performances of QOSTBC, OSTBC, and the proposed technique for fourth-order diversity.

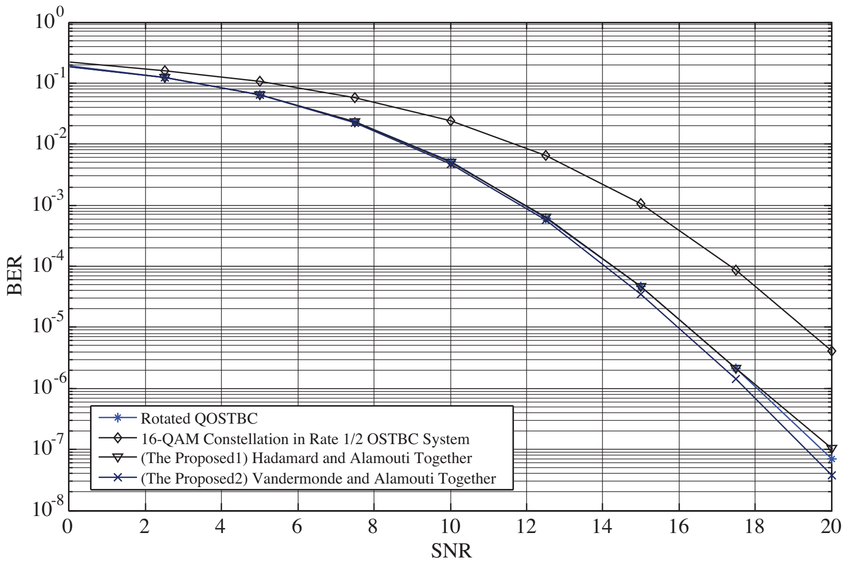

Figure 2. The performances of QOSTBC, OSTBC, and the proposed technique for eighth-order diversity.

Figures 5 and 6 show the performances of the proposed technique and the rotated QOSTBC for symbol durations of 280 μs and 1120 μs, respectively. The diversity order was 4. The proposed technique could provide 2.5 dB more gain than rotated QOSTBC, as can be seen from Figure 5. It could provide 7 dB more gain than rotated QOSTBC when the symbol duration was 1120 μs, as seen from Figure 6.

Figures 7 and 8 show the performances of the proposed technique and the rotated QOSTBC for symbol durations of 280 μs and 1120 μs, respectively. The diversity order was 8. The proposed technique could provide 5 dB more gain than rotated QOSTBC when the symbol duration was 280 μs, as seen from Figure 7. It could provide 11 dB more gain from rotated QOSTBC when the symbol duration was 1120 μs, as seen from Figure 8.

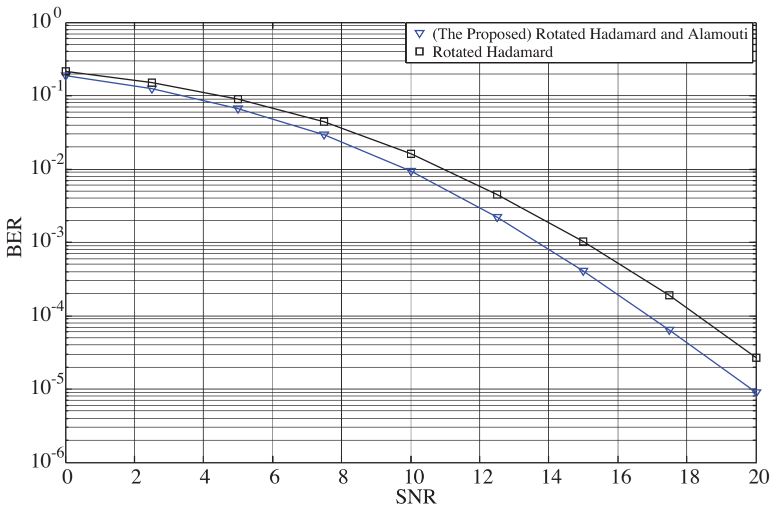

Figure 3. The performances of the proposed and spreading diversity for fourth-order diversity.

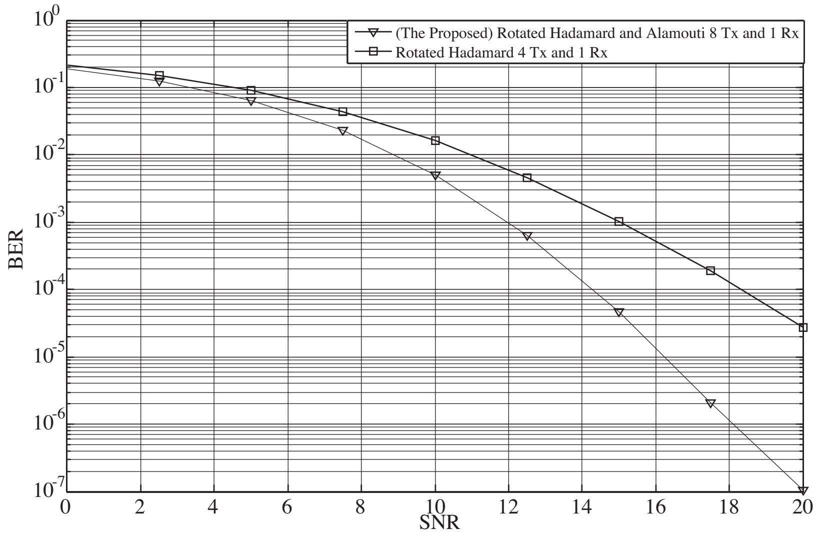

Figure 4. The performances of the proposed and spreading diversity at the same order of detection complexity.

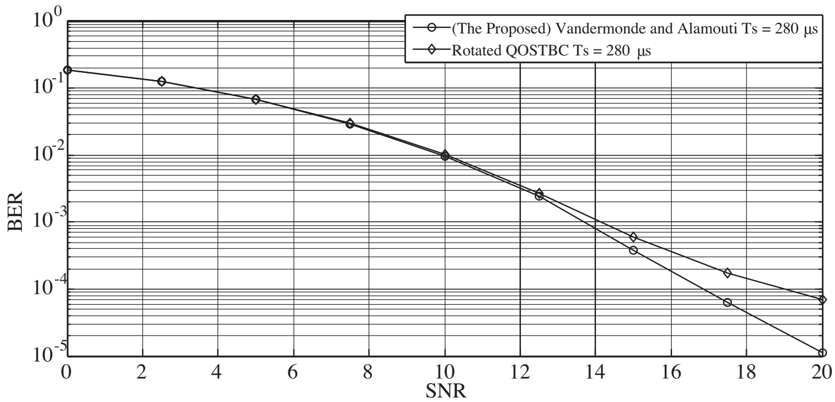

Figure 5. The performances of the proposed and the rotated QOSTBC for fourth-order diversity and 280 μs of symbol duration.

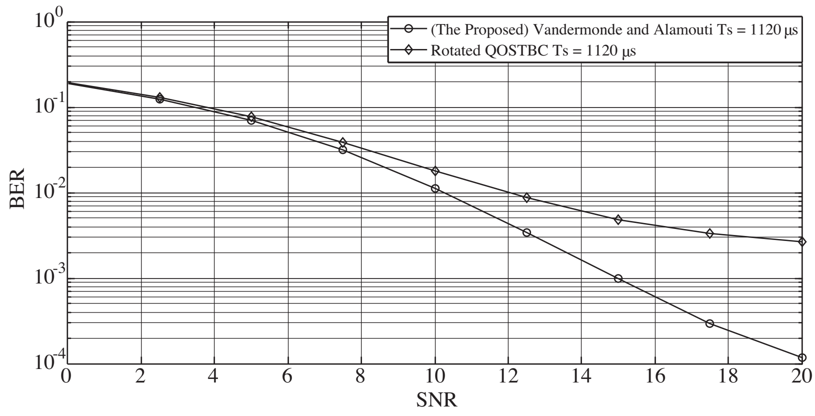

Figure 6. The performances of the proposed and the rotated QOSTBC for fourth-order diversity and 1120 μs of symbol duration.

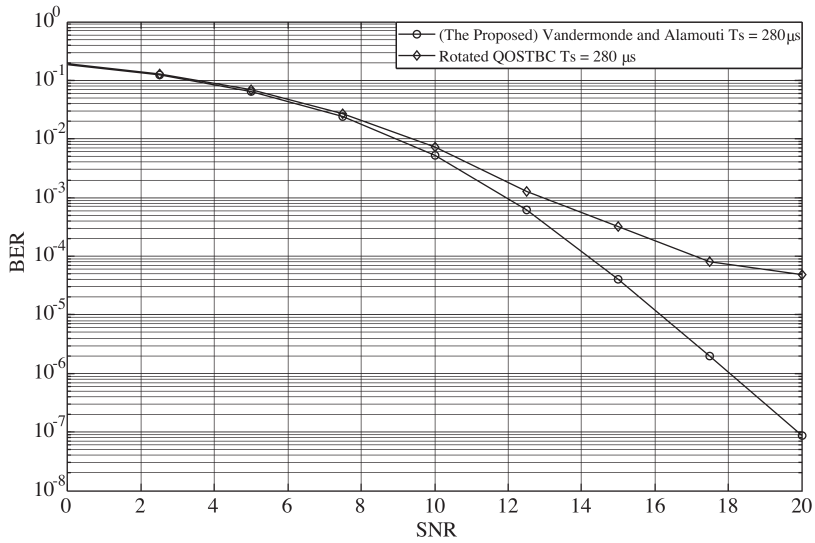

Figure 7. The performances of the proposed and the rotated QOSTBC for eighth-order diversity and 280 μs of symbol duration.

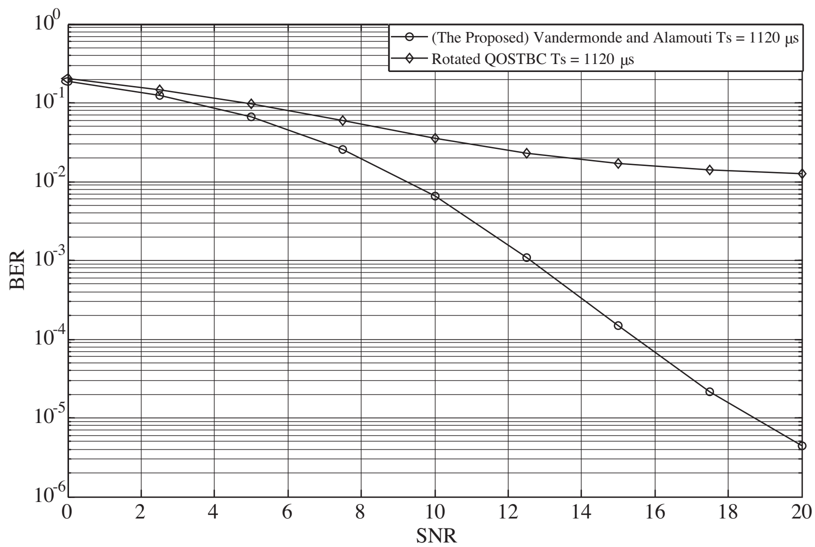

Figure 8. The performances of the proposed and the rotated QOSTBC for eighth-order diversity and 1120 μs of symbol duration.

The proposed technique and rotated QOSTBC had the same performance for space-frequency diversity for the given conditions, as seen from Figure 9. It provided approximately 1.5 dB more gain than the CSTD, as seen from Figure 10. It also performed a little higher than rotated QOSTBC for the given conditions, as also seen in Figure 10.

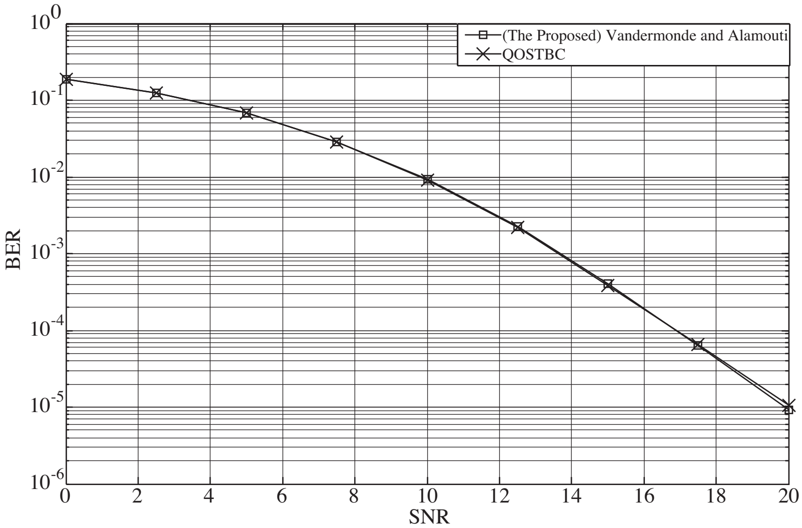

Figure 9. Comparison of the proposed technique and rotated QOSTBC as space-frequency coding for fourth-order diversity.

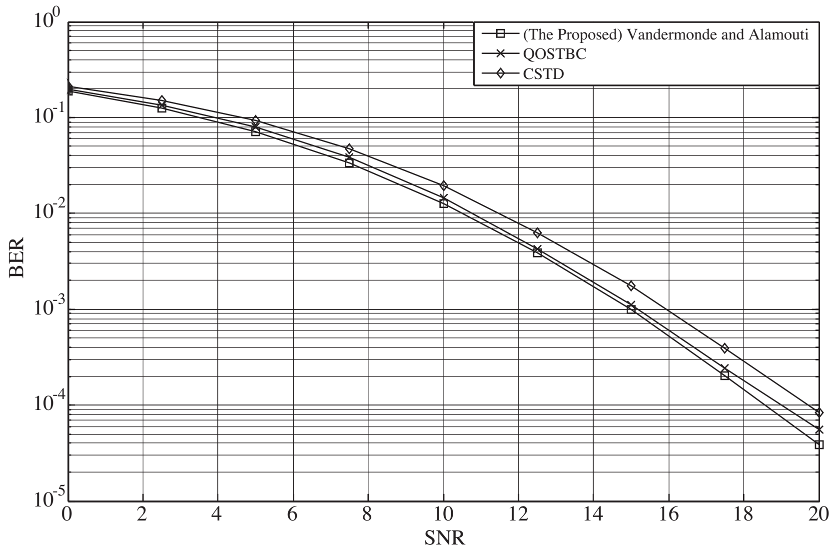

Figure 10. Comparison of the proposed technique, rotated QOSTBC, and CSTD as frequency diversity for fourth-order diversity.

6. Conclusions

In this paper, a rate 1 transmitter diversity technique was proposed. The proposed technique uses spreading transform and space-time block coding (STBC) together and can employ any order of diversity gain without reducing the transmission rate. A space-time, space-frequency, and frequency diversity application of the proposed technique were shown for a single carrier system and orthogonal frequency division multiplexing (OFDM).

The proposed technique causes an increase of the detection complexity because joint detection of the symbols is required. The number of the symbols that are required to be detected jointly is equal to half of the transmit diversity order. The same number of symbols is required to be detected jointly for QOSTBC at the same order of diversity. However, the proposed diversity technique has a systematically designable code matrix structure and requires the channel coefficients to stay constant over transmission of 2 rows of the code matrix regardless of size of the code matrix. While QOSTBC coding matrices cannot be designed systematically for any number of transmit antennas, QOSTBC requires the channel coefficients to stay constant over the transmitting of the entire code matrix. OFDM symbol durations may be very large and the channels may change significantly in a few OFDM symbol durations. According to the simulation results, the proposed technique can provide significantly more gain than QOSTBC if the symbol duration is large.

When the spreading transform is used to obtain frequency diversity gain in OFDM, the detection complexity increases exponentially with the size of the spreading matrix and a larger spreading matrix is required to obtain a higher order diversity gain. Because both Alamouti coding and spreading transform are utilized by the proposed technique, we can reach the same order of transmit diversity with half the size of the spreading matrix, making the decoding less complex and the minimum product distance larger when compared to CSTD. Therefore, the proposed technique can reach a higher SNR-BER gain than the carrier spreading transform. According to simulation results, the proposed technique can provide more gain than CSTD as a frequency diversity technique.

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