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Research Article

Canceling Interferences for High Data Rate Time Reversal MIMO UWB System: A Precoding Approach

Taotao Wang and Tiejun Lv

School of Information and Communication Engineering, Beijing University of Posts and Telecommunications (BUPT), Beijing 100876, China

Correspondence should be addressed to Tiejun Lv, lvtiejun@bupt.edu.cn

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An ultra-high data rate time reversal (TR) multiple-input multiple-output (MIMO) ultra-wideband (UWB) communication system with space-time precoding is proposed. When the symbol duration is set to approach the duration of UWB monocycles, the data rate is close to the limit, resulting in the severe intersymbol interference (ISI). The zero-forcing (ZF) criterion-based space-time precoding presented in this paper eliminates both ISI and multistream interference (MSI) caused by spatial multiplexing at the sampling time. With less demand for the degree of freedom (the number of antennas) than other existing schemes, the proposed scheme enables the data rate to reach the order of Gbps without losing bit error rate (BER) performance. Since TR signal preprocessing and the proposed precoding both require the channel state information (CSI), a simple but effective channel estimation algorithm is also proposed to evaluate the impact of channel estimation on the proposed scheme.

1. Introduction

Ultra-wideband (UWB) impulse radio communications, as a promising candidate for location-aware indoor communications, wireless sensor networks (WSN) and wireless personal area network (WPAN), has received significant attention in both academia and industry in recent years [1, 2]. The most attractive feature of UWB is its potential to offer great capacity in theory as compared with the narrowband systems. However, the conventional UWB system shows much lower data rate than expectation. This is because capturing the energy of dense multipath channel [3] and combating severe intersymbol interference (ISI) caused by large maximum excess delay of the channel [4] will increase receiver complexity which limits both detection performance and data rate under the condition that receivers with high complexity are not preferred in UWB short-range applications. To reduce receiver complexity, noncoherent scheme is developed to bypass the complicated treatments on UWB channel, whereas the deterioration of detection performance and the reduction of data rate are inevitable [5]. On the other hand, the system complexity can be shifted from the receiver to the transmitter, where the power and computation resources are generally enough to implement signal processing. Since preprocessing the signal before transmission may cope with the deteriorating effects of the channel, the receiver can keep a simple structure without losing detection performance and data rate. In particular, signal preprocessing scheme is desirable in the networks where a central node with sufficient power and computation resources serves many distributed nodes with extremely stringent limits on complexity and power consumption [6].

A time reversal (TR) (TR is also referred to as pre-Rake diversity combining [7]) preprocessing-based system with minimum mean-squared error (MMSE) equalizer is firstly applied to combat ISI in UWB communications [8]. TR preprocessing is that the transmitter takes the time reversed channel impulse response (CIR) as a filter to prefilter the original signal before transmission. If the prefiltered signal is radiated into the channel, it convolves with the CIR and leads to a strong peak at the output of the channel at one particular instant. As a result, the receiver can be simplified significantly and meanwhile makes full use of the energy from all paths of the channel. Recently, the TR-based UWB system and its variations have been investigated in [9–14].

Multiple-input multiple-output (MIMO) technique, employing multiple antennas at the transmitter and receiver, is capable of increasing data transmission rate by spatial multiplexing without expanding the bandwidth. In order to transmit parallel data streams simultaneously (spatial multiplexing), the multistream interference (MSI) of MIMO channel must be mitigated. The potential of TR-based UWB system with multiple antennas to increase data rate is studied in [9]. In [10], a TR-based scheme for MSI suppression is proposed for MIMO-UWB system without considering ISI. (To be exact, the schemes in [10] are proposed for multiuser UWB system, which consists of an access point with multiple transmit antennas and several single-antenna radio terminals. Obviously, it is equivalent to a MIMO-UWB system without cooperation among receive antennas.) Further, TR is proposed to cope with both MSI and ISI in MIMO-UWB system in [11]. It is worthwhile to note that the interferences are not absolutely eliminated by TR in [10, 11] though they are mitigated to a certain extent, which becomes the principal factor to cause error for the large signal-to-noise ratio (SNR) and results in the deterioration of bit error rate (BER) performance ultimately.

In this paper, we propose an ultra-high data rate TR-MIMO-UWB system with space-time precoding. Multiple antennas can increase data rate, whereas the occurrence of MSI degrades the system performance. The ultra-high data rate UWB transmission usually requires extremely short symbol, thus ISI is very strong. In order to implement a TR-based UWB system, the channel state information (CSI) must have been available at transmitter. Therefore, the interferences (MSI and ISI) of TR-MIMO-UWB system should be canceled by using CSI at transmitter rather than at receiver. In [15, 16], the precoding scheme, which is extensively applied in narrowband system, has been employed to eliminate the MSI of TR-MIMO-UWB system. Since the effect of ISI is not taken into account, their system performances are degenerating rapidly as symbol duration is shorted. In this work, the space-time precoding matrix based on zero-forcing (ZF) criterion is originally derived, which is independent on the degree of freedom (the number of antenna). The proposed space-time precoding can effectively eliminate both ISI and MSI, and it is beneficial to achieve the high data rate of TR-MIMO-UWB system up to the order of Gbps without losing BER performance.

Since TR signal preprocessing requires the CSI, a simple but effective channel estimation algorithm for TR-MIMO-UWB system is also presented in this work. In [13, 14], the authors utilize a feedback channel to send the estimated channel information from the receiver to the transmitter. Because the UWB channel is characterized by dense multi-path, that is, the number of multipath is large, the required bandwidth for the feedback channel is huge. Hence, the implementation of feedback channel is unfeasible. The proposed channel estimation exploits the reciprocity of the UWB channel which has experimentally been demonstrated in [12]. That is to say that the receiver sends training symbols, and the channel estimation algorithm is performed at transmitter. As the channel estimation algorithm is introduced to acquire CSI, the imperfection of CSI inevitably presents at this more practical UWB system. Since the proposed scheme can more effectively use the CSI to cancel the interferences, it shows more robustness to the error of channel estimation.

The rest of this paper is organized as follows. The system model of ultra-high data rate TR single-input single-output (SISO) UWB is described in Section 2, and the TR-MIMO-UWB system with space-time precoding is proposed in Section 3. In Section 4, we address the channel estimation problem for TR-MIMO-UWB system. Section 5 presents the simulation results. Finally, conclusions are drawn in Section 6.

Notation. The boldface letters denote vector or matrix. $\mathbf{0}_{m\times n}$ is a matrix of size $m \times n$ with all entries being zeros. $\otimes$ represents convolution operation. $\lfloor \cdot \rfloor$ stands for integer floor operation. vec($\mathbf{A}$) returns $\mathbf{A}$ transformed into a column vector with one column stacked onto the next. $\|\mathbf{A}\|$, $\mathbf{A}^T$ and $\mathbf{A}^{-1}$ stand for the Euclidian norm, the transpose and the inverse of matrix $\mathbf{A}$, respectively.

2. System Model of TR-SISO-UWB

In this section, a peer-to-peer TR-SISO-UWB system is described. The UWB impulse radio signal with binary pulse amplitude modulation (BPAM) is

$$z(t) = \sqrt{E_b} \sum_{j=-\infty}^{+\infty} b_j \omega(t - jT_s), \tag{1}$$

where $\omega(t)$ is the monocycle pulse waveform with very short duration T_ω and normalized energy, $T_s = N_s T_\omega$ is the symbol duration which is assumed to be an integer multiple of the pulse waveform duration, $b_j \in \{\pm 1\}$ is the jth binary symbol and E_b denotes the bit energy. $z(t)$ is prefiltered by the time reversed CIR before transmission, then the transmitted signal is

$$x(t) = z(t) \otimes \hat{g}(-t) = \sqrt{E_b} \sum_{j=-\infty}^{+\infty} b_j s(t - jT_s), \tag{2}$$

where $\hat{g}(t)$ is the estimate of the UWB channel impulse response $g(t)$ and

$$s(t) = \omega(t) \otimes \hat{g}(-t) \tag{3}$$

is the transmitted waveform for one binary symbol.

The dense multipath environment, such as the industrial and indoor office [17], is considered in this paper, and the CIR $g(t)$ is modeled as

$$g(t) = \sum_{l=0}^{L_g-1} \alpha_l \delta(t - l\Delta), \tag{4}$$

where δ is the Dirac delta function, L_g is the number of resolvable multipath components (MPCs), α_l is the fading coefficient of the lth MPC, and Δ is the minimum multipath resolution, which is equal to the duration of $\omega(t)$ ($\Delta = T_\omega$), as any two paths whose relative delay is less than T_ω are not resolvable. The maximum excess delay of the channel is denoted by $T_g = (L_g - 1)\Delta$. In conventional UWB systems, the symbol duration T_s is usually set large enough ($T_s > T_g + T_\omega$) to avoid or alleviate ISI. However, in this paper, T_s is set much smaller than T_g ($T_s \ll T_g$) to achieve an ultra-high data rate. It can be found that the waveform duration of $s(t)$ is $T_g + T_\omega$, and thus the transmitted waveform for one symbol overlaps that of other symbols.

The transmitted signal is radiated into the channel, and it convolves with the CIR. The received signal is

$$r(t) = x(t) \otimes g(t) + n(t) = \sqrt{E_b} \sum_{j=-\infty}^{+\infty} b_j \omega(t - jT_s) \otimes R(t) + n(t), \tag{5}$$

where

$$R(t) = \hat{g}(-t) \otimes g(t) \tag{6}$$

is the correlation function between $\hat{g}(t)$ and $g(t)$, $n(t)$ is the zero-mean additive white Gaussian noise (AWGN) with two sided power spectral density (PSD) $N_0/2$. Substituting $\hat{g}(t) = \sum_{l=0}^{L_g-1} \hat{\alpha}_l \delta(t - l\Delta)$ and (4) into (6), we have

$$R(t) = \begin{cases} \delta(t - l\Delta) \displaystyle\sum_{i=0}^{l+L_g-1} \hat{\alpha}_{i-l}\alpha_i, & -L_g + 1 \le l \le 0, \\ \delta(t - l\Delta) \displaystyle\sum_{i=0}^{L_g-1-l} \hat{\alpha}_i \alpha_{i+l}, & 1 \le l \le L_g - 1. \end{cases} \tag{7}$$

Since the dense multipath channel is regarded as an equally spaced model, (7) is actually a sequence of delta functions with regular spacings. This channel model is employed for the only purpose of facilitating the analysis for ISI. And if more general channel model is involved, the validity of our proposal is still supported. If $\hat{g}(t)$ is the perfect estimation of $g(t)$, the peak of $R(t)$ is $R(0)$, and $R(0) = \sum_{i=0}^{L_g-1} \alpha_i^2 = 1$ due to channel energy normalization. A simple filter is designed to capture the desired energy at the positions of peak as follows:

$$y_j = \int_0^{T_\omega} r(t + jT_s)\omega(t)dt = b_j\sqrt{E_b}R(0) + I_j + n_j, \tag{8}$$

where y_j is the decision statistic for b_j, $n_j = \int_0^{T_\omega} n(t + jT_s)\omega(t)dt$ is the noisy component, and I_j is the ISI component for b_j. For the purpose of analyzing the interference pattern at receiver, we define the received waveform $c(t)$ for one symbol as

$$c(t) = s(t) \otimes g(t) = \omega(t) \otimes R(t). \tag{9}$$

The process that signal transmits from transmitter to receiver in the absence of noise is illustrated in Figure 1. Since the duration of $c(t)$ ranges from $-T_g$ to $T_g + T_\omega$, any symbol is interfered by its following M_1 symbols and preceding M_2 symbols at receiver, where $M_1 = \lfloor T_g/T_s \rfloor = \lfloor (L_g - 1)/N_s \rfloor$ and $M_2 = \lfloor (T_g + T_\omega)/T_s \rfloor = \lfloor L_g/N_s \rfloor$. From (5), $R(t)$ can be considered as an equivalent channel impulse response (ECIR), and we define discrete form of the equivalent channel as a $(M_1 + M_2 + 1) \times 1$ vector $\mathbf{f} = [f_{-M_1}, \ldots, f_{-1}, f_0, f_1, \ldots, f_{M_2}]^T$, where

$$f_i = \int_0^{T_\omega} c(t - iT_s)\omega(t)dt = R(iT_s), \tag{10}$$

$i = -M_1, \ldots, -1, 0, 1, \ldots, M_2$, is the ith sampling value of ECIR. Using the discrete form of $R(t)$, the ISI component in (8) can be expressed as

$$I_j = \sum_{\substack{i=j-M_2 \\ i \ne j}}^{j+M_1} f_{-i} b_i. \tag{11}$$

It can be observed in (11) that the concerned interferences are only dependent on the sampling value of the ECIR, yet they do not relate to the value of ECIR at any other time.

3. TR-MIMO-UWB with Space-Time Precoding

3.1. System Description. A TR-MIMO-UWB communication system includes a transmitter equipped with N_t antennas and a receiver equipped with N_r antennas. N_r parallel data streams are transmitted simultaneously. In typical indoor environments, the UWB channel is quasistatic [17, 18]. That means UWB channels remain invariant over a block of symbols duration, but they are allowed to change from block to block. Therefore, block transmission is adopted in the proposed scheme. We consider a block of $N_r \times L$ bit binary symbols, which is represented by L column vectors $\mathbf{b}_i = [b_{i,1}, b_{i,2}, \ldots b_{i,N_r}]^T$ for $i = 1, 2, \ldots, L$. The L column vectors are stacked in one $N_rL \times 1$ column vector which is $\mathcal{B} = \mathrm{vec}([\mathbf{b}_1, \mathbf{b}_2, \ldots, \mathbf{b}_L])$.

The $N_rL \times N_rL$ space-time precoding matrix is denoted by $\mathcal{P}$. After using $\mathcal{P}$ to prefilter $\mathcal{B}$, we get an $N_rL \times 1$ column vector

$$\mathcal{X} = \alpha\mathcal{P}\mathcal{B}, \tag{12}$$

where $\alpha = N_rL/\|\mathcal{P}\mathcal{B}\|$ is the energy normalization factor which guarantees the average transmitted energy to be $\sqrt{E_b}$ for one binary symbol.

$\mathcal{X}$ is fed into a parallel-to-serial converter to get L column vectors $\mathbf{x}_i$ of size $N_r \times 1$ for $i = 1, 2, \ldots, L$ ($\mathcal{X} = \mathrm{vec}([\mathbf{x}_1, \mathbf{x}_2, \ldots, \mathbf{x}_L])$). The $N_r \times (M_1 + L + M_2)$ transmit symbol matrix $\mathcal{D}$ is constructed by padding M_1 zero guard vectors at the front of $\mathbf{x}_1$ and M_2 zero guard vectors at the end of $\mathbf{x}_L$ (the size of all zero guard vectors is $N_r \times 1$); that is, $\mathcal{D} = [\mathbf{0}_{N_r \times M_1}, \mathbf{x}_1, \mathbf{x}_2, \ldots, \mathbf{x}_L, \mathbf{0}_{N_r \times M_2}]$. The (k, j)th entry of $\mathcal{D}$ is denoted by $d_{k,j}$, which is the jth transmit symbol of kth data stream. The TR signal radiated by the pth antenna at transmitter in a block duration $(M_1 + L + M_2)T_s$ for $p = 1, 2, \ldots, N_t$ is given by

$$x_p(t) = \sqrt{E_b} \sum_{j=1}^{M_1+L+M_2} \frac{1}{\sqrt{N_t}} \sum_{k=1}^{N_r} d_{k,j}\omega(t - (j-1)T_s) \otimes \hat{g}_{k,p}(-t), \tag{13}$$

where $\hat{g}_{k,p}(t)$ is the estimation of $g_{k,p}(t)$ which stands for the impulse response of the multipath channel between the

pth antenna at transmitter and the kth antenna at receiver. In Section 4, a channel estimation algorithm is proposed to obtain $\hat{g}_{k,p}(t)$. It is worthwhile to note that all N_r parallel data streams are simultaneously transmitted from the pth antenna at transmitter.

The signal received by the qth antenna at receiver for $q = 1, 2, \ldots, N_r$ is expressed as

$$\begin{aligned} r_q(t) &= \sum_{p=1}^{N_t} x_p(t) \otimes g_{q,p}(t) + n_q(t) \\ &= \sqrt{E_b} \sum_{j=1}^{M_1+L+M_2} \sum_{k=1}^{N_r} d_{k,j}\,\omega(t-(j-1)T_s) \otimes R_{q,k}(t) \\ &\quad + n_q(t), \end{aligned} \tag{14}$$

where $n_q(t)$ is the AWGN at the qth receive antenna and $R_{q,k}(t)$ is the sum of N_t correlation functions which is defined as

$$R_{q,k}(t) = \frac{1}{\sqrt{N_t}} \sum_{p=1}^{N_t} \hat{g}_{k,p}(-t) \otimes g_{q,p}(t), \tag{15}$$

for $q, k = 1, 2, \ldots, N_r$. It can be noticed in (14) that the qth receive antenna receives all N_r parallel data streams simultaneously from N_r equivalent channels which is represented by its impulse response $R_{q,k}(t)$. At the back-end of the qth receive antenna, a simple filter which matches to $\omega(t)$ captures the energy as follows:

$$y_{j,q} = \int_0^{T_\omega} r_q(t+(j-1)T_s)\,\omega(t)\,dt, \tag{16}$$

where $y_{j,q}$ is the jth decision statistic at the qth receive antenna, $j = 1, 2, \ldots, M_1 + L + M_2$, and $q = 1, 2, \ldots, N_r$. Substituting (14) into (16), we have

$$y_{j,q} = z_{j,q} + n_{j,q}, \tag{17}$$

where

$$\begin{aligned} z_{j,q} &= \int_0^{T_\omega} \left\{ \sqrt{E_b} \sum_{i=1}^{M_1+L+M_2} \sum_{k=1}^{N_r} d_{k,i}\,\omega(t) \otimes R_{q,k}(t) \right\} \omega(t)\,dt \\ &= \sqrt{E_b} \sum_{k=1}^{N_r} \sum_{i=-M_1}^{M_2} d_{k,j+i} \int_0^{T_\omega} \left[\omega(t) \otimes R_{q,k}(t+iT_s) \right] \omega(t)\,dt \end{aligned} \tag{18}$$

is the sampling value of signal and

$$n_{j,q} = \int_0^{T_\omega} n_q(t+(j-1)T_s)\,\omega(t)\,dt \tag{19}$$

is the discrete noise component.

3.2. Space-Time Precoding Matrix Design. In order to transmit N_r parallel data streams simultaneously at a very high data rate without losing performance, the MSI and ISI must be eliminated. As the CSI is already available for the implementation of TR signal preprocessing, we can use the CSI to calculate the precoding matrix $\mathcal{P}$. In this paper, we seek the solution based on ZF criterion.

The jth decision statistic vector of size $N_r \times 1$ is given by $\mathbf{y}_j = [y_{j,1}, y_{j,2}, \ldots, y_{j,N_r}]^T$, and the corresponding signal vector and noise vector are $\mathbf{z}_j = [z_{j,1}, z_{j,2}, \ldots, z_{j,N_r}]^T$, $\mathbf{n}_j = [n_{j,1}, n_{j,2}, \ldots, n_{j,N_r}]^T$, respectively. Moreover, the $M_1 + L + M_2$ column vectors $\{\mathbf{y}_j\}_{j=1}^{M_1+L+M_2}$ are stacked in one $N_r(M_1 + L + M_2) \times 1$ column vector $\mathcal{Y}$; that is, $\mathcal{Y} = \operatorname{vec}([\mathbf{y}_1, \mathbf{y}_2, \ldots, \mathbf{y}_{M_1+L+M_2}])$. Similarly, we get $\mathcal{Z} = \operatorname{vec}([\mathbf{z}_1, \mathbf{z}_2, \ldots, \mathbf{z}_{M_1+L+M_2}])$ and $\mathcal{N} = \operatorname{vec}([\mathbf{n}_1, \mathbf{n}_2, \ldots, \mathbf{n}_{M_1+L+M_2}])$. Notably, the desired decision statistic vector for bit vector $\mathbf{b}_j$ is $\mathbf{y}_{M_1+j}$, $j = 1, 2, \ldots, L$. We stack the desired decision statistic vectors in an $N_r L \times 1$ column vector $\widetilde{\mathcal{Y}} = \operatorname{vec}([\mathbf{y}_{M_1+1}, \mathbf{y}_{M_1+2}, \ldots, \mathbf{y}_{M_1+L}])$, which is the decision statistic for $\mathcal{B}$. Now, we can establish the discrete input-output relationship between $\mathcal{B}$ and $\mathcal{Y}$ as

$$\mathcal{Y} = \mathcal{Z} + \mathcal{N} = \sqrt{E_b}\mathcal{H}\mathcal{X} + \mathcal{N} = \alpha\sqrt{E_b}\mathcal{H}\mathcal{P}\mathcal{B} + \mathcal{N}, \tag{20}$$

where $\mathcal{H}$ is the space-time channel matrix (STCM). Via extending ECIR to MIMO channel, the ECIR matrix of size $N_r \times N_r$ is given by

$$\mathbf{H}(t) = \begin{bmatrix} R_{1,1}(t) & \cdots & R_{1,N_r}(t) \\ \vdots & \ddots & \vdots \\ R_{N_r,1}(t) & \cdots & R_{N_r,N_r}(t) \end{bmatrix}. \tag{21}$$

Then, the STCM $\mathcal{H}$ in (20) can be represented as an $N_r(M_1 + L + M_2) \times N_r L$ block Toeplitz matrix

$$\mathcal{H} = \begin{bmatrix} \mathbf{H}_{-M_1} & 0 & \cdots & 0 \\ \mathbf{H}_{-M_1+1} & \mathbf{H}_{-M_1} & \ddots & \vdots \\ \vdots & \mathbf{H}_{-M_1+1} & \ddots & 0 \\ \mathbf{H}_{-1} & \vdots & \ddots & \mathbf{H}_{-M_1} \\ \mathbf{H}_0 & \mathbf{H}_{-1} & \ddots & \mathbf{H}_{-M_1+1} \\ \mathbf{H}_1 & \mathbf{H}_0 & \ddots & \vdots \\ \vdots & \mathbf{H}_1 & \ddots & \mathbf{H}_{-1} \\ \mathbf{H}_{M_2} & \vdots & \ddots & \mathbf{H}_0 \\ 0 & \mathbf{H}_{M_2} & \ddots & \mathbf{H}_1 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \mathbf{H}_{M_2} \end{bmatrix}, \tag{22}$$

where $\mathbf{H}_i = \mathbf{H}(iT_s)$, $i = -M_1, \ldots, -1, 0, 1, \ldots, M_2$ is the ith sampling value of ECIR matrix. From (20), it can be found that the space-time MIMO relationship between the transmitted information bits and the sampling values of

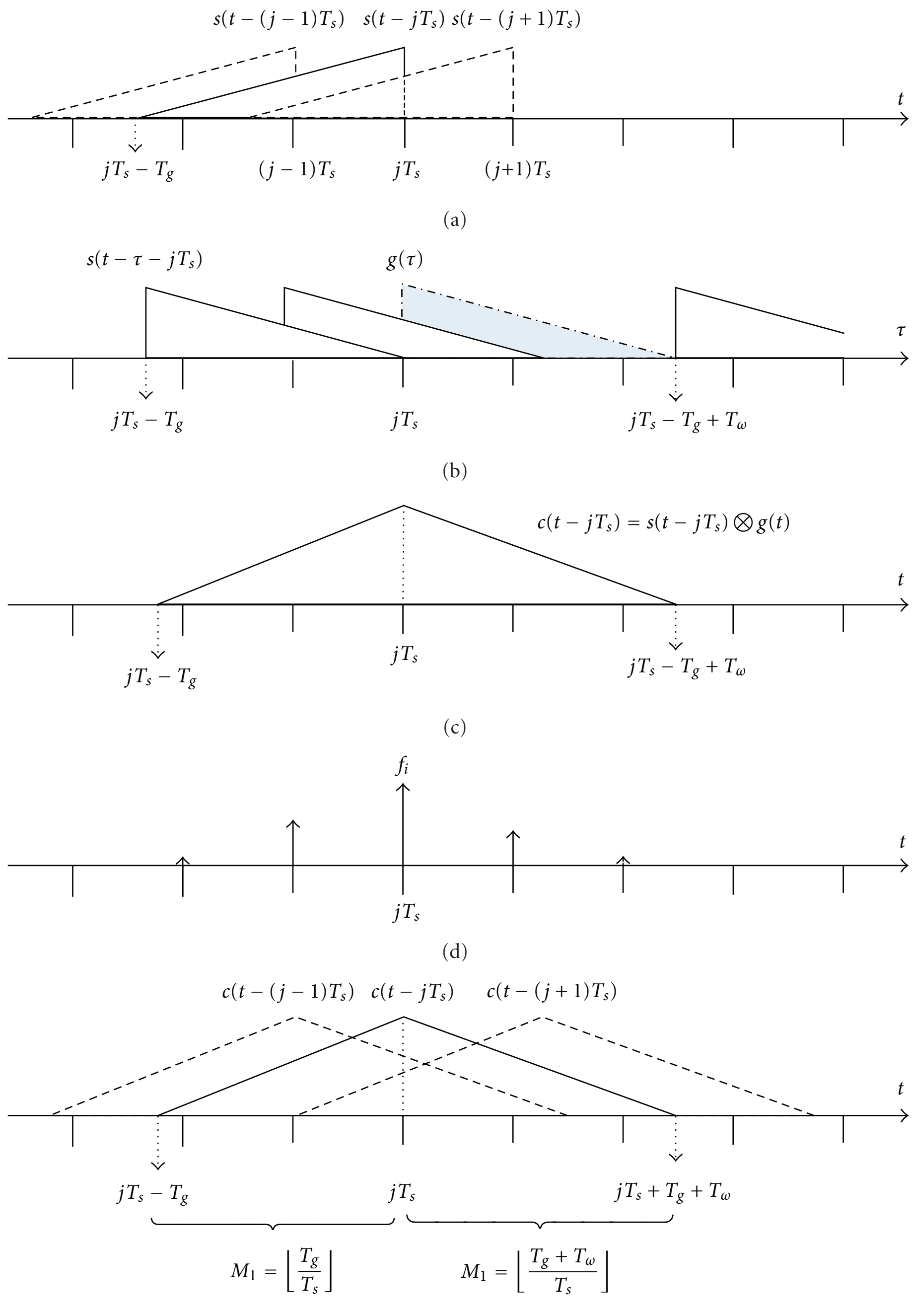

Figure 1: The process that signal transmits from the transmitter to receiver in the absence of noise. (a) Signal at the output of transmitter. (b) The process of transmitted signal convolving with CIR. (c) The received waveform for one symbol. (d) The discrete form of ECIR. (e) Received waveforms are interfered by each other.

received signal is constructed. Therefore, the interference in time domain (ISI) and the interference in spatial domain (MSI) can be eliminated at the same time by employing ZF precoding matrix $\mathcal{P}$ to diagonalize $\mathcal{H}$.

Since the right pseudoinverse of $\mathcal{H}$ is inexistent ($N_r(M_1 + L + M_2) > N_r L$), ZF-based precoding matrix $\mathcal{P}$ cannot be directly solved from (20). On the other hand, $\mathcal{H}$ can be rewritten as $\mathcal{H} = [\mathbf{h}_1^T, \mathbf{h}_2^T, \ldots, \mathbf{h}_{N_r(M_1+L+M_2)}^T]^T$, where $\mathbf{h}_j$ is the jth row of $\mathcal{H}$. The desired statistic $\widetilde{\mathcal{Y}}$ is a part of $\mathcal{Y}$, which consists of the elements ranging from the $(N_r M_1 + 1)$th to the $(N_r M_1 + N_r L)$th within the vector $\mathcal{Y}$. In (20), $\widetilde{\mathcal{Y}}$ is related to the rows ranging from the $(N_r M_1 + 1)$th to the $(N_r M_1 + N_r L)$th within the matrix $\mathcal{H}$. Therefore, the input-output relationship between $\mathcal{B}$ and $\widetilde{\mathcal{Y}}$ is given as

$$\widetilde{\mathcal{Y}} = \alpha\sqrt{E_b}\widetilde{\mathcal{H}}\mathcal{P}\mathcal{B} + \widetilde{\mathcal{N}}, \tag{23}$$

where $\widetilde{\mathcal{H}} = [\mathbf{h}^T_{M_1N_r+1}, \mathbf{h}^T_{M_1N_r+2}, \ldots, \mathbf{h}^T_{M_1N_r+LN_r}]^T$ is an $N_rL \times N_rL$ matrix which consists of the (N_rM_1+1)th to the $(N_rM_1+N_rL)$th row in the matrix $\mathcal{H}$ and $\widetilde{\mathcal{N}} = \mathrm{vec}([\mathbf{n}_{M_1+1}, \mathbf{n}_{M_1+2}, \ldots, \mathbf{n}_{M_1+L}])$. According to (23), the ZF-based precoding matrix which diagonalizes $\widetilde{\mathcal{H}}$ is given as

$$\mathcal{P} = \widetilde{\mathcal{H}}^T \left(\widetilde{\mathcal{H}}\widetilde{\mathcal{H}}^T\right)^{-1}. \tag{24}$$

Since $\widetilde{\mathcal{H}}$ is a square matrix, its right pseudoinverse exists, and (24) can be calculated.

However, the actual ECIR matrix $\mathbf{H}(t)$ and corresponding STCM $\mathcal{H}$ cannot be obtained at transmitter, we can only achieve the estimations of them. The estimated ECIR matrix at transmitter is calculated as

$$\widehat{\mathbf{H}}(t) = \begin{bmatrix} \widehat{R}_{1,1}(t) & \cdots & \widehat{R}_{1,N_r}(t) \\ \vdots & \ddots & \vdots \\ \widehat{R}_{N_r,1}(t) & \cdots & \widehat{R}_{N_r,N_r}(t) \end{bmatrix}, \tag{25}$$

where

$$\widehat{R}_{q,k}(t) = \frac{1}{\sqrt{N_t}} \sum_{p=1}^{N_t} \widehat{g}_{k,p}(-t) \otimes \widehat{g}_{q,p}(t) \tag{26}$$

is the estimated ECIR between the kth equivalent transmit antenna and pth receive antenna. Replacing $\mathbf{H}(t)$ with $\widehat{\mathbf{H}}(t)$ in (22), the estimated STCM $\widehat{\mathcal{H}}$ is immediately obtained and used to calculate the precoding matrix $\mathcal{P}$. Obviously, if the estimations are perfect, the interference can be effectively eliminated; otherwise, the imperfect estimations may result in the residual interferences.

Some remarks about the TR-MIMO-UWB system with ZF space-time precoding are essential.

(i) From (13), all N_r parallel data streams are simultaneously transmitted from one antenna. That means the number of transmitted parallel data streams is independent on N_t, but only lies on N_r. This is not in common with ordinary MIMO systems in which $\min\{N_t, N_r\}$ parallel data streams can be normally transmitted. This is because the TR MIMO system in this paper is a wideband system, where the TR processing filters or the CIR act as orthogonal codes to spread information bits (they are actually quasiorthogonal and after TR preprocessing the interferences are mitigated to a certain extent). Therefore, the data stream number is not constrained by $\min\{N_t, N_r\}$.

(ii) There is no cooperation among receive antennas in the proposed scheme so that the scheme can be naturally extended to multiuser UWB system.

(iii) The motivation to insert the zero guard vectors is to prevent the interference between blocks. Admittedly, this operation will result in some data rate reduction. In fact, the data rate of the proposed system is $R_b = N_rL/(M_1+L+M_2)T_s$ bits per second (bps). Owing to that, the coherence time of the typical indoor UWB channel is rather larger than the maximum excess delay of the channel, N_rL is of the same order as $M_1 + L + M_2$. Therefore, the data rate is mainly dependent upon symbol duration T_s.

(iv) A ZF prefiltering scheme for MSI suppression is proposed in [10], which forces received interference to zero within the whole symbol duration. Since our ZF space-time precoding only forces the received interference to zero at the sampling time within one symbol duration, the proposed precoding scheme needs less degree of freedom than ZF prefiltering. For example, when $N_t \le N_r$, our scheme can work well, but ZF prefiltering is inapplicable, and it needs more transmit antennas.

4. Channel Estimation Algorithm

As indicated in last section, the operation of the canceling interferences requires knowledge of the channels. This information must be provided by channel estimation. In this section, we address the channel estimation problem for TR-MIMO-UWB system.

The reciprocity of UWB channel has been experimentally demonstrated in [12]. Consequently, the channel from transmitter to receiver can be estimated by sending training symbols from the receiver and performing channel estimation algorithm at the transmitter. This scheme shuns the implementation of feedback channel which is unfeasible in UWB system. The gist of the proposed algorithm is that the channel is sounded by sending pilot pulses. During the estimation process, the ISI is avoided by letting the pulse repetition interval be larger than T_g, and the MSI is avoided by using orthogonal training symbols. An orthogonal training symbol set is defined as $\mathcal{A} = \{\mathbf{a}_p\}_{p=1}^{N_r}$, where each training symbol is represented as a vector with elements $\{a_{p,n}\}_{n=1}^{N_r}$ taking values of ± 1 and the orthogonality of the set guarantees the relationship

$$\mathbf{a}_p \cdot \mathbf{a}_{p'} = \sum_{n=1}^{N_r} a_{p,n} a_{p',n} = N_r \delta(p - p') \tag{27}$$

holds. In the initialization stage of one block, the training symbol $\mathbf{a}_p$ is sent by the pth antenna at receiver. The training pulses waveform radiated by the pth antenna at receiver is expressed as

$$s^t_p(t) = \sum_{n=1}^{N_r} a_{p,n} \omega(t - (n-1)T'_s), \tag{28}$$

for $p = 1, 2, \ldots, N_r$. In (28), the repetition interval of the training pulses $T_s{}'$ is larger than T_g to avoid interference between training pulses.

The training pulses waveform received by the qth antenna at transmitter is written as

$$r^t_q(t) = \sum_{p=1}^{N_r} s^t_p(t) \otimes g_{q,p}(t) + n_q(t), \tag{29}$$

where $g_{q,p}(t) = \sum_{l=0}^{L_g-1} \alpha_l^{q,p} \delta(t - l\Delta)$ is the channel between the qth antenna at transmitter and the pth antenna at receiver. The transmitter correlates and samples at every Δ time instant on the received training pulses waveform to get

$$v_q(n, l) = \int_{l\Delta}^{(l+1)\Delta} r^t_q(t + (n-1)T'_s)\omega(t)dt. \tag{30}$$

Substituting (29) into (30), we have

$$v_q(n,l) = \sum_{p=1}^{N_r} a_{p,n}\alpha_l^{q,p} + N_q(n,l), \tag{31}$$

where $N_q(n,l) = \int_{l\Delta}^{(l+1)\Delta} n_q(t + (n-1)T_s')\omega(t)dt$ is zero mean Gaussian noise with variance $N_0/2$. The estimated fading coefficient of the channel between the qth antenna of transmitter and the p'th antenna of receiver can be obtained by

$$\hat{\alpha}_l^{q,p'} = \frac{1}{N_r}\sum_{n=1}^{N_r} a_{p',n}v_q(n,l), \tag{32}$$

for $q = 1,2,\ldots,N_t$ and $p' = 1,2,\ldots,N_r$. Inserting (31) into (32) and using (27), we have

$$\hat{\alpha}_l^{q,p'} = \alpha_l^{q,p'} + N_{q,p'}(l), \tag{33}$$

where $N_{q,p'} = (1/N_r)\sum_{n=1}^{N_r} a_{p',n}N_q(n,l)$ the estimation noise with zero mean and variance $N_0/2N_r$. The training symbols can be repeated to send N_c times to get N_c estimations of each fading coefficient. Then, N_c estimation results are averaged to reduce the estimation noise. The result of the averaged estimated fading coefficient is $\overline{\hat{\alpha}}_l^{q,p'} = \alpha_l^{q,p'} + \overline{N}_{q,p'}$, where $\overline{N}_{q,p'}$ is the averaged estimation noise with variance $N_0/2N_cN_r$. The corresponding estimated CIR is $\hat{g}_{q,p'}(t) = \sum_{l=0}^{L_g-1}\overline{\hat{\alpha}}_l^{q,p'}\delta(t - l\Delta)$. Since the UWB short-range applications always occur in the indoor entironment, where the surrounding objects and UWB transceiver are nearly quiescent [17, 18], the coherent time of channel is very long. Therefore, we can increase N_c to reduce the estimation noise within the channel coherent time; however, this will result in a data throughput reduction. When N_c goes to infinity, the estimation noise goes to zero and the estimated channel tends to perfection. The impact of channel estimation on TR-MIMO-UWB system with ZF precoding is investigated by simulations in Section 5.

5. Simulation Results

In this section, simulations and comparisons are performed to validate the proposed scheme. In all cases, the MIMO-UWB channel is generated according to IEEE 802.15.3a channel model recommendation CM4 [19] and truncated to $T_g = 100$ ns. Although the channel model CM4 is designed for single-input single-output (SISO) scenario, the extension to a MIMO configuration is achieved by assuming that the MIMO channel parameters are independent and identically distributed realizations from the same statistical model. The used impulse shape is the second derivative of a Gaussian function $\omega(t) = A_H(1 - 4\pi(t/\tau_m)^2)\exp(-2\pi(t/\tau_m)^2)$, where A_H is the energy normalized parameter and $\tau_m = 0.2788$ ns is the pulse shaping parameter. The duration of $\omega(t)$ is set as $T_\omega = 0.5$ ns so that the minimum multipath resolution of channel is $\Delta = 0.5$ ns.

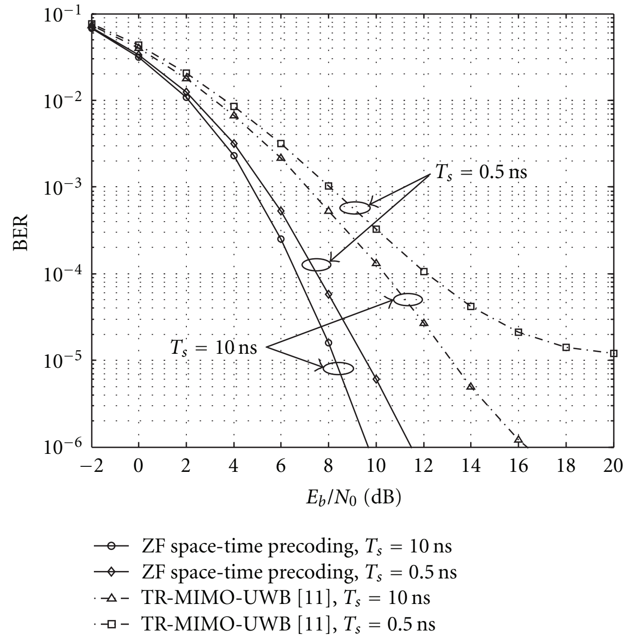

Figure 2: BER performance comparison between the proposed TR-MIMO-UWB system with space-time precoding and the TR-MIMO-UWB system. $N_t = 2$, $N_r = 2$.

TEST 1: BER Performance Comparison between the Proposed Scheme and the Spatial Multiplexed TR-MIMO-UWB System Proposed in [11]. First, we evaluate the BER performance of TR-MIMO-UWB system with ZF precoding proposed in this paper and compare it with the spatial multiplexed TR-MIMO-UWB system proposed in [11]. In this case, both the transmitter and receiver are equipped with $N_t = N_r = 2$ antennas and $N_r = 2$ parallel data streams are transmitted from transmitter simultaneously. The symbol duration T_s is set as 0.5 ns and 10 ns, respectively, which are much smaller than $T_g = 100$ ns. L is set as 200. In this test case, we assume the CSI is perfect. The BER versus E_b/N_0 curves are plotted in Figure 2. It is observed that the BER performance is improved by the proposed space-time precoding scheme. When ISI is strong ($T_s = 0.5$ ns), the BER curve of the spatial multiplexed TR-MIMO-UWB system [11] suffers a floor at high E_b/N_0, while the proposed scheme can obtain a remarkable gain. When $T_s = 0.5$ ns, $M_1 = \lfloor (T_g + T_\omega)/T_s \rfloor = 201$ and $M_2 = \lfloor T_g/T_s \rfloor = 200$. We can compute the bit rate $R_b = N_rL/(M_1 + L + M_2)T_s \approx 4/3$ Gbps.

TEST 2: BER Performance Comparison between the Proposed Scheme and ZF Prefiltering Scheme [10]. Then, the comparison between the proposed scheme and ZF prefiltering scheme [10] is given. In order to meet the needs of degree of freedom for ZF prefiltering, we set the parameters $N_t = 4$, $N_r = 2$, and the length of prefiltering 400 chips. The CSI is perfect for both schemes. From Figure 3, the proposed scheme outperforms ZF prefiltering in terms of BER when both schemes choose the same deployment of antenna ($N_t = 4$, $N_r = 2$). The proposed precoding scheme focuses energy

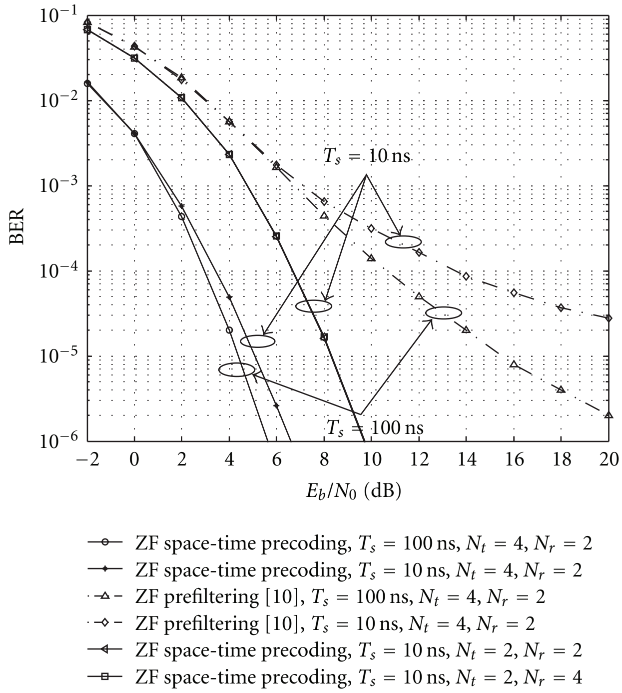

Figure 3: BER performance comparison between the proposed ZF-based space-time precoding for TR-MIMO-UWB system and ZF-based prefiltering scheme.

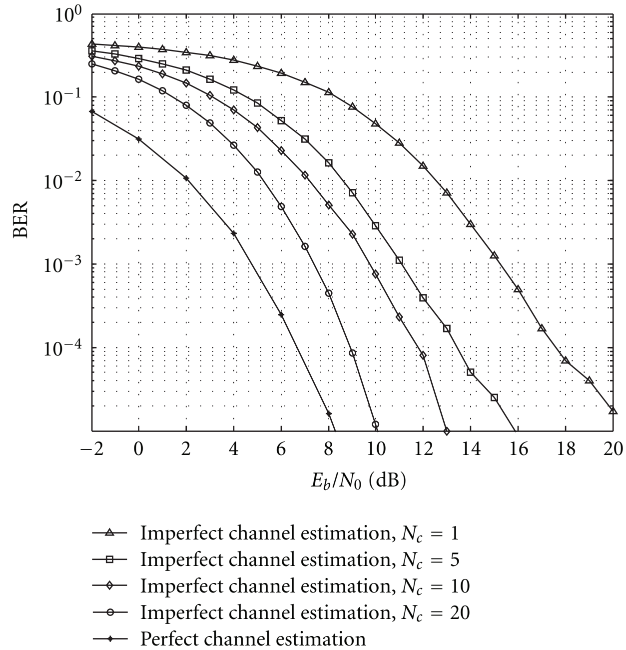

Figure 4: The impact of channel estimation on the proposed scheme. $N_t = 2$, $N_r = 2$, and $T_s = 10$ ns.

on the sampling time to eliminate interferences and ignores other time; therefore, it has higher energy efficiency than ZF prefiltering. Since ZF prefiltering does not consider ISI, the BER curve suffers a floor at high E_b/N_0 when ISI is severe. In order to show that the proposed scheme demands less degree of freedom than ZF prefiltering, the BER performances of the proposed scheme when $N_t = 2$, $N_r = 2$ and $N_t = 2$, $N_r = 4$ are also evaluated. It can be shown the proposed scheme outperforms ZF prefiltering even though less transmit antennas are used. When more transmit antennas are employed, the proposed scheme obtains a considerable gain due to a higher energy efficiency provided by more degree of freedom. It is worthwhile to point out that when $N_t = 2$ and $N_r = 4$, the proposed scheme can transmit $N_r = 4$ parallel data streams normally, this is not in common with ordinary MIMO systems. It is shown in Figure 3 that the performance of the proposed scheme with $N_t = 2$, $N_r = 2$ and that with $N_t = 2, N_r = 4$ are uniform. This is because the same number of transmit antennas offers the same degree of freedom to eliminate interference and results in the same performance. However, the data rate with $N_r = 4$ is twice as high as that with $N_r = 2$.

TEST 3: The Impact of Channel Estimation on the Proposed Scheme. We have so far assumed the CSI is perfect. In this case, the impact of imperfect channel estimation on the proposed scheme is investigated. Both the transmitter and receiver are equipped with $N_t = N_r = 2$ antennas, and $N_r = 2$ parallel data streams are transmitted from transmitter simultaneously. The data symbol duration T_s is set as 10 ns. The channel estimation algorithm proposed in Section 4 is employed in the initialization stage of one block. Orthogonal training symbol set $\mathcal{A} = \{\mathbf{a}_1, \mathbf{a}_2\} = \{\{1,1\},\{1,-1\}\}$ is used. The repetition interval of the training pulse is set as $T_s' = 100$ ns to avoid interference between training pulses. The repetition time of training symbols is set as $N_c = 1, 5, 10$, and 20, respectively. Increasing N_c will improve the accuracy of channel estimation. The simulation results are shown in Figure 4. The BER performance which corresponds to the perfect channel estimation is also plotted. As N_c increases, the BER performance gets better at the price of data throughput reduction. Therefore, there is a tradeoff between performance and data throughput. The imperfect estimation brings out the residual interferences, and the BER curve suffers a floor at high E_b/N_0. That is because the residual interferences become the principal factor to cause error at high E_b/N_0. When $N_c = 20$, the estimation noise is small, and the corresponding BER is close to that of perfect channel estimation.

TEST 4: BER Performance Comparison between the Proposed Scheme and Other Schemes When Imperfect CSI Presents. Finally, the dependence of three schemes (the proposed scheme, the spatial multiplexed TR-MIMO-UWB system [11] and ZF prefiltering scheme [10]) on channel estimation is investigated. The system parameters are set as follows: $N_t = 4$, $N_r = 2$, and $T_s = 10$ ns. The orthogonal training symbol set used to execute channel estimation algorithm is the same as TEST 3 and $T_s' = 100$ ns. The repetition time of training symbols is set as $N_c = 1, 5$, respectively. Figure 5 presents the simulation results. When the transmitter can only use imperfect CSI to implement preprocessing (this comes nearer to practical situation), the improvement of BER performance obtained by the proposed scheme is

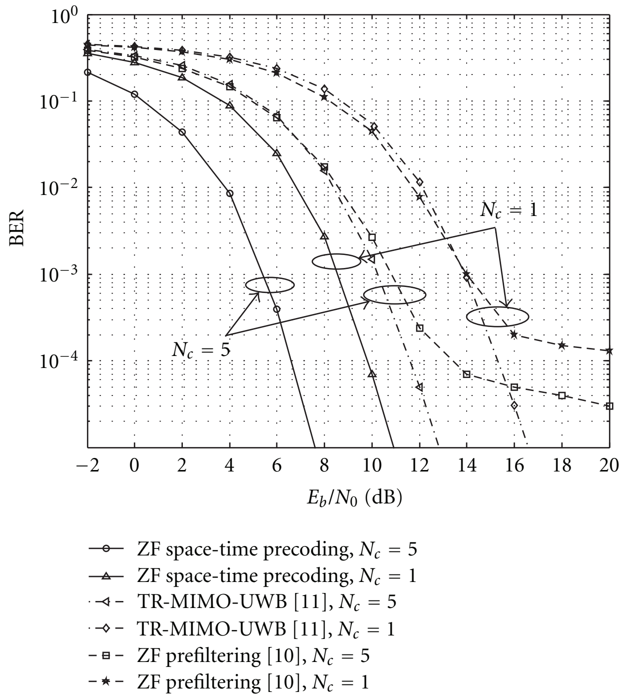

Figure 5: BER performance comparison between the proposed scheme and other schemes when imperfect CSI presents. $N_t = 4$, $N_r = 2$, and $T_s = 10$ ns.

remarkable. From Figure 5, when imperfect CSI presents, the performance of the proposed scheme is also solid and outperforms other schemes. Notably, the performance of the proposed scheme with $N_c = 1$ still outperforms the two other schemes with $N_c = 5$. That is because the CSI is more effectively used to cancel the interferences by the proposed scheme, and the residual interferences are least. The spatial multiplexed TR-MIMO-UWB system [11] can suppress the ISI and MSI to a certain extend, and it shows some robustness to imperfect CSI. Since ZF prefiltering scheme leaves ISI out of consideration [10], its performance becomes worst at high E_b/N_0, where the ISI and the residual MSI are strong.

6. Conclusion

An ultra-high data rate TR-MIMO-UWB system with space-time precoding is proposed in this paper. After the system model of TR-MIMO-UWB is investigated, the computation of the ZF criterion-based space-time precoding matrix is originally derived. With less demand for degree of freedom than other schemes, the proposed space-time precoding scheme can effectively eliminate both ISI and MSI. As a result, the TR-MIMO-UWB system achieves ultra-high data rate of the order of Gbps and keeps BER performance well. The performance of the proposed scheme is evaluated through computer simulations. It is shown that the proposed scheme outperforms the spatial multiplexed TR-MIMO-UWB system and ZF prefiltering scheme. A simple but effective channel estimation algorithm is proposed to provide the estimated CSI for preprocessing. The impact of channel estimation on the proposed scheme is also investigated by simulations. The results confirm that the CSI is more effectively used to remove the interferences by the proposed scheme.

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