Space-Time Diversity Schemes with Blind Channel Estimation for TH-PPM UWB MIMO Communications

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Abstract —For a time-hopping (TH) pulse position modulation (PPM) based ultra-wideband (UWB) multi-antenna system, we propose four coherent space-time (ST) diversity schemes, corresponding to which the correlation template signals with blind channel estimation are also presented. First of all, the issue of channel identifiability based on first-order statistics is addressed, aspired by the single-antenna case [1]. Then, we shed light on that the determination of space-time structure is related to both channel estimation and signal detection. After that, four ST diversity structures are presented to guarantee both the feasibility of blind channel estimation and no multi-antenna interference introduced into signal detection. Finally, substantial simulations show that the proposed ST diversity schemes can significantly outperform the traditional single-antenna system especially for high signal-to-noise ratios whether channel state information is perfect or not.

Index Terms—Ultra-wideband; MIMO; blind channel estimation; first-order statistics; space-time diversity; coherent detection

I. Introduction

Recently, it comes to light that ultra-wide band (UWB) wireless communication has been the subject of extensive research due to its unique capabilities and potential applications. The most attractive point of UWB technology is the ultra-wide bandwidth, which brings about such-and-such advantages. UWB techniques are expected to be applied extensively in practical environments, e.g., wireless personal area networks (WPANS), sensor networks, imaging systems and vehicular radar systems so on [2]. Based on the different signal formats, the candidate transmission and detection methods for the physical layer of UWB communications include three categories: the single carrier impulse radio (IR) using the direct sequence technology introduced; the multiband OFDM technology supported by Multiband OFDM Alliance; a cyclic prefixed single-carrier technique using block transmission with frequency-domain equalization [3]. However, due to the ultra-low transmission power density, the data transmission rates of UWB are very limited. Since multiple-input multiple-output (MIMO) technology boosts the channel capacity largely and provides big performance improvement with low power consumption, it is one viable solution to solve the power limit problem. UWB systems combined with the MIMO technique can provide increased data rates and range so as to achieve the goal of more than 1 Gb/s rate [4]-[8]. Thus, one of our goals in this work has been to investigate several space-time (ST) diversity schemes in the case of a pulse based UWB system over the extremely frequency selective wireless channel.

Diversity transmission technology of Alamouti's space-time block-coding (STBC) scheme [9] has been firstly proposed for flatly fading MIMO systems so as to achieve full spatial diversity at a full rate. After that, an Alamouti-like scheme for SC-FDE systems [10] is developed so as to achieve significant diversity gains over frequency-selective fading channels. To guarantee the performance improvement, the knowledge of instant channel state information (CSI) is required to implement the coherent detection of the received signal in these networks [11]. However, the CSI is unknown and needs to be estimated in the receiver before signal detection. Generally, the IR-UWB systems have two basic signal modulation formats: one is time-hopping (TH) pulse position modulation (PPM); the other is direct sequence (DS) pulse amplitude modulation (PAM). Since DS-PAM solution has the extremely similar signal structure to DS CDMA, the traditional channel estimation methods proposed for DS-CDMA can be extended into the DS-PAM UWB case directly. Hence, current researches on channel estimation focus on the TH-PPM UWB systems, which can be classified into three categories: training or pilot-based non-blind estimation technique [12]-[14], blind estimation technique [15], [16] and semi-blind estimation technique [17], [18].

The main advantage of blind estimation methods over non-blind methods is that no known signal like training or pilot is required. Since both the mean and covariance of the received TH-PPM signal contain the information of channel parameters, blind channel estimation methods could be presented based on the second-order statistics (SOS) [19], [20] or even the first-order statistics (FOS) [1] of received signal. In [1], a blind channel estimation method based on the FOS is developed for the TH-PPM UWB system with single antenna. Unlike the SOS based

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methods, it has not only a low complexity but also avoids the ambiguity problem and slow convergence rate of the estimated channel. Nevertheless, in [1], the reasonable choice of tolerance threshold in channel estimation is not taken into consideration over the large signal-to noise ratio (SNR) range. In the event that the tolerance threshold is set overlarge, the estimated channel frequency response (CFR) coefficients smaller than the tolerance threshold will be regarded as noise and filtered out. On the contrary, the overmuch noise will be retained in the estimated CFR coefficients so as to secure a high mean square error (MSE). As such, the other goal of our paper is to propose a blind channel estimation method with a low complexity based on FOS for MIMO systems. Besides, different to the SISO case, since the received signal contains the all signals from transmitters through channels, we not only need to discuss the existence of channel estimation solution but also consider the multi-antenna interference in coherent signal detection for the proposed UWB MIMO systems. Thus, the main contributions of this paper are described as follows:

- A general transceiver for TH-PPM based UWB with multiple antennas is presented, where blind channel estimation is proposed based on FOS and coherent signal detection is also implemented.
- Channel identification conditions on blind estimation based on FOS are provided under the consideration of the influence of multi-antenna interference, and correspondingly, four ST diversity schemes are proposed for TH-PPM UWB MIMO systems.
- The influence of tolerance threshold on channel estimation performance is discussed here and the optimal values varying from SNR and the data record length can be determined by the numerical simulations. Thus, channel estimation performance can be improved further;
- The correlation template signals are given for realizing the coherent reception for each branch at the receiver. After that, simple equal gain combining (EGC) method is adopted to achieve receiver diversities.

The rest of this paper is structured as follows. Section II first presents the system model of TH-PPM-based UWB MIMO. Then, Section III provides the details of the proposed blind channel estimation algorithm and the corresponding studies on channel identification are also presented. Four ST diversity schemes and their signal detection methods are given in Section IV and Section V, respectively. In Section VI, substantial simulation results are presented for demonstrating the validness of our studies. Finally, conclusions are drawn in Section VII.

II. System Model of TH-PPM Based UWB MIMO

We consider a single-user UWB environment equipped with receiver antennas, where for simplicity, the case of two transmit antennas is only taken into account. Multiple antennas' extension is direct by applying the orthogonal design principle. Every UWB signal is composed of short length pulses of width T_p in the order of nanoseconds (ns). The TH-PPM format using 2-binary is applied to our proposed UWB MIMO method. The two transmitted spectral shaping pulses may be alike, or different. The delays between the transmitters and the receivers are assumed to be zero for simplicity. The transmitted spectral shaping pulse at the m^{th} transmit antenna can be represented as [12]

$$p_m(t)=\sum_{j=0}^{N_f-1} w_m(t-jT_f-c_jT_f), m=1,2 \qquad (1)$$

where $w_m(t)$ is the monocycle at the mth transmit antenna, which is normalized to have unit energy, T_f is the frame duration with $T_f \gg T_p$, N_f is the number of frames per symbol and $\{c_f, f=0,1,\cdots N_f-1\}$ is the user-specific code between 0 and 1, i.e. time-hopping code in order to avoid the catastrophic collisions in the multiple access environment. Furthermore, if $w_1(t)\neq w_2(t)$, then we assume that two shape pulses must be mutually orthogonal, i.e.,

$$\int_0^{T_f} w_1(t)w_2(t)dt=0 \qquad (2)$$

and the minimum time resolution bin equals to pulse duration with T_p. Then, The transmitted waveform at the transmit antenna is expressed as [1], [21]

$$\begin{aligned} s_m(t) &= \sum_{i=-\infty}^{\infty} s_m^i(t) \\ &= \sqrt{E_s}\, p_m^e(t) * \sum_{i=-\infty}^{\infty} \delta(t-2iT_s-b_m^{2i}\Delta) \\ &+ \sqrt{E_s}\, p_m^o(t) * \sum_{i=-\infty}^{\infty} \delta(t-(2i+1)T_s-b_m^{2i+1}\Delta) \end{aligned} \qquad (3)$$

where $s_m^i(t)$ represents the information symbol of the m^{th} transmit antenna at the ith time instant, the symbol energy E_s is normalized to one and is omitted herein, $p_m^e(t)=p_m^i(t),\ i\in\text{even}$ and $p_m^o(t)=p_m^i(t),\ i\in\text{odd}$ is the transmitted spectral shaping pulse of even symbol instant and odd symbol instant at the mth transmit antenna, respectively. T_s is the duration of one symbol per transmit, which is equal to N_fT_f , b_m^i 's are the ith equip-probable binary data symbols with the value of 0 or 1 from the mth transmit antenna, and a PPM parameter Δ is a fixed constant. Usually, $T_p \le \Delta < T_f$ is assumed. Finally, $\delta(t)$ is the Dirac delta function.

The impulse response of the channel between the mth transmit antenna and the nth receive antenna can be modeled as [21], [22]

$$h_{mn}(t)=\sum_{l=0}^{L-1}\theta_l^{mn}\beta_l^{mn}\delta(t-lT_p),\ m=1,2;\ n=1,2,\cdots,N_r \qquad (4)$$

where L is the number of resolvable multipath components, the parameter θ_l^{mn} is the value of -1 or 1 with equal probability 0.5 accounting for the random pulse inversion, β_l^{mn} denotes the sub-channel fading amplitude which is lognormally distributed with a standard deviation of 3-5dB [22], and the minimum multipath resolution is equal to the pulse-width T_p. Then, the power of the lth multipath component can be modeled as $E\left\{\left|h_{mn}(l)\right|^2\right\}=\Omega_0\exp(-\rho l)$, where ρ is the power decay factor and Ω_0 is the normalized factor of channel power.

The information symbol waveforms are transmitted according to the certain encoding form. The encoding sequence needs to satisfy the conditions on channel identification and signal detection (as seen later). At a given symbol period, two signals are simultaneously transmitted using the two different or same pulses from the two antennas. Then, the received signal at the n^{th} receive antenna is given by

$$y_n(t)=\sum_{m=1}^{2}s_m(t)*h_{mn}(t)+v_n(t), n=1,\ldots,N_t \quad (5)$$

where $v_n(t)$ is the additive white Gaussian noise AWGN with zero mean and a two-sided power spectral density of $N_0/2$ at the n^{th} receive antenna and '*' denotes linear convolution. For simplicity, the MAI is ignored in (5). Thus, similar to [1], let $x_{mn}(t)=p_m(t)*h_{mn}(t)$, the received signal of (5) can be expressed by

$$y_n(t)=\sum_{i=-\infty}^{\infty}y_n^i(t), n=1,2,\cdots,N_t \quad (6)$$

where

$$y_n^i(t)=\sum_{m=1}^{2}x_{mn}(t-iT_s-b_m^i\Delta)+v_n^i(t), n=1,2,\cdots,N_t \quad (7)$$

and $v_n^i(t)$ represents the AWGN over the ith symbol period. It is assumed that the duration of every sub-channel is less than symbol duration; in other words, the effect of ISI can be ignored. Perfect timing is also assumed. Blind channel estimation will be completed based on $2I$ consecutive received symbols across which all sub-channels' fading remains constant. Subsequently, they can be employed to construct the template signals for complementing the coherent detection in the receiver.

III. Blind Channel Estimation

Aspired by the method reported in [1], the proposed blind channel estimation method is based on the FOS of the received signal, i.e. the statistical expectation of the received signal across the multiple symbols. The average received signal $y_n(t)$ at the nth receive antenna over the information symbols b_m^i's is expressed by

$$\begin{aligned}\bar{y}_n(t)&=E\{y_n(t)\}\\&=\sum_{m=1}^{2}\sum_{i=-\infty}^{\infty}\frac{1}{2}\left[x_{mn}(t-iT_s)+x_{mn}(t-iT_s-\Delta)\right]\\&\text{for } n=1,2,\cdots,N_t\end{aligned} \quad (8)$$

With the aid of the PPM signal format, the expectation of the received signal is periodic with the period T_s. Then, by $x_{mn}(t)=p_m(t)*h_{mn}(t)$, we have

$$\bar{y}_n(t)=\frac{1}{2}\sum_{m=1}^{2}\left[\left(p_m(t)+p_m(t-\Delta)\right)*\tilde{h}_{mn}(t)\right],\ n=1,2,\cdots,N_t \quad (9)$$

where

$$\tilde{h}_{mn}(t)=\sum_{i=-\infty}^{\infty}h_{mn}(t-iT_s),\ m=1,2;\ n=1,2,\cdots,N_t \quad (10)$$

which is a periodically extended channel of $h_{mn}(t)$. Now, let

$$pa_m(t)=\frac{1}{2}\left(p_m(t)+p_m(t-\Delta)\right),\ m=1,2 \quad (11)$$

and

$$pa_m(t)=\sum_{i=-\infty}^{\infty}pa_m(t-iT_s),\ m=1,2 \quad (12)$$

Then, we can rewrite $\bar{y}_n(t)$ in (8) as

$$\bar{y}_n(t)=\sum_{m=1}^{2}pa_m(t)\otimes\tilde{h}_{mn}(t),\ n=1,2,\cdots,N_t \quad (13)$$

where '$\otimes$' denotes the circular convolution operation. Then, the Fourier transform of both sides of (13) can be given by

$$\bar{Y}_n[k]=\sum_{m=1}^{2}Pa_m[k]H_{mn}[k],\ n=1,2,\cdots,N_t;k=0,\pm1,\ldots \quad (14)$$

Notably, the method of [1] cannot be applied directly to estimate the UWB MIMO channel blindly since the equation (14) is hard to be solved due to too many unknown channel parameters. Hence, we both utilize the time diversity and space diversity of the MIMO system to solve this problem. Thus, the received signals of the even instants and the odd ones from (14) are collected by

$$\begin{aligned}\bar{Y}_n^e[k]&=\frac{1}{I}\sum_{i=0}^{I-1}Y_n^{2i}[k]\\&=Pa_1^e[k]\tilde{H}_{1n}[k]+Pa_2^e[k]\tilde{H}_{2n}[k]\\&\text{for } n=1,2,\cdots,N_t;\ k=0,\pm1,\ldots\end{aligned} \quad (15)$$

and

$$\begin{aligned}\bar{Y}_n^o[k]&=\frac{1}{I}\sum_{i=0}^{I-1}Y_n^{2i+1}[k]\\&=Pa_1^o[k]\tilde{H}_{1n}[k]+Pa_2^o[k]\tilde{H}_{2n}[k]\\&\text{for } n=1,2,\cdots,N_t;\ k=0,\pm1,\ldots\end{aligned} \quad (16)$$

respectively, where

$$pa_m^e(t) = \sum_{i=0}^{I-1} pa_m(t - 2iT_s),\ m = 1,2 \quad (17)$$

and

$$pa_m^o(t) = \sum_{i=0}^{I-1} pa_m(t - (2i+1)T_s),\ m = 1,2 \quad (18)$$

Since the sub-channels are assumed invariant during the $2I$ consecutive received symbols, we have $\tilde{H}_{mn}^i[k] = \tilde{H}_{mn}^{i+1}[k] = \tilde{H}_{mn}[k],\ m = 1,2;\ n = 1,2,\cdots,N_t$; $k = 0,\pm1,\cdots$. $\bar{Y}_n^e[k]$ and $\bar{Y}_n^o[k]$ represents the arithmetic average values of the frequency domain received signals over the I even information symbols $\{Y_n^0[k], Y_n^2[k],\cdots,Y_n^{2I-2}[k]\}$ and the I odd ones $\{Y_n^1[k], Y_n^3[k],\cdots,Y_n^{2I-1}[k]\}$ at the nth receive antenna, respectively. Then, (15) and (16) can be rewritten by a matrix form as follows

$$\underbrace{\begin{bmatrix} \bar{Y}_n^e[k] \\ \bar{Y}_n^o[k] \end{bmatrix}}_{\bar{\mathbf{Y}}_n[k]} = \underbrace{\begin{bmatrix} Pa_1^e[k] & Pa_2^e[k] \\ Pa_1^o[k] & Pa_2^o[k] \end{bmatrix}}_{\mathbf{Pa}[k]} \underbrace{\begin{bmatrix} \tilde{H}_{1n}[k] \\ \tilde{H}_{2n}[k] \end{bmatrix}}_{\tilde{\mathbf{H}}_n[k]}, k = 0,\pm1\cdots \quad (19)$$

Then, the channel vector $\tilde{\mathbf{H}}_n[k] = [\tilde{H}_{1n}[k], \tilde{H}_{2n}[k]]^{\mathrm{T}}$ $k = 0,\pm1,\cdots$ at the kth frequency component of the nth receiver can be estimated by solving (19) as

$$\tilde{\mathbf{H}}_n^e[k] = \mathbf{Pa}^{-1}[k]\bar{\mathbf{Y}}_n[k], k = 0,\pm1,\cdots,n = 1,2,\cdots,N_t \quad (20)$$

Notably, the determinant of the matrix $\mathbf{Pa}[k]$ is required to be nonzero so that the matrix $\mathbf{Pa}[k]$ is invertible. In other words, the solution (20) of (19) exists only if the following channel identification conditions can be satisfied by

$$Pa_1^e[k]Pa_2^o[k] \neq Pa_1^o[k]Pa_2^e[k],\ k = 0,\pm1,\cdots \quad (21)$$

Similar to [1], in order to avoid the large noise enhancement due to the small elements of $Pa_1^e[k]Pa_2^o[k] - Pa_1^o[k]Pa_2^e[k], k = 0,\pm1,\cdots$, the conditions of (21) can be modified to

$$\left| Pa_1^e[k]Pa_2^o[k] - Pa_1^o[k]Pa_2^e[k] \right| > \gamma,\ k = 0,\pm1,\ldots \quad (22)$$

where '$|\cdot|$' stands for the absolute value operation and the tolerance threshold γ is a small positive value. The influence of γ on the system performance will be explored as seen later. Then, the channel vector $\tilde{\mathbf{H}}_n[k] = [\tilde{H}_{1n}[k], \tilde{H}_{2n}[k]]^{\mathrm{T}}$, $k = 0,\pm1,\cdots$, can be estimated through (20) if (22) is satisfied. Otherwise, it will be replaced by a zero vector. Also, based on the condition of (21), several space-time diversity schemes are presented in the following subsection. Thus, the proposed blind channel estimation method for the UWB MIMO systems can be described as follows:

1) Design the space-time diversity scheme to guarantee the invertible matrix $\mathbf{Pa}[k]$ according to (21);
2) To compute Fourier series $Pa_1[k]$ and $Pa_2[k]$ of $pa_1(t)$ and $pa_2(t)$ according to (11) and (12), $k = 0,\pm1,\cdots$;
3) To obtain the Fourier series $\hat{\bar{Y}}_n^e(k)$ and $\hat{\bar{Y}}_n^o(k)$ of the expectations of the received signals by (15) and (16)
4) To estimate $\tilde{\mathbf{H}}_n^e[k]$ as (20) if (22) is satisfied; otherwise, $\tilde{\mathbf{H}}_n^e[k]$ is set zero. After that, $\tilde{h}_{mn}^e(t)$ can be obtained by the inverse Fourier transform of $\tilde{\mathbf{H}}_n^e[k]$;
5) Further to reduce the noise effect on the $\tilde{h}_{mn}^e(t)$ by the following low-pass filtering

$$\hat{h}_{mn}(t) = \begin{cases} \tilde{h}_{mn}^e(t), & \text{if} \quad 0 \le t < LT_p \\ 0, & \text{otherwise} \end{cases} \quad (23)$$

IV. Proposed Several Space-Time Diversity Schemes

It should be mentioned that not all of the space-time diversity structures are suitable to the channel identability condition of (21). In this section, several reasonable space-time diversity schemes are presented and analyzed.

A. *Space-Time Diversity Scheme 1 (ST1)*

If one is to let

$$\begin{matrix} Pa_1^e[k] = Pa_1[k] & Pa_2^e[k] = Pa_2[k] \\ Pa_1^o[k] = -Pa_2[k] & Pa_2^o[k] = Pa_1[k] \end{matrix}, \quad (24)$$

and

$$\begin{matrix} b_1^i = b_2^{i+1} \\ b_2^i = b_1^{i+1} \end{matrix}, i = 0,2\cdots,2I-2 \quad (25)$$

then, (21) and (22) can be rewritten by

$$\left(Pa_1[k]\right)^2 + \left(Pa_2[k]\right)^2 \neq 0 \quad (26)$$

and

$$\left| \left(Pa_1[k]\right)^2 + \left(Pa_2[k]\right)^2 \right| > \gamma \quad (27)$$

respectively. There are two possibilities for the monocycle used at each transmitter to meet the equation (26) above. One is to let $w_1(t) = w_2(t)$; the other is $w_1(t) \neq w_2(t)$. Obviously, the first solution is not feasible for the space-time structure (24). This is because the correlation receiver for implementing the coherent signal detection suffers from a severe performance loss due to the multiple antenna interference (MAI) caused by the same monocycles for all transmitters. Therefore, we assume that different transmitters use the different monocycles and they are orthogonal to each other as (2). This scheme is named after ST1 in this paper. Actually, ST1 for $N_t = 2$ is equivalent to the method reported in [23] and similar to the Alamouti-like STBC architecture [10]. Differently, the influence of threshold will be considered as seen later. Besides, another similar space-time diversity structure to (24) is given by

$$\begin{array}{l} Pa_1^e[k] = Pa_1[k],\ Pa_2^e[k] = Pa_2[k] \\ Pa_1^o[k] = Pa_2[k],\ Pa_2^o[k] = Pa_1[k] \end{array} \tag{28}$$

Only if $\left(Pa_1[k]\right)^2 \neq \left(Pa_2[k]\right)^2$ or $\left|\left(Pa_1[k]\right)^2 - \left(Pa_2[k]\right)^2\right| > \gamma$ holds, the estimated channel vector $\tilde{\mathbf{H}}_n^e[k], k = 0, \pm 1, \ldots$ can be derived from (20). Herein, (28) is named after the modified ST1 (mST1) scheme.

B. Space-Time Diversity Scheme 2 (ST2)

It is worthwhile to note that the monocycle for even symbol instant is different from that of odd symbol instant at each transmitter for ST1 and mST1 schemes. Hence, compared to the single antenna case, they can obtain both time and space diversity but at the cost of increasing the complexities of transmitter since two types of monocycle are required for each transmitter branch. For the purpose of overcoming this drawback, we have the following space-time diversity scheme 2 (named as ST2),

$$\begin{array}{l} Pa_1^e[k] = Pa_1[k],\ Pa_2^e[k] = Pa_2[k] \\ Pa_1^o[k] = -Pa_1[k],\ Pa_2^o[k] = Pa_2[k] \end{array} \tag{29}$$

and

$$\begin{array}{l} b_1^i = b_1^{i+1} \\ b_2^i = b_2^{i+1} \end{array},\ i = 0, 2 \cdots, 2I-2 \tag{30}$$

Notably, since each transmitting antenna makes use of the same monocycle for all symbol instants, ST2 will be realized more easily than ST1 and mST1. Nevertheless, for a good signal recovery, $w_1(t)$ and $w_2(t)$ are still assumed orthogonal to each other as (2) so that the MAI is avoided. From (29), (21) and (22) can be given by

$$Pa_1[k] Pa_2[k] \neq 0 \tag{31}$$

and

$$\left|2Pa_1[k] Pa_2[k]\right| > \gamma \tag{32}$$

respectively. It can be easily observed that the existence condition of channel estimation solution in ST2 is stricter than that of ST1 or mST1.

C. Space-Time Diversity Scheme 3 (ST3)

Further to simplify the transmitter, we introduce the another ST diversity scheme (named after ST3)

$$\begin{array}{l} Pa_1^e[k] = Pa_1[k],\ Pa_2^e[k] = Pa_1[k] \\ Pa_1^o[k] = -Pa_1[k],\ Pa_2^o[k] = Pa_1[k] \end{array} \tag{33}$$

and

$$\begin{array}{ll} b_1^i = b_2^i,\ for & i = 0, 1, \cdots, 2I-1 \\ b_m^i \neq b_m^{i+1},\ for & i = 0, 2 \cdots, 2I-2;\ m = 1, 2 \end{array} \tag{34}$$

where only one monocycle ($w_1(t) = w_2(t)$) is employed for all transmitters and hence no MAI exists. Moreover, the interference between two different monocycles in the schemes above can be avoided. Equation (34) is assumed to ensure the coherent signal detection without MAI. If using (33), (21) and (22) can be expressed by

$$2Pa_1^2[k] \neq 0 \tag{35}$$

and

$$\left|2Pa_1^2[k]\right| > \gamma \tag{36}$$

respectively. The threshold γ is a small positive number. In the subsequent simulations, the influence of the choice of γ on the performance will be discussed. Supposing that (35) or (36) can be satisfied, the channel vector $\tilde{\mathbf{H}}_n[k], k = 0, \pm 1, \cdots$ can be recovered from (20) by the ST3 scheme above.

V. Signal Detection for Different ST Structures

In order to recover the information symbols, the difference signals should be firstly constructed by the estimated channel $\hat{h}_{mn}(t)$ for the correlate receiver. Then, the output of all branches after correlation operations can be combined at certain combination principle such as maximal-ratio combining (MRC) and EGC, etc. For simplicity, EGC method is adopted here to achieve receiver diversities at a cost of performance loss. Finally, the output signal of the combiner will pass through the decision detector to achieve the estimated value of the transmitted signals.

For all ST diversity schemes above, the difference signals are expressed by

$$x_d^{mn(i)}(t) = \left[p_m^i(t) - p_m^i(t-\Delta)\right] * \hat{h}_{mn}(t) \tag{37}$$

where $p_m^i(t),\ i = 0, 1, 2, \cdots;\ m = 1, 2$ for various space-time diversity schemes are presented in Table I. We define $\mathrm{mod}(m, 2)$ as the operation of modulus-2. Thus, for $i = 0, 2, 4 \cdots 2I-2;\ m = 1, 2$, using EGC in space and time, we obtain the decision statistics of ST1 or mST1 by

$$\begin{aligned} d_m^i = \sum_{n=1}^{N_t} \int_{-\infty}^{\infty} & y_n^i(t) x_d^{mn(i)}(t - iT_s) + \\ & y_n^{i+1}(t) x_d^{(\mathrm{mod}(m,2)+1)n(i+1)}(t - (i+1)T_s) dt \end{aligned} \tag{38}$$

The decision statistics of ST2 under EGC can be expressed by

$$\begin{aligned} d_m^i = \sum_{n=1}^{N_t} \int_{-\infty}^{\infty} & y_n^i(t) x_d^{mn(i)}(t - iT_s) + \\ & y_n^{i+1}(t) x_d^{mn(i+1)}(t - (i+1)T_s) dt \end{aligned} \tag{39}$$

Differently, the decision statistics of ST3 using EGC should be constructed as

$$d_m^i = d_1^i = d_2^i = \sum_{n=1}^{N_t}\sum_{m=1}^{2}\int_{-\infty}^{\infty} y_n^i(t)\, x_d^{mn(i)}(t - iT_s)dt \quad (40)$$

As such, the information symbols b_m^i 's can be detected as the following decision rule

$$\hat{b}_m^i = \begin{cases} 0, & \text{if} \quad d_m^i \geq 0 \\ 1, & \text{otherwise} \end{cases} \quad (41)$$

TABLE I: SPECTRAL SHAPING PULSE ALLOCATED FOR VARIOUS ST SCHEMES

ST Schemes	Spectral Shaping Pulse			
	$p_1^e(t)$	$p_2^e(t)$	$p_1^o(t)$	$p_2^o(t)$
ST1	$p_1(t)$	$p_2(t)$	$-p_2(t)$	$p_1(t)$
mST1	$p_1(t)$	$p_2(t)$	$p_2(t)$	$p_1(t)$
ST2	$p_1(t)$	$p_2(t)$	$-p_1(t)$	$p_2(t)$
ST3	$p_1(t)$	$p_1(t)$	$-p_1(t)$	$p_1(t)$

VI. NUMERICAL SIMULATIONS

In this section, we numerically study the performance of our proposed ST schemes as well as the corresponding blind channel estimation algorithm under TH-PPM based UWB MIMO scenarios. Simulation parameters are provided in Table II. The monocycle waveform $w_1(t)$ is selected as the first derivative of Gaussian pulse, while the monocycle waveform $w_2(t)$ is set by the second derivative of Gaussian pulse. Thus, they are orthogonal to each other. The normalized mean square error (NMSE) of channel estimation is defined as

$$\text{NMSE} = \sum_{m=1}^{2}\sum_{n=1}^{N_t}\left\|\hat{h}_{mn} - h_{mn}\right\|^2 \Big/ L\sum_{m=1}^{2}\sum_{n=1}^{N_t}\left\|h_{mn}\right\|^2 \quad (42)$$

TABLE II: SIMULATION PARAMETERS

Name	**Value/Notation**
System	TH-PPM UWB
Pulse width	T_p=0.5ns
Sampling frequency	2GHz
Modulation	BPSK
Frame duration	T_f=11ns
Frame number/symbol	N_f=12
PPM parameter	Δ =1.5ns
TH code $c_j T_f$	[8,5,0,7,6,6,6,0,3,3,7,5]ns
Multipath number	L=179
σ_{dB}	3dB
Decay factor ρ	0.02

Fig. 1 gives the numerical simulation results of NMSE performance against tolerance threshold γ as E_b/N_0 varies under different data record lengths for the regular SISO scheme. Results indicate that if E_b/N_0 range is set to be large enough, the optimal tolerance threshold certainly exists. Moreover, this optimal value is related to both E_b/N_0 value and data record length. As shown in Fig. 1, for E_b/N_0 =10dB and I=500, γ_{opt} is approximately 0.56 whereas for E_b/N_0 =30dB with I=500 and E_b/N_0 =10dB under I=1000, γ_{opt} changes to about 0.06 and 0.36, respectively. Also, as E_b/N_0 value or data record length I increases, the optimal tolerance thresholds gradually move toward to low value region. In other words, the larger E_b/N_0 value or data record length is, the smaller the optimal tolerance threshold is. It indicates that channel estimation performance can be improved if the tolerance threshold is chosen appropriately.

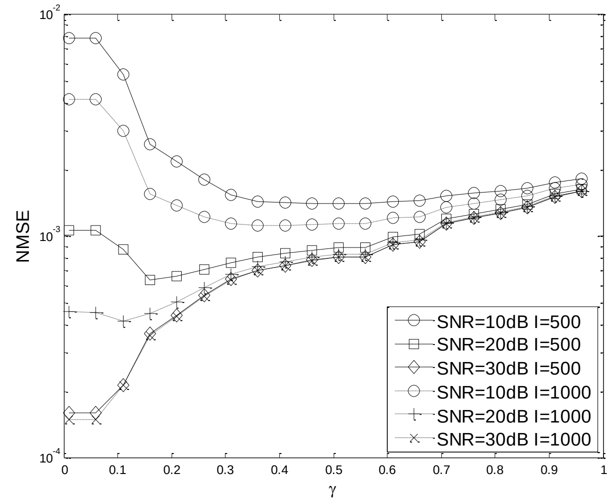

Fig. 1. NMSE against γ of the regular SISO scheme for different SNR and I values.

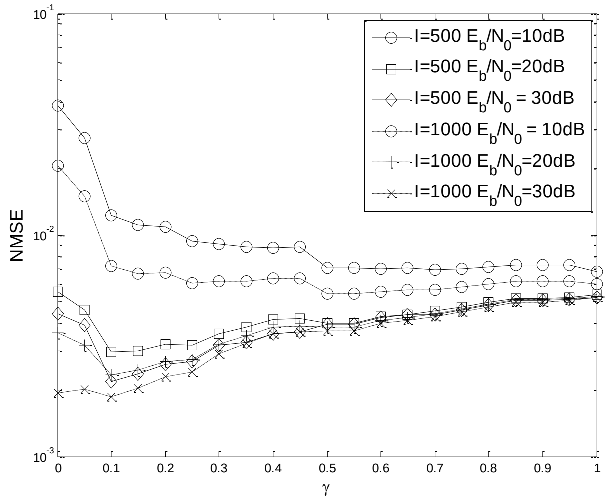

Fig. 2. NMSE against γ of the proposed ST1 scheme for different SNR and I values.

Fig. 2 shows the results for proposed ST1, from which we can observe the identical phenomenon. For the simplicity and space limit of paper, the proposed mST1, ST2 and ST3 schemes are omitted since the same results can be obtained. Correspondingly, blind channel estimation method based on FOS is utilized in different

ST diversity schemes. As a result, the choice of tolerance threshold greatly influences the NMSE performance of blind channel estimation for all proposed schemes.

Table III shows the optimal values of tolerance threshold varying from E_b/N_0 for various ST schemes under different data lengths by adopting same numerical methods above. In the subsequent simulations, these optimal values will be compared to the non-optimal case. This non-optimal case means that only a fixed tolerance threshold is used in blind channel estimation over the whole E_b/N_0 range.

TABLE III: THE OPTIMAL TOLERANCE VALUES FOR DIFFERENT SCHEMES

Schemes	I	E_b/N_0 (dB)		
		10	20	30
SISO	500	0.56	0.16	0.06
	1000	0.36	0.11	0.01
ST1	500	1.00	0.10	0.10
	1000	0.50	0.10	0.10
mST1	500	0.30	0.05	0.01
	1000	0.18	0.02	0.00
ST2	500	0.31	0.09	0.01
	1000	0.21	0.03	0.01
ST3	500	0.30	0.08	0.00
	1000	0.20	0.05	0.00

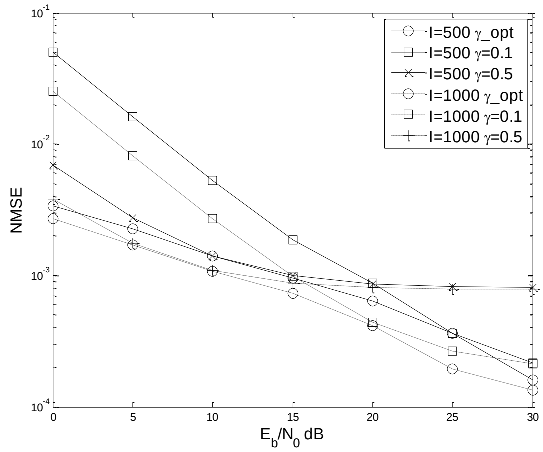

Fig. 3. NMSE against E_b/N_0 of the regular SISO scheme for different γ and I values

Fig. 3 shows the NMSEs of blind channel estimation method [1] for SISO systems with I=500, 1000 as a function of E_b/N_0 for optimal tolerance threshold and non-optimal ones (i.e., 0.1 and 0.5, respectively). In Figs. 4-7, we show the NMSEs of the proposed channel estimation methods for different ST diversity schemes with I=500, 1000 with respect to E_b/N_0, respectively. The NMSE curves of various thresholds (i.e., 0.1 and 0.5) are also compared in these figures. First, it appears that whether for the SISO case or for the MIMO cases, the channel estimation methods with γ_{opt} can gain the best performance over the whole E_b/N_0 range than those with non-optimal tolerance thresholds. This is consistent with our results before. Further, the NMSE performance gaps of γ_{opt} between I=500 and I=1000 are less than those of non-optimal tolerance thresholds. It means that the proposed methods not only boost the channel estimation performance but also gain the low computational complexity by setting the optimal tolerance thresholds.

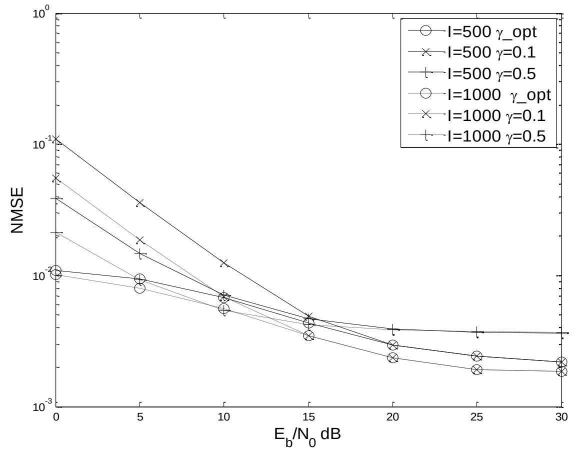

Fig. 4. NMSE against E_b/N_0 of the proposed ST1 scheme for different γ and I values

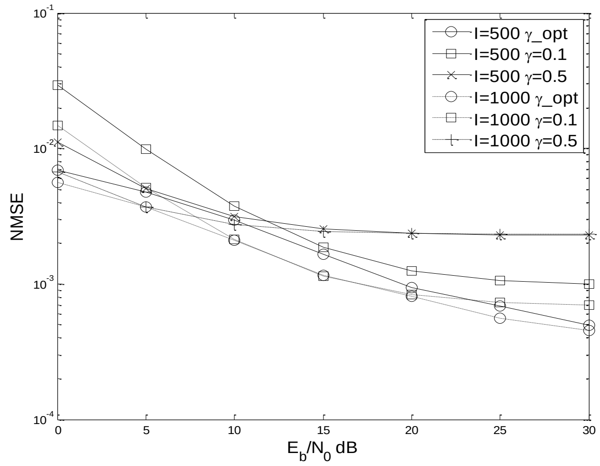

Fig. 5. NMSE against E_b/N_0 of the proposed mST1 scheme for different γ and I values

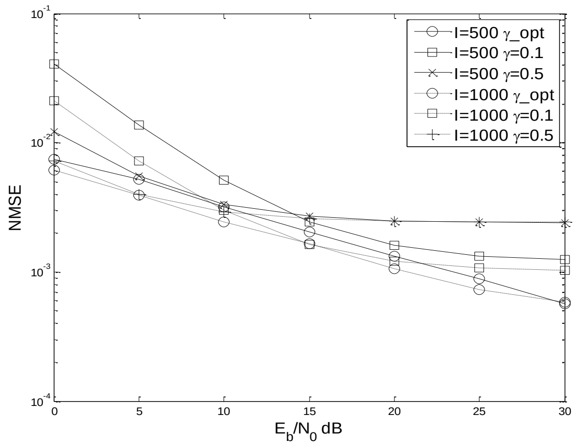

Fig. 6. NMSE against E_b/N_0 of the proposed ST2 scheme for different γ and I values

Fig. 8 shows the NMSE performance of blind channel estimation against E_b/N_0 for the proposed four ST schemes with γ_{opt} and I=500. The SISO case with γ_{opt} is also considered under blind channel estimation. First, it can be seen that the SISO case has the best estimation performance than the others due to much fewer channel estimation parameters. Second, we can see that the proposed mST1 scheme has the closest NMSE performance to the SISO case while the proposed ST1 much worse than the others. The proposed ST2 and ST3 schemes have the almost same performance to the mST1 scheme for $E_b/N_0 < 10\text{dB}$. For $E_b/N_0 < 15\text{dB}$, the proposed ST3 scheme performs better than the proposed ST1 and ST2 schemes, is second only to the proposed mST1 scheme for $15\text{dB} < E_b/N_0 < 25\text{dB}$. Although for $E_b/N_0 > 25\text{dB}$, the proposed ST3 scheme gradually begins to degrade slightly, it is still the best one in the proposed schemes in terms of implementation since it only uses one monocycle pulse for all transmitters as discussed in Section IV.

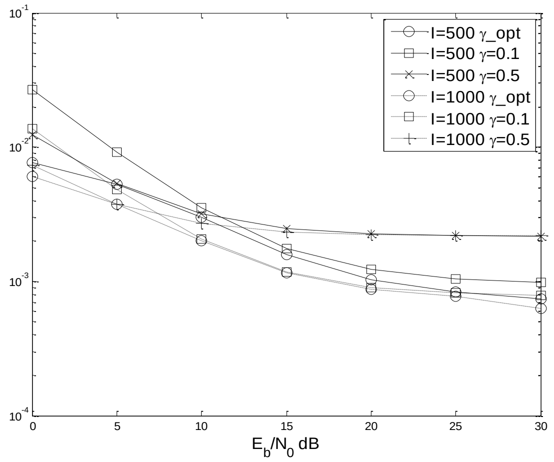

Fig. 7. NMSE against E_b/N_0 of the proposed ST3 scheme for different γ and I values

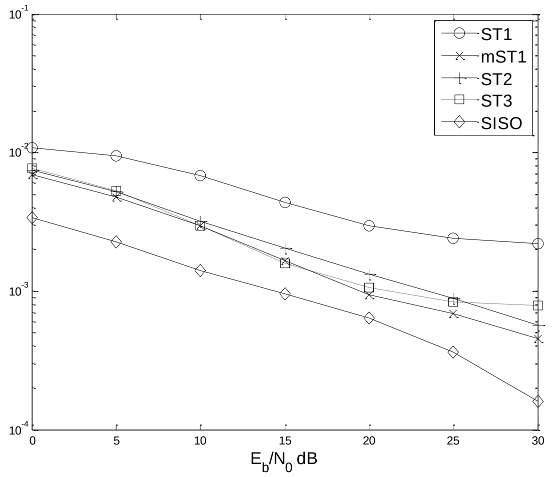

Fig. 8. NMSE against E_b/N_0 for different γ and I values under different schemes

In Fig. 9, we show the BER performance against E_b/N_0 of our developed ST schemes and their comparisons with the conventional SISO cases for I=500, where the case for perfect CSI is considered as a benchmark. Therein, the corresponding correlation template signals are constructed by channel estimates or perfect CSI for coherent detection. It can be noted that the proposed schemes using correlation reception with EGC yield a smaller BER than the SISO cases by exploiting the ST diversity whether CSI is perfect or not. Besides, under the perfect CSI, the proposed ST schemes have the almost identical performance while different in the case of estimated channel. It means that the variation of ST diversity structure in the transmitter almost has a great influence on blind channel estimation based on FOS.

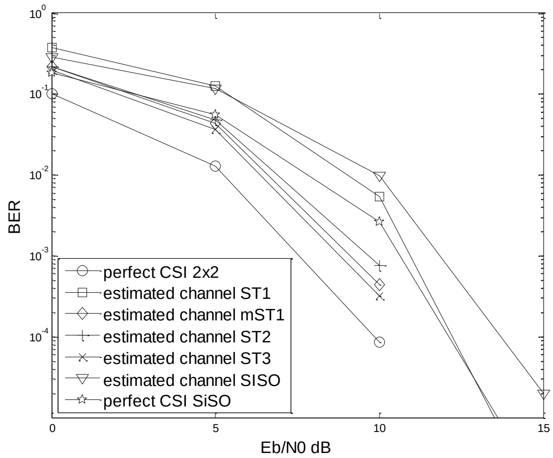

Fig. 9. BER against E_b/N_0 for γ_{opt} and I values under different schemes

VII. CONCLUSIONS

This paper proposed several ST diversity structures for a TH-PPM based UWB MIMO system, where the corresponding correlation receivers with EGC were employed for coherent signal detection. Both channel identification condition and blind channel estimation method based on FOS were also derived. Herein, there was no ambiguity problem, not as much as that of the SOS methods. More importantly, the influence of tolerance threshold on channel estimation was considered over large SNR range and its optimal values were also determined. Simulation results demonstrate that the proposed ST schemes yield a smaller BER value compared with the SISO case.

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REFERENCES

[1] Z. D. Wang and X. F. Yang, "Blind channel estimation for ultra wide-band communications employing pulse position modulation," *IEEE Signal Processing Letters*, vol. 12, no. 7, pp. 520-523, Jul. 2005.

[2] S. Kwon, S. Lee, and J. Kim, "A joint timing synchronization, channel estimation, and SFD detection for IR-UWB systems," *Journal of Communications and Networks*, vol. 12, no. 5, pp. 501-509, 2012.

[3] Y. Yao, X. Dong, and N. Tin, "Design and analysis of timing synchronization in block transmission UWB systems," *IEEE Transactions on Communications*, vol. 59, no. 6, pp. 1686-1696, 2011.

[4] T. Kaiser, F. Zheng, and E. Dimitrov, "An overview of ultra-wide-band systems with MIMO," *Proceedings of the IEEE*, vol. 97, no. 2, pp. 285-312, 2009.

[5] S. Kim, C. Uhm, D. Kang, and K. Kim, "A study on the wireless MIMO-UWB transceiving techniques for WPAN," in *Proc. IEEE International Conference on Consumer Electronics*, 2005, pp. 91-92.

[6] Z. W. Lin, P. Benjamin, and A. S. Madhukumar, "MMSE detection for high data rate UWB MIMO systems," in *Proc. IEEE 60th Vehicular Technology Conference*, 2004, pp. 1463-1467.

[7] C. Wang, T. Lv, H. Gao, *et al.*, "Generalized likelihood ratio test multiple-symbol detection for MIMO-UWB: A semidefinite relaxation approach," in *Proc. IEEE Wireless Communications and Networking Conference*, Istanbul, Turkey, Apr. 2014, pp. 1276-1280

[8] H. Nigam and M. Kumar, "Design and performance analysis of UWB MIMO OFDM system using microstrip antenna," *Advanced Communication Technology*, Korea (South), Feb. 2014, pp. 971-977.

[9] S. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE Journal on Selected Areas in Communications*, vol. 16, no. 8, pp. 1451-1458, 1998.

[10] N. Al-Dhahir, "Single-carrier frequency-domain equalization for space-time-coded transmissions over broadband wireless channels," *IEEE Communications Letter*, vol. 5, no. 7, pp. 304-306, 2001.

[11] V. Sharma and S. Kumar, "Recent developments in MIMO channel estimation techniques," in *Proc*. *IEEE/Digital Information and Communication Technology and its Applications*, Bangkok, Thailand, 2012, pp. 1-6.

[12] V. Lottici, A. D'Andrea, and U. Mengali, "Channel estimation for ultra-wideband communications," *IEEE Journal on Selected Areas in Communications*, vol. 20, no. 9, pp. 1638-1645, 2002.

[13] D. Wang, L. Jiang, and C. He, "Robust noise variance and channel estimation for SC-FDE UWB systems under narrowband interference," *IEEE Transactions on Wireless Communications*, vol. 8, no. 6, pp. 3249-3259, 2009.

[14] C. Vaz, K. Mun, and D. G. Daut, *et al*., "Estimation of communications channels using discrete wavelet transform-based deconvolution," *IEEE Transactions on Communications*, vol. 61, no. 10, pp. 4186-4195, Oct. 2013.

[15] L. Sheng and R. C. D. Lamare, "Blind reduced-rank adaptive receivers for DS-UWB systems based on joint iterative optimization and the constrained constant modulus criterion," *IEEE Transactions on Vehicular Technology*, vol. 60, no. 6, pp. 2505-2518, 2011.

[16] A. Coskun and I. Kale, "All-adaptive blind matched filtering for the equalization and identification of multipath channels-a practical approach," *IEEE Transactions on Circuits and Systems*, vol. 60, no. 1, pp. 232-242, 2013.

[17] J. Cai, X. S. Shen, and J. W. Mark *et al*., "Semiblind channel estimation for pulse-based ultra-wideband wireless communication systems," *IEEE Transactions on Vehicular Technology*, vol. 55, no. 1, pp. 95-103, 2006.

[18] S. Franz, C. Carbonelli, and U. Mitra, "Joint semi-blind channel and timing estimation for generalized UWB transmitted reference systems," *IEEE Transactions on Wireless Communications*, vol. 6, no. 1, pp. 229-232, 2007.

[19] Z. Y. Xu and P. Liu, "A subspace approach to blind estimation of ultra-wideband channels," in *Proc. Thirty-Seventh Asilomar Conference on Signals, Systems and Computers, Pacific Grove*, CA, United States, 2003, pp. 1249-1253.

[20] P. Liu and Z. Y. Xu, "POR-based channel estimation for UWB communications," *IEEE Transactions on Wireless Communications*, vol. 4, no. 6, pp. 2968-2982, 2005.

[21] H. P. Liu, "Error performance of pulse amplitude and position modulated ultra-wideband system over lognormal fading channels," *IEEE Communications Letters*, vol. 7, no. 11, pp. 531-533, 2003.

[22] R. Dilmaghani, M. Hernandez, and M. Ghavami, "Optimum sequence design and performance evaluation of channel estimation techniques for ultra-wideband systems," *IEEE Transactions on Communications*, vol. 152, no. 2, pp. 241-245, 2005.

[23] D. Wang, L. Yang, and J. X. Pu, "TH-PPM UWB systems with space-time diversity based on channel estimation using first-order statistics," *Video Engineering*, vol. 35, no. 17, pp. 93-96, 2011.

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