A MULTIUSER TRANSMITTED REFERENCE UWB TRANSCEIVER FOR HIGH DATA RATE COMMUNICATION

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ABSTRACT

A conventional transmitted reference (TR) transceiver suffers from a data rate penalty and performance loss in the presence of the typically severe multipath distortion. It is also ineffective in suppressing multiuser interference (MUI). Thus a multiuser TR (MTR) UWB transceiver is designed for high data rate communication in this work. A doublet is transmitted in each frame. The first pulse is a reference pulse, but the second pulse carries two bits of user information in amplitude and position simultaneously using bi-orthogonal signaling. Such a hybrid modulation doubles data rate. Since no restrictive constraint is imposed on pulse spacing, data rate can be further increased to four times that of the conventional TR transceiver as pulse spacing approaches zero. To improve resilience to MUI, each pulse is amplitude modulated by a pseudo-random (PN) sequence. Moreover, a time hopping sequence can be applied to control the delay of each data pulse. At the receiver, the signal waveform can be reliably estimated from the noisy received signal despite severe inter-pulse interference and MUI. Then two templates are constructed for simple correlation detection that independently detects both bits, while complicated joint detection is unnecessary [1].

INTRODUCTION

Ultra wideband (UWB) communication technology has attracted considerable attention [1]. However, to fully deliver its appealing features, it calls for low complexity transmitter and receiver (transceiver) design techniques in the presence of severe multipath distortion and multiuser interference (MUI), while not sacrificing much in detection performance and data rate.

Transmitted reference (TR) modulation is an effective means to combat unknown channel distortion by transmitting some reference signals known to the receiver [2], [3]. This technique has been applied to UWB transceiver design [4], [5], [6], [7]. The first pulse of each doublet is a reference pulse, and the second (delayed) pulse is data modulated in its amplitude or delay. A low complexity correlation receiver adopts the received signal corresponding to the reference pulse as a template to demodulate the data pulse [5], [6]. However, minimum pulse spacing larger than channel delay spread inevitably sacrifices data rate. In addition, the template may be very noisy, limiting the conventional TR performance. If small spacing between pulses is desirable in order to increase the data rate, then inter-pulse interference (IPI) further contaminates the template and may consequently yield poor detection performance.

In order to achieve near full-rate data transmission, pulse spacing may be permitted to be arbitrarily small at the price of induced severe IPI at the receiver. A mean-based smoothing technique has been developed to obtain a clean template from interference-contaminated received signals [8]. Similar to the conventional TR schemes, the method focuses on single user transmission. The method is ineffective in suppressing MUI since it simply relies on variable delays of the data pulses. Thus a multiuser TR (MTR) scheme is proposed in [9]. The data pulse is modulated either in amplitude by pulse amplitude modulation (PAM), or in delay by pulse position modulation (PPM). Meanwhile, a pair of pseudo-random (PN) sequences is assigned to the reference and data pulses to modulate their amplitudes. In this scheme, the data rate advantage of the single-user counterpart is still preserved while being able to support multiple access. At the receiver, during template acquisition, interference from all users' data signals and other

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users' reference signals is mitigated, while during data acquisition, interference from all interfering users' data signals and all users' reference signals is suppressed, thanks to the property of PN codes in addition to pulse delays.

It is interesting to investigate whether the data rate can be further increased for the MTR scheme where single modulation is employed. If bi-orthogonal signaling is used to modulate pulse amplitude as well as position, the information rate will be doubled, while other appealing features are retained. This modulation technique has been applied to single user transceiver design without transmitting a reference pulse [10]. In that scheme, the receiver must know channel distortion for data detection. If a TR scheme is considered [11], then detection can be based solely on the received signal. Extension to tri-orthogonal signaling is also discussed, and the design of high-level orthogonal signaling still remains an open topic. However, the too noisy template causes unsatisfactory detection performance in typical UWB environments. Even for a single user system, only 10^{-1} bit error rate (BER) is achieved at $10dB$ signal to noise ratio (SNR).

In this work, bi-orthogonal signaling and MTR ideas are integrated into UWB transceiver design in order to increase data rate and support multiuser communication. High data rate is realized not only by the smaller pulse spacing, but also by the orthogonal data modulation. Besides using distinct delays that are set for data pulses of different users, MUI is tackled by user and pulse dependent PN sequences. Accordingly, a simple receiver structure is suggested for independent detection of the two bits based on purified templates.

HIGH-RATE MTR UWB SYSTEM

The conventional single-user TR scheme suffers from a rate penalty as well as a performance loss. The proposed single-user near full-rate TR scheme can increase the data rate by decreasing the pulse spacing and improved detection performance due to the use of a noise-reduced template. The data rate can be further increased by orthogonal signaling [10], [11]. If TR scheme is adopted, the first pulse is a reference pulse, and the second pulse is modulated by two information symbols $I_{k,n}^{(1)}$ and $I_{k,n}^{(2)}$ for user k in its amplitude and delay using PAM and PPM respectively. For simplicity, consider binary modulation only and assume $I_{k,n}^{(1)}, I_{k,n}^{(2)} \in \{\pm 1\}$. This can be easily generalized to high order modulations as in [11]. The transmitted signal of power $\mathcal{P}_k$ due to user k in a K-user system can be described by

$$s_k(t) = \sqrt{\frac{\mathcal{P}_k}{2}} \sum_n [A_{k,n} w(t - nT_f) + I_{k,\lfloor n/N_f \rfloor}^{(1)} B_{k,n} w(t - nT_f - \tau_k - b_{k,\lfloor n/N_f \rfloor}^{(2)} \sigma)], \quad (1)$$

where $w(t)$ is the monopulse, T_f is frame duration, $A_{k,n}$ and $B_{k,n}$ are two frame-rate binary PN sequences taking $\{\pm 1\}$, $\tau_k = c_k T_c$ is the nominal delay of the data pulse in multiples of T_c that is controlled by a hopping code c_k, $b_{k,\lfloor n/N_f \rfloor}^{(2)} \triangleq \frac{1}{2}|I_{k,\lfloor n/N_f \rfloor}^{(1)} - I_{k,\lfloor n/N_f \rfloor}^{(2)}|$, σ is a PPM modulation parameter. There exist four possible pairs of $(I_{k,n}^{(1)}, I_{k,n}^{(2)})$ as $\{(1,1),(1,-1),(-1,1),(-1,-1)\}$. If a template $w(t) + w(t-\sigma)$ is considered for correlation detection of $I_{k,n}^{(1)}$ and σ is chosen such that orthogonal signaling is adopted, then corresponding outputs become $\{1,1,-1,-1\}$ that correctly reflect the polarity of $I_{k,n}^{(1)}$. Therefore a simple threshold detector suffices to perform this task. Similarly, if a template $w(t) - w(t-\sigma)$ is considered for correlation detection of $I_{k,n}^{(2)}$, then four corresponding outcomes become $\{1,-1,1,-1\}$. The correlator's output can be directly fed into a threshold detector as well. Under this signaling, both bits are independently decoded. No complicated joint detection is necessary. See [10] for further discussion of this.

Not surprisingly, the model subsumes a conventional TR system if $K = 1$, PN codes take value 1 and delay τ_k becomes a constant T_d that is larger than the channel delay spread. It also represents a near full-rate single-user scheme if τ_k is sufficiently small. Incorporating only one modulation format leads to a near full-rate MTR scheme. If only PAM is considered, then $I_{k,n}^{(2)} = I_{k,n}^{(1)}$, while for PPM, $I_{k,n}^{(1)} = 1$. The model also represents a superimposed training transmission scheme when the nominal delay τ_k is set to zero. However, introducing τ_k brings more design flexibility and enhances resilience to MUI. It should be mentioned that although only this unbalanced modulation is presented, balanced modulation is possible by distributing two information bits to two pulses respectively to achieve some benefits such as in a single user case [12]. We mainly focus on unbalanced modulation due to lack of space.

After propagating through a multipath channel $\theta_k(t)$, each pulse is distorted by the transmitter antenna, channel, receiver antenna and bandpass filter $g(t)$ so that the effective channel is $h_k(t) = \sqrt{\frac{\mathcal{P}_k}{2}} w(t) \star \theta_k(t) \star g(t)$. Then the received signal takes the following form

$$r(t) = \sum_{k,n} [A_{k,n} h_k(t - nT_f)$$

$$+ \quad I^{(1)}_{k,\lfloor n/N_f \rfloor} B_{k,n} h_k(t - nT_f - \tau_k - b^{(2)}_{k,\lfloor n/N_f \rfloor}\sigma)] \\ + \quad v(t), \qquad (2)$$

where $v(t) = n(t) \star g(t)$ is the output of the filter $g(t)$ whose input is additive white Gaussian noise (AWGN) $n(t)$ with double-sided power spectral density $\frac{N_0}{2}$.

Suppose $h_k(t)$ has support in $(0, T_h)$ where $T_h \gg \tau_k$. It is clear that both the reference signal and the data signal interfere with each other at the receiver due to the small pulse spacing. Assume $T_h < T_f$, but the analysis can be generalized to other situations. The receiver must detect both bits in each symbol interval based only on the received noisy signals. If the instantaneous received signal is adopted as the template as in the conventional TR system [4], then the interference to the reference signal introduced by all users' data signals, interfering users' reference signals and noise causes the template to be very noisy. Thus a clean template needs to be retrieved, following an approach presented in [9]. Both template acquisition and data detection processes will be facilitated by exploiting the desired user's PN codes. Without loss of generality, assume user k is the user of interest. Its bits $I^{(1)}_{k,n}$ and $I^{(2)}_{k,n}$ are to be decoded. For simplicity, term the former as the PAM bit and the latter as the PPM bit.

DATA DEMODULATION

Conventional correlation detection of the PAM bit uses the template $h_k(t)$, while $h_k(t) - h_k(t-\sigma)$ is used for the PPM bit [13]. For orthogonal signaling, $h_k(t) + h_k(t-\sigma)$ has to be used for the PAM bit in order to achieve independent detection of the two bits by simple threshold detection [10], [11]. Clearly, template acquisition in each case requires reliable estimation of the waveform $h_k(t)$ from the received noisy signal. This waveform repeats from frame to frame, but is corrupted by various interferences. Rather than directly using the noisy received signal, we use a smoothing technique to obtain a cleaner template [8], [9]. Let's consider the received signal $r(t)$ in N_s symbol intervals, and partition it into segments each of frame duration T_f. Totally, there are $N_p \triangleq N_f N_s$ segments. The m'-th ($m' = 1, \cdots, N_p$) segment is denoted by $r_{m'}(t) \triangleq r(t + m'T_f)$ for $t \in [0, T_f)$, and $r_{m'}(t) \triangleq 0$ elsewhere. Similarly, define $v_{m'}(t)$ for the noise.

The waveform $h_k(t)$ characterizes the reference signal after channel propagation of the reference pulse. According to (2) and assisted by the first PN sequence $A_{k,m'}$ that takes values $\{\pm 1\}$, one can find

$$E\{A_{k,m'} r_{m'}(t)\} = h_k(t) + \sum_{l \neq k} A_{k,m'} A_{l,m'} h_l(t), \qquad (3)$$

because $E\{I^{(1)}_{l,\lfloor m'/N_f \rfloor}\} = 0$ and $E\{v_{m'}(t)\} = 0$. For notational convenience, we have dropped the limits $l = 1, \cdots, K$ (possibly l_1 and l_2 later on) for summation over user index. We will also follow similar practice for frame index $m' = 1, \cdots, N_p$ in N_s symbol intervals, frame index $m = nN_f, \cdots, (n+1)N_f - 1)$ in the n-th symbol interval, and binary PPM modulation index $i = 0, 1$ (possibly i_1 and i_2). So, in the mean, interference in (3) is attributed to all users' reference signals only. It can be further reduced after considering the property of PN codes. The time average of $A_{k,m'} A_{l,m'}$ over N_p frame intervals favorably approaches zero as N_p increases. Therefore, according to (3), an estimate of the waveform can be obtained along the lines of [8], [9] as follows

$$\widehat{h}_k(t) = \frac{1}{N_p} \sum_{m'} A_{k,m'} r_{m'}(t). \qquad (4)$$

The estimator requires delay elements, multipliers, and adders. It can be implemented in analog circuitry by delay-and-add operations or in mixed analog/digital circuitry.

Once the waveform is estimated, detection of both PAM and PPM bits can be performed. The n-th symbol interval is comprised of N_f frame intervals ($m = nN_f, \cdots, (n+1)N_f - 1$). For each frame segment, a generic signal $\widetilde{r}_{k,m}(t)$ is obtained from $r_m(t)$ by subtracting estimated reference signal, that is then shifted and multiplied by $B_{k,m}$

$$\bar{r}_{k,m}(t) = B_{k,m} \widetilde{r}_{k,m}(t + c_k T_c), \\ \widetilde{r}_{k,m}(t) = r_m(t) - A_{k,m} \widehat{h}_k(t). \qquad (5)$$

As explained earlier, two different templates are constructed to detect two bits respectively. To detect PAM bit $I^{(1)}_{k,n}$, construct a template $\widehat{\Psi}_{k,0}(t)$ according to $\widehat{\Psi}_{k,d}(t) = \widehat{h}_k(t - dT_c) + \widehat{h}_k(t - dT_c - \sigma)$ at lag dT_c. It is correlated with $\bar{r}_{k,m}(t)$ to yield an output signal whose polarity is used for estimation of $I^{(1)}_{k,n}$

$$\widehat{I}^{(1)}_{k,n} = \mathrm{sign}(y_{k,n,1}), \\ y_{k,n,1} = \frac{1}{N_f} \sum_m \int_0^{T_f} \widehat{\Psi}_{k,0}(t) \bar{r}_{k,m}(t) dt. \qquad (6)$$

To detect PPM bit $I^{(2)}_{k,n}$, we construct a template $\widehat{\Phi}_{k,0}(t)$ according to $\widehat{\Phi}_{k,d}(t) = \widehat{h}_k(t - dT_c) - \widehat{h}_k(t - dT_c - \sigma)$ at lag dT_c and correlate it with $\bar{r}_{k,m}(t)$

$$\widehat{I}_{k,n}^{(2)} = \text{sign}(y_{k,n,2}),$$

$$y_{k,n,2} = \frac{1}{N_f}\sum_m \int_0^{T_f} \widehat{\Phi}_{k,0}(t)\bar{r}_{k,m}(t)dt. \qquad (7)$$

Modulation parameter $\sigma > 0$ affects the detection performance of both bits. Regardless of channel distortion, the rule of thumb is to consider the autocorrelation of $w(t)$ [13]. For the current demodulation criterion, it is chosen such that $w(t)$ and $w(t-\sigma)$ are orthogonal (zero-correlation) to balance detection performance between two bits. An example in the simulation section shows that such a choice is relatively insensitive to channel distortions.

PERFORMANCE STUDY

It is clear that the quality of the waveform estimator affects each detector's performance, and this will be studied first. To quantify waveform estimation performance, define a waveform estimation error $\delta h_k(t) = \widehat{h}_k(t) - h_k(t)$ for user k, and the corresponding MSE as

$$MSE_k = \int_0^{T_f} E\{[\delta h_k(t)]^2\}dt. \qquad (8)$$

The BER of each detector with an imperfect template will be evaluated. For tractable analysis, binary PN sequences are approximated as random sequences with zero mean and unit variance. This assumption can yield reliable results for large sample size, as for example demonstrated in an aperiodic CDMA system [14].

Waveform Estimation MSE

MSE evaluation requires evaluation of $E\{[\delta h_k(t)]^2\}$. Define $\sigma = \alpha T_c$. Based on (2) and (4), we obtain

$$N_p \delta h_k(t) = \sum_{l\neq k, m'} A_{k,m'}A_{l,m'}h_l(t) + \sum_{m'} A_{k,m'}v_{m'}(t) + \sum_{l,m'} A_{k,m'}B_{l,m'}I^{(1)}_{l,\lfloor m'/N_f \rfloor} h_l(t - c_l T_c - b^{(2)}_{k,\lfloor m'/N_f \rfloor}\alpha T_c). \qquad (9)$$

It involves noise, reference signals of interfering users, and data signals of all users. To facilitate later derivation of the BER, let's examine a general term $E\{\delta h_k(t+a)\delta h_k(\tau+b)\}$. Then it suffices to set $a = b = 0$ and $t = \tau$ in order to evaluate the MSE given by (8). Invoking the assumptions on PN codes, inputs and noise, we obtain

$$E\{\delta h_k(t+a)\delta h_k(\tau+b)\} = \frac{1}{N_p}\sum_{l\neq k} h_l(t+a)h_l(\tau+b) + \frac{1}{N_p}E\{v(t+a)v(\tau+b)\} + \frac{1}{2N_p}\sum_{i,l} h_l(t+a-c_lT_c - i\alpha T_c) \times h_l(\tau+b-c_lT_c-i\alpha T_c). \qquad (10)$$

For the noise term, $v(t) = n(t) \star g(t)$ and $E\{n(t)n(\tau)\} = \frac{N_0}{2}\delta(t-\tau)$. If the ideal bandpass filter $g(t)$ has unit frequency response over $f \in [-\frac{\mathcal{B}}{2}, \frac{\mathcal{B}}{2}]$, then one can easily find that

$$E\{v(t)v(\tau)\} = \sigma_v^2 \phi(t-\tau),$$

where

$$\sigma_v^2 \triangleq \frac{N_0}{2}\mathcal{B}, \ \phi(t) \triangleq \text{sinc}(\pi \mathcal{B} t).$$

After defining a deterministic cross correlation of waveforms of users l_1 and l_2 at offsets d_1T_c and d_2T_c as

$$\mathcal{E}_{l_1,l_2,d_1,d_2} \triangleq \int_0^{T_f} h_{l_1}(t-d_1T_c)h_{l_2}(t-d_2T_c)dt,$$

substituting (10) in (8), and letting $a = b = 0$ and $t = \tau$, the MSE becomes

$$MSE_k = \sum_{l\neq k}\frac{\mathcal{E}_{l,l,0,0}}{N_p} + \sum_{i,l}\frac{\mathcal{E}_{l,l,c_l+i\alpha,c_l+i\alpha}}{2N_p} + \frac{\sigma_v^2 T_f}{N_p}. \quad (11)$$

It consists of autocorrelations of interfering users' waveforms at zero offset, those of all users' waveforms at offsets equal to possible delays of their data pulses, and noise contribution. It is inversely proportional to the number of frame segments N_p. If only one segment from one symbol interval is used in the estimator, then the MSE level may be unacceptable and the template too noisy. That is the case of the conventional detector. Our windowed smoothing of received signals across multiple symbol intervals significantly reduces waveform estimation error and will lead to improved detection quality. Compared to the near full rate MTR scheme [9], the MSE shows the same result as the PPM modulation case therein, also involving average autocorrelations of the data signal at two possible lags due to PPM data modulation. It is not surprising that the binary PAM bit shows no impact because of the zero mean and independent assumption on the PN codes across frames and users. However, BER performance is dependent of the bit to be detected (the PAM bit and the PPM bit) since adopted templates for those bits are different.

BER Performance

Given PN sequence $A_{k,m}$, estimated reference signals of user k are subtracted to obtain N_f generic signal segments $\tilde{r}_{k,m}(t)$ $(m = nN_f, \cdots, nN_f + N_f - 1)$ in the n-th symbol interval

$$\begin{aligned}\tilde{r}_{k,m}(t) &= I_{k,n}^{(1)} B_{k,m} h_k(t - c_k T_c - b_{k,n}^{(2)} \alpha T_c) \\ &+ \sum_{l \neq k} [A_{l,m} h_l(t) + I_{l,n}^{(1)} B_{l,m} h_l(t - c_l T_c - b_{l,n}^{(2)} \alpha T_c)] \\ &- A_{k,m} \delta h_k(t) + v_m(t),\end{aligned} \quad (12)$$

and subsequently (5) becomes

$$\bar{r}_{k,m}(t) = I_{k,n}^{(1)} h_k(t - b_{k,n}^{(2)} \alpha T_c) + u_{k,m}(t), \quad (13)$$

where $u_{k,m}(t)$ represents waveform estimation error, MUI, plus noise,

$$\begin{aligned}u_{k,m}(t) &= -A_{k,m} B_{k,m} \delta h_k(t + c_k T_c) \\ &+ B_{k,m} v_m(t + c_k T_c) + \sum_{l \neq k} [A_{l,m} B_{k,m} h_l(t + c_k T_c) \\ &+ B_{l,m} B_{k,m} I_{l,n}^{(1)} h_l(t + c_k T_c - c_l T_c - b_{l,n}^{(2)} \alpha T_c)].\end{aligned} \quad (14)$$

Detection of the two bits requires separate discussions since the templates are different.

PAM bit: Expressing the estimated waveform as $h_k(t) + \delta h_k(t)$, the template $\widehat{\Psi}_{k,0}(t)$ is given by

$$\widehat{\Psi}_{k,0}(t) = \Psi_{k,0}(t) + \delta \Psi_{k,0}(t),$$

$$\delta \Psi_{k,0}(t) = \delta h_k(t) + \delta h_k(t - \alpha T_c),$$

where $\Psi_{k,d}(t)$ $(d = 0)$ without an hat represents the perfect template for the PAM bit corresponding to the clean waveform $h_k(t)$. This convention will similarly apply to $\Phi_{k,d}(t)$ for the PPM bit. Substituting this and (13) into the signal (6) upon which the detector is based, signal and noise components can be identified as

$$z_s = I_{k,n}^{(1)} \int_0^{T_f} \Psi_{k,0}(t) h_k(t - b_{k,n}^{(2)} \alpha T_c) dt, \quad (15)$$

$$\begin{aligned}z_n &= I_{k,n}^{(1)} \int_0^{T_f} \delta \Psi_{k,0}(t) h_k(t - b_{k,n}^{(2)} \alpha T_c) dt \\ &+ \sum_m \int_0^{T_f} \left[\Psi_{k,0}(t) + \delta \Psi_{k,0}(t)\right] \frac{u_{k,m}(t)}{N_f} dt.\end{aligned} \quad (16)$$

The noise component is due to imperfect template, interference and noise. Assume z_n is a Gaussian random variable. According to the central limit theorem, this assumption is reasonable when N_p is large since $\delta h_k(t)$ given by (9) stems from the sum of many terms, and it directly contributes to both $u_{k,m}(t)$ and z_n. Then the BER of the detector depends on the signal to noise ratio. Define

$$\mathcal{F}_{l_1,l_2,d_1,d_2} \triangleq \int_0^{T_f} h_{l_1}(t - d_1 T_c) \Psi_{l_2,d_2}(t) dt. \quad (17)$$

The signal power is easily found to be

$$\epsilon_s = E\{z_s^2\} = \frac{1}{2} \sum_i \mathcal{F}_{k,k,i\alpha,0}^2. \quad (18)$$

To evaluate the power of z_n, statistics of $\delta \Psi_{k,0}(t)$ and $u_{k,m}(t)$ are required. If those N_p segments used for waveform estimation exclude N_f segments in the current (n-th) symbol interval, clearly all terms in the expression of $u_{k,m}(t)$ in (14), except the first, are independent of $\delta \Psi_{k,0}(t)$. Even if those N_f segments are used, it is still plausible to assume that they are independent of $\delta \Psi_{k,0}(t)$, since the waveform may be typically estimated based on $N_p \gg N_f$ segments. Under this assumption, we obtain the power $\epsilon_n = E\{z_n^2\}$ as

$$\begin{aligned}\epsilon_n &= \frac{1}{2} \sum_i \iint_0^{T_f} E\{\delta \Psi_{k,0}(t) \delta \Psi_{k,0}(\tau)\} \\ &\qquad \times h_k(t - i\alpha T_c) h_k(\tau - i\alpha T_c) dt\, d\tau \\ &+ \frac{1}{N_f} \iint_0^{T_f} \Big[\Psi_{k,0}(t) \Psi_{k,0}(\tau) + E\{\delta \Psi_{k,0}(t) \delta \Psi_{k,0}(\tau)\}\Big] \\ &\qquad \times E\{u_{k,m}(t) u_{k,m}(\tau)\} dt\, d\tau.\end{aligned} \quad (19)$$

Although statistics of $\delta h_k(t)$ have been derived, simplification of ϵ_n requires statistics of $\delta \Psi_{k,0}(t)$ and $u_{k,m}(t)$. The former $E\{\delta \Psi_{k,0}(t) \delta \Psi_{k,0}(\tau)\}$ can be derived from (9), and the latter $E\{u_{k,m}(t) u_{k,m}(\tau)\}$ from (14) and applying (10). After straightforward manipulations, (19) is simplified as

$$\begin{aligned}\epsilon_n &= \left(\frac{\sigma_v^2}{N_f} + \frac{\sigma_v^2}{N_f N_p}\right) \mathcal{Q}_{k,0} + \frac{\sigma_v^4}{N_f N_p} \mathcal{X} + \sum_i \frac{\sigma_v^2}{2N_p} \mathcal{R}_{k,i\alpha} \\ &+ \sum_{l \neq k} \Big[\left(1 + \frac{1}{N_p}\right) \frac{\mathcal{F}_{l,k,-c_k,0}^2}{N_f} + \frac{\sigma_v^2 \mathcal{R}_{l,-c_k}}{N_f N_p} + \frac{\sigma_v^2 \mathcal{Q}_{l,0}}{N_f N_p}\Big] \\ &+ \sum_{i,l} \left(\frac{\mathcal{F}_{l,k,c_l - c_k + i\alpha,0}^2}{2N_f N_p} + \frac{\sigma_v^2}{2N_f N_p} \mathcal{Q}_{l,c_l + i\alpha}\right) \\ &+ \sum_{i,l \neq k} \left(\frac{\sigma_v^2 \mathcal{R}_{l,c_l - c_k + i\alpha}}{2N_f N_p} + \frac{\mathcal{F}_{l,k,c_l - c_k + i\alpha,0}^2}{2N_f} + \frac{\mathcal{F}_{k,l,i\alpha,0}^2}{2N_p}\right) \\ &+ \sum_{i_1,i_2,l} \frac{\mathcal{F}_{k,l,i_1\alpha,c_l + i_2\alpha}^2}{4N_p} + \sum_{l_1 \neq k, l_2 \neq k} \frac{\mathcal{F}_{l_1,l_2,-c_k,0}^2}{N_f N_p} \\ &+ \sum_{i,l_1 \neq k, l_2 \neq k} \frac{\mathcal{F}_{l_1,l_2,c_{l_1} - c_k + i\alpha,0}^2}{2N_f N_p} + \sum_{i,l_2,l_1 \neq k} \frac{\mathcal{F}_{l_1,l_2,-c_k,c_{l_2} + i\alpha}^2}{2N_f N_p} \\ &+ \sum_{i_1,i_2,l_2,l_1 \neq k} \frac{\mathcal{F}_{l_1,l_2,c_{l_1} - c_k + i_1\alpha,c_{l_2} + i_2\alpha}^2}{4N_f N_p},\end{aligned} \quad (20)$$

where all terms of order $\frac{1}{N_p^2}$ have been ignored, $\mathcal{Q}_{k,d}$, $\mathcal{X}$ and $\mathcal{R}_{k,d}$ are defined as

$$\mathcal{Q}_{k,d} \triangleq \iint_0^{T_f} \phi(t-\tau)\Psi_{k,d}(t)\Psi_{k,d}(\tau)dt\,d\tau, \qquad (21)$$

$$\mathcal{X} \triangleq \iint_0^{T_f} \Big[2\phi(t-\tau) + \phi(t-\tau+\alpha T_c) + \phi(t-\tau-\alpha T_c)\Big]\phi(t-\tau)dt\,d\tau, \qquad (22)$$

$$\mathcal{R}_{k,d} \triangleq \iint_0^{T_f} \Big[2\phi(t-\tau) + \phi(t-\tau+\alpha T_c) + \phi(t-\tau-\alpha T_c)\Big]h_k(t-dT_c)h_k(\tau-dT_c)dt\,d\tau. (23)$$

The power ϵ_n depends on autocorrelations as well as cross correlations of all users' waveforms at different offsets controlled by hopping codes. Introducing different hopping codes helps to reduce interference. It is expected that if asynchrony is incorporated in the received data model to characterize different time of arrivals, then ϵ_n will be further decreased. Most terms in (20) are inversely proportional to N_p (or equivalently sample size N_s for given N_f), so increasing N_p will decrease ϵ_n as well. However, the following lower bound exists

$$\epsilon_n = \frac{\sigma_v^2 \mathcal{Q}_{k,0}}{N_f} + \sum_{l\neq k} \frac{\mathcal{F}^2_{l,k,-c_k,0}}{N_f} + \sum_{i,l\neq k} \frac{\mathcal{F}^2_{l,k,c_l-c_k+i\alpha,0}}{2N_f},$$

which corresponds to the limiting case $N_p \to \infty$. This means that even in the absence of waveform estimation error ($MSE_k \to 0$ as $N_p \to \infty$ according to (11)), interference from other users' reference as well as data pulses plus noise in those N_f segments of the n-th symbol interval are non-trivial. In this case, ϵ_n is inversely proportional to N_f. So, increasing N_f is desirable while meeting the data rate requirement. The BER of our detector can be evaluated as $Q(\sqrt{\frac{\epsilon_s}{\epsilon_n}})$ where $Q(x)$ is the Q-function given by $Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-\frac{x^2}{2}} dx$.

PPM bit: Detection of PPM bit is different from that of PAM bit because a different template is used. For this case, the derivation follows along similar lines, using Φ to replace Ψ in (17), (21), and change two positive signs in each of (22) and (23) before ϕ by negative signs. Then ϵ_s and ϵ_n have the same forms as before.

NUMERICAL EXAMPLES

Performance of the proposed detectors is tested and corresponding analytical results are verified by computer simulation. The second derivative of Gaussian pulse with $D_g = 0.7ns$ and $\tau_m = 0.2877ns$ is adopted as the transmitted pulse [13]. Its autocorrelation is plotted in Figure 1 from which σ is chosen as $0.27ns$ for orthogonal signaling. Except when stated otherwise, the following typical parameters are set: $N_s = 500$ symbols, $N_f = 2$ frames per symbol, $T_c = 1ns$, $K = 4$ users, $E_b/N_0 = 10dB$. Each TH code c_k is chosen randomly from a set $\{D, \cdots, D_{max}\}$ in each of 100 independent channel realizations where channels are generated according to the IEEE UWB CM1 channel model [15], $D = 3$, $D_{max} = D + K$. T_f is set to be slightly larger than the maximum channel delay spread, typically in the range $[20, 40]ns$ for this channel model. With a relatively small D, severe IPI at the receiver occurs. Effects of varying sample size N_s from 1 to 3000 are shown in Figure 2. For the proposed transceiver, experimental result converges to both the analytical one and the clairvoyant bound (where the true noise-free waveform is used in the detector) as the sample size increases to about 1000. At $N_s = 1$, it degrades to the conventional receiver. A small difference is due to the subtraction of the reference signal in the proposed method. At $N_s = 100$, the BER is close to 10^{-2}. The slight detection advantage of the PAM bit stems from the fact that the delay spread channel destroys the orthogonality of the PPM waveform. This can be observed in Figure 3 that shows the effects of σ. At $\sigma = 0.27ns$, the results are consistent with convergence levels in the previous figure. If σ is chosen as $0.156ns$ for smallest autocorrelation of $w(t)$ and $w(t-\sigma)$ as in [13], then the detector for the PPM bit is much better than that for the PAM bit. For $\sigma > 0.4ns$, detection performance is almost balanced. Since the larger autocorrelation favors detection of the PAM bit but the smaller one favors detection of the PPM bit, this figure is consistent with Figure 1. Noise effects are shown in Figure 4. At $10dB$, BER of 10^{-2} is still achievable while the conventional receiver fails. [2]

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[2]The views and conclusions contained in this document are those of the authors and should not be interpreted as representing the official policies, either expressed or implied, of the Army Research Laboratory or the U.S. Government.

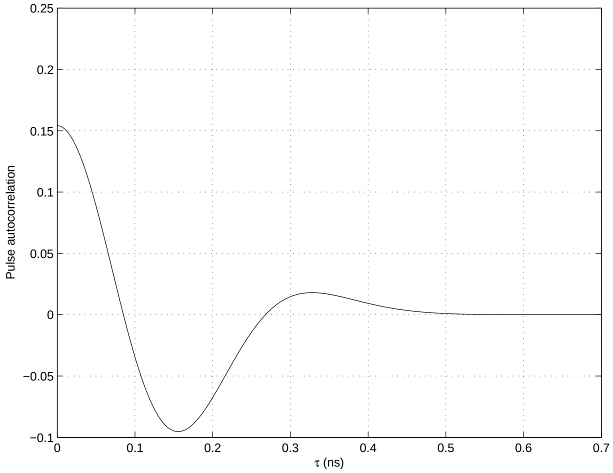

Fig. 1. Pulse autocorrelation.

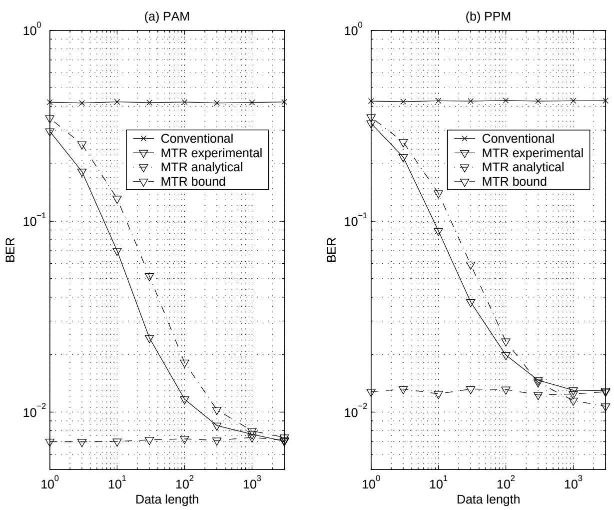

Fig. 2. BER versus data length.

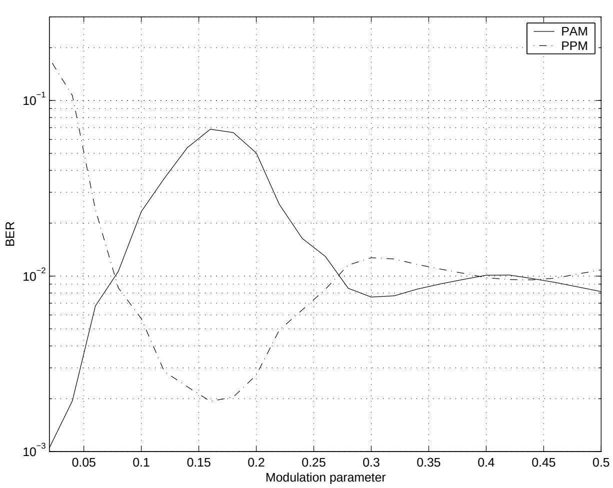

Fig. 3. Effects of modulation parameter σ on BERs for two bits.

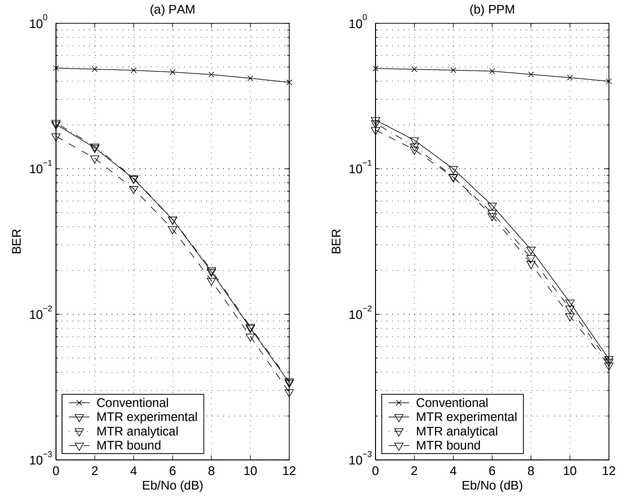

Fig. 4. BER versus Eb/No.

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