CODE AIDED NEAR FULL RATE MULTIUSER TR-UWB SYSTEMS

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ABSTRACT

A mean estimator has been proposed to improve template for the correlation receiver employed in a single user transmitted reference (TR) ultra wideband (UWB) system. The receiver yields satisfactory detection performance even in the presence of inter-pulse interference (IPI) while transmission rate is increased to near full rate compared to a 50% rate loss in a conventional TR system. This paper further extends the TR modulation to a multiuser scenario by assigning a pair of frame rate pseudo-random (PN) sequences to each user, modulating the amplitude of each pulse in the doublet respectively, irrespective of adopted data modulation formats. Given a user's PN sequence that modulates the reference pulse, the mean estimator is still applicable to estimate channel-distorted waveform for that user in the presence of IPI and multiple access interference. Then correlation based data demodulation is assisted by the other PN sequence modulating the data pulse, and constructed template from estimated waveform. Waveform estimation error and bit error rate detection performance are studied analytically and experimentally.

1. INTRODUCTION

Transmitted reference (TR) modulation, proposed a few decades ago for a narrowband system [1]-[3], is an effective means to combat multipath distortion. It is recently applied to ultra wideband (UWB) communications [4]-[6]. The second of the two well separated consecutive pulses conveys information in either amplitude (pulse amplitude modulation - PAM), phase (binary phase shift keying - BPSK), or position (pulse position modulation - PPM). Spacing between two pulses is designed large enough to avoid inter-pulse interference (IPI) at the receiver. Thus data rate is usually upper bounded to at most 50%. In order to improve template corrupted by noise, [5] and [7] propose to average signals within one symbol interval to minimize noise effect. Consequently, better detection performance is achieved than a conventional receiver built upon a noisy template.

The noise effect can be further alleviated by statistically averaging signals over multiple symbol intervals [8]. Although the reference pulse is distorted by channel yielding a channel-distorted waveform, that waveform can be estimated by utilizing the first order statistic of received signals. Then estimated waveform can be either directly used as a template to detect a PAM symbol or be used to construct a template to detect a PPM symbol [5]. Utilization of the mean of received signals requires less intensive computations. We allow negligibly small time between the data pulse and the reference pulse, introducing IPI at the receiver. However, IPI can be suppressed in both waveform estimation and data detection processes. During estimation, IPI stems from the data signal and can be maximally eliminated by statistical averaging, yielding a more purified waveform. During detection, IPI can be subtracted after the waveform template is estimated. The waveform estimator shows decreasing mean square error (MSE) with increased size of observation windows, significantly smaller than those from other existing methods and conventional TR scheme. Consequently, detection performance improves significantly. It has also been observed that PAM is slightly better than PPM in terms of detector's bit error rate (BER) performance.

Since method in [8] is designed for a single user TR-UWB system and relies on digital processing, this paper further extends that scheme to a multiuser scenario. In order to enable multiple access and increase capacity, proper coding is introduced, similar to [9] motivated by an overlaying idea for a code division multiple access (CDMA) system [10], [11]. Since both reference and data pulses need to be discriminated across users, two frame-rate pseudo-random (PN) sequences are assigned to each user. The first one modulates the amplitude of the reference pulse, while the second one the data pulse. Then using a user' first PN sequence, the waveform of that user can be estimated by a mean estimator because contributions of interference from both interfering users reference signals and all users data signals can be minimized, attributed to the PN property of each PN sequence. In the case of PAM signaling, zero mean of the PAM symbol enhances minimization of the interference from all data signals. That estimated waveform is either directly used as a template to detect the PAM symbol or used to construct a template to detect the PPM symbol by a correlation receiver. The waveform estimator employs delay elements, adders and multipliers, and meanwhile the correlation receiver performs addition, integration and symbol rate dumping operations. Thus the receiver can be implemented in mixed analog/digital circuits. In order to further alleviate multiple access interference (MAI), a hopping code is adopted to adjust the delay of the data pulse of each user. Performance of waveform estimation and detection is analyzed and studied numerically as well.

2. NEAR FULL RATE MULTIUSER TR-UWB SYSTEMS

A conventional TR-UWB system considers single user transmission and detection [4]. A user transmits a doublet in each frame of T_f seconds. The first pulse serves as a reference and is information free. The second pulse is modulated by either a PAM symbol or PPM symbol and delayed by T_d seconds. Denote the pulse by $w(t)$ with duration T_w. Each symbol repeats N_f frames, so symbol period is $T_s = N_f T_f$. In order to accommodate multiuser communication and easily obtain templates for each of two data modulation cases, we propose to modulate the amplitude of the reference pulse by a unique PN sequence at the frame rate. Meanwhile, the other PN sequence is applied to modulate the amplitude of the data pulse and also protect each user's data. Those PN

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sequences help to reduce the waveform estimation MSE and suppress MAI as illustrated later. To enhance mitigation of MAI, the delay of the second pulse is controlled by another user dependent time hopping (TH) sequence. For easy illustration of proposed methods later, we only focus on binary PAM and PPM signaling although it is straightforward to generalize discussions to high order modulation.

2.1. PAM signaling

Denote the binary PAM symbol by $I_{k,n} \in \{\pm 1\}$. The transmitted signal with power $\mathcal{P}_k$ from user k in a K-user system is proposed in the following form

$$\begin{aligned} s_k(t) &= \sqrt{\frac{\mathcal{P}_k}{2}} \sum_n \Big[A_{k,n} w(t - nT_f) \\ &+ I_{k,\lfloor n/N_f \rfloor} B_{k,n} w(t - nT_f - \tau_{k,n}) \Big], \end{aligned} \tag{1}$$

where $A_{k,n}$ and $B_{k,n}$ are frame-rate binary PN sequences taking values ± 1. They can also be chosen randomly from a ternary set $\{+1, 0, -1\}$ with pre-specified probabilities, providing more flexibility to MAI rejection and multipath mitigation [12]. Notation $\lfloor \cdot \rfloor$ is an integer floor operator. Delay $\tau_{k,n} = c_{k,n} T_c$ of the second pulse is designed to minimize MAI as well where $c_{k,n} \in \{D, D+1, \cdots, D_{max}\}$ is the hopping code, T_c is the chip duration, $T_f = N_c T_c$. The minimum spacing of two pulses is $T_d \overset{\Delta}{=} D T_c$. It can be arbitrarily small to achieve near full rate data transmission. It thus eliminates a 50% rate penalty, similar to single-user modeling without PN coding [8]. The model subsumes a superimposed training transmission scheme corresponding to $\tau_{k,n} = 0$, as well as a conventional TR system if $K = 1$, $A_{k,n}$ and $B_{k,n}$ take values 1, and delay $\tau_{k,n}$ is set as T_d. For a single-user system, $B_{k,n}$ can still be introduced to reduce waveform estimation error illustrated later. For small D, signals resulting from reference and data pulses after multipath propagation may severely interfere with each other, causing IPI. If we denote a multipath channel impulse response by $\theta_k(t)$, and transmitter-receiver front end bandpass filter by $g(t)$, the received signal becomes

$$\begin{aligned} r(t) &= \sum_k \sum_n \Big[A_{k,n} h_k(t - nT_f) \\ &+ I_{k,\lfloor n/N_f \rfloor} B_{k,n} h_k(t - nT_f - c_{k,n} T_c) \Big] + v(t), \end{aligned} \tag{2}$$

where $h_k(t) = \sqrt{\frac{\mathcal{P}_k}{2}} w(t) \star \theta_k(t) \star g(t)$ is the unknown waveform, $\star$ denotes convolution, $v(t) = n(t) \star g(t)$ and $n(t)$ represents zero mean Gaussian noise with two-sided power spectral density $\frac{N_0}{2}$. Propagation delay for each user is ignored for simplicity, but is analytically unnecessary. Indeed, it creates the worst communication scenario when other users maximally interfere with the desired user. Generally, MAI may be reduced if the users signals are mis-aligned. This simple reasoning suggests the worst-case detection performance (BER upper bound) based on this model. Suppose all $h_k(t)$ have support in $(0, T_h)$. Since both reference and data pulses propagate through the same channel, $h_k(t)$ is not only the received signal due to the reference pulse, but also the waveform of the data symbol after delay $\tau_{k,n}$. Though technically unnecessary, assume $T_h + \tau_{k,n} < T_f$ for simplified analysis of the methods proposed later. Even so however, severe IPI results. Hence, if a noisy template directly taken from received signal is used for a correlation receiver as in a conventional TR system, it leads to a large data demodulation error. Therefore, a mean-based estimation technique will be proposed to clean the "dirty" template based on an observation window spanning multiple symbol intervals.

2.2. PPM signaling

In this system, the second pulse conveys information by the pulse position. Similarly, after propagating through a multipath channel, received signal has the following form

$$\begin{aligned} r(t) &= \sum_k \sum_n \Big[A_{k,n} h_k(t - nT_f) \\ &+ B_{k,n} h_k(t - nT_f - c_{k,n} T_c - \tau_{I_{k,\lfloor n/N_f \rfloor}}) \Big] + v(t), \end{aligned} \tag{3}$$

where $h_k(t)$ is the waveform defined before, $\tau_{I_{k,n}} = I_{k,n} \sigma_d = I_{k,n} \alpha T_c$ is the modulation delay controlled by a binary information sequence $I_{k,n}$ that takes $\{0, 1\}$, σ_d is a modulation parameter that can be properly designed [13].

Our goal is to decode information sequence $I_{k,n}$ in the unknown channel for either PAM or PPM modulation based on model (2) or (3) respectively. Waveform $h_k(t)$ will be estimated from received signal $r(t)$ first, and then used for symbol detection.

3. TEMPLATE ACQUISITION AND SYMBOL DETECTION

A conventional TR correlation receiver uses the instantaneously received signal in the first segment in each frame as a template to detect the PAM symbol. Such a template is very noisy even in a single user system. In [7], averaging of $r(t)$ over N_f frames within one symbol period is performed to reduce noise. In a multiuser system, the situation deteriorates because of interference from reference signals of other users, all users data signals, and background noise. Thus a clean template is required. Exploiting the PN sequences and zero mean property of the noise, statistical averaging of segments of $r(t)$ (normalized by $A_{k,n}$) from different frames across multiple symbol intervals significantly reduces interference (even for the non-zero mean PPM symbol case).

Because of repetitive transmission of a reference pulse, each user's waveform $h_k(t)$ repeats from frame to frame. It is reasonable to partition the received signal $r(t)$ into segments of duration T_f, in order to estimate the waveform. Let's consider user k and estimate $h_k(t)$. Take $r(t)$ in N_s symbol intervals, yielding a total of $N_p \overset{\Delta}{=} N_f N_s$ segments. The m'-th ($m' = 1, \cdots, N_p$) segment is defined as $r_{m'}(t) \overset{\Delta}{=} r(t + m' T_f)$ for $t \in [0, T_f)$, and $r_{m'}(t) \overset{\Delta}{=} 0$ elsewhere. Similarly, define $v_{m'}(t)$ for the noise. According to (2) or (3), and assisted by the first PN sequence of this user that takes values ± 1, we find the following expected value

$$\begin{aligned} E\{A_{k,m'} r_{m'}(t)\} &= h_k(t) + \sum_{l \neq k, l=1}^{K} A_{k,m'} A_{l,m'} h_l(t) \\ &+ \kappa \sum_{i=0}^{1} \sum_{l=1}^{K} \frac{1}{2} A_{k,m'} B_{l,m'} h_l(t - c_{l,m'} T_c - i \alpha T_c), \end{aligned} \tag{4}$$

where $\kappa = 0$ for PAM signaling, and $\kappa = 1$ for PPM signaling, expected value of the PPM modulated data pulse has been explicitly evaluated with equally probable values in $\{0, 1\}$, and zero mean of PAM symbols and noise has been used. So, in the mean, interference is attributed to reference signals only with PAM signaling, or both reference and data signals with PPM signaling. It can be further reduced after considering the PN property. The time average of each of $A_{k,m'}A_{l,m'}$ and $A_{k,m'}B_{l,m'}$ over N_p frame intervals favorably approaches zero as N_p increases. Therefore, an estimate of the waveform for a multiuser system can be described along the lines of a single-user waveform estimator in [8] as follows

$$\widehat{h}_k(t) = \frac{1}{N_p}\sum_{m'=1}^{N_p} A_{k,m'} r_{m'}(t). \tag{5}$$

The estimator requires delay elements, multipliers, and adders.

Detection of either PAM or PPM symbol continues straightforwardly employing the estimated waveform. Consider detection of $I_{k,n}$, the n-th symbol of user k. There are N_f segments $r_m(t)$ for $m = nN_f, \cdots, (n+1)N_f - 1$. Reference signal can be subtracted after waveform is estimated. If we assume all $A_{l,m}$ ($l = 1, \cdots, K$) are known to the receiver such as in the uplink, then $A_{l,m}h_l(t)$ can be subtracted from $r_m(t)$ after waveforms $h_l(t)$ are estimated. In a case when only $A_{k,m}$ is known, such as in the downlink, only the desired user's reference signal is subtracted. Denote the generic signal after subtraction by $\widetilde{r}_{k,m}(t)$. Then the PAM symbol can be estimated based on outputs of N_f correlators in the n-th symbol interval via

$$\widehat{I}_{k,n} = sign\Big(\frac{1}{N_f}\sum_{m=nN_f}^{(n+1)N_f-1}\int_0^{T_f}\widehat{h}_k(t)\bar{r}_{k,m}(t)dt\Big), \tag{6}$$

$$\bar{r}_{k,m}(t) \triangleq B_{k,m}\widetilde{r}_{k,m}(t + c_{k,m}T_c). \tag{7}$$

To simplify notations, we will omit the upper and lower limits in each summation later, but follow the same convention for time indices m' and m, given by m' from 1 to N_p and m from nN_f to $(n+1)N_f - 1$. Others include user index l (possibly additional ones as l_1, l_2) from 1 to K, and PPM modulation index i (possibly additional ones as i_1, i_2) from 0 to 1. Similarly, for the PPM modulated input, construct a template from the estimated waveform, and also perform a simple mapping from $\{\pm 1\}$ to $\{0, 1\}$ as [13]

$$\widehat{I}_{k,n} = \frac{1}{2}(1 - y_{k,n}), \tag{8}$$

where $y_{k,n}$ is the detector's output taking $\{\pm 1\}$

$$y_{k,n} = sign\Big(\frac{1}{N_f}\sum_m\int_0^{T_f}[\widehat{h}_k(t) - \widehat{h}_k(t - \alpha T_c)]\bar{r}_{k,m}(t)dt\Big). \tag{9}$$

In the next section we jointly analyze PAM and PPM detector performance with waveform estimate given by (5).

4. PERFORMANCE STUDY

Given N_p received signal segments, our waveform estimator depends on statistics of received signals. Consequently, detectors' performance based on constructed template is also dependent on corresponding statistics. To quantify estimation performance, define the waveform estimation error as $\delta h_k(t) = \widehat{h}_k(t) - h_k(t)$ and corresponding MSE as

$$\xi_k = \int_0^{T_f} E\big\{[\delta h_k(t)]^2\big\}dt. \tag{10}$$

Then BER of each detector with imperfect template will be evaluated. For tractable analysis, approximate PN sequences as random binary sequences with zero mean and unit variance. This assumption can yield fairly reliable results for large sample size, as already demonstrated in an aperiodic CDMA system [14]. Also for concise analytical results, each TH sequence is assumed periodic, allowing dropping the time dependent index of $c_{k,m}$ as c_k.

4.1. PAM signaling

Using (2) to obtain $r_m(t)$ and subsequently substituting in (5), we find

$$\begin{aligned}\delta h_k(t)N_p \quad &= \quad \sum_{l,m',l\neq k} A_{k,m'}A_{l,m'}h_l(t) + \sum_{m'} A_{k,m'}v_{m'}(t)\\ &+ \quad \sum_{l,m'} A_{k,m'}B_{l,m'}I_{l,\lfloor m'/N_f\rfloor}h_l(t - c_lT_c).\end{aligned} \tag{11}$$

Invoking assumptions on PN codes, inputs and noise, we obtain

$$\begin{aligned}&E\{\delta h_k(t)\delta h_k(\tau)\} = \frac{1}{N_p}\sum_{l,l\neq k} h_l(t)h_l(\tau)\\ &+\frac{\sigma_v^2}{N_p}\phi(t-\tau) + \frac{1}{N_p}\sum_l h_l(t - c_lT_c)h_l(\tau - c_lT_c)\end{aligned} \tag{12}$$

where $\sigma_v^2 \triangleq \frac{N_0}{2}\mathcal{B}$, $\phi(t) \triangleq \mathrm{sinc}(\pi\mathcal{B}t)$, $\frac{\mathcal{B}}{2}$ is the bandwidth of the filter $g(t)$. Define a deterministic cross correlation of waveforms of users l_1 and l_2 ($l_1, l_2 = 1, \cdots, K$) at offsets d_1T_c and d_2T_c as

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

Applying (12), the MSE (10) becomes

$$\xi_k = \frac{1}{N_p}\sum_{l,l\neq k}\mathcal{E}_{l,l,0,0} + \frac{1}{N_p}\sum_l \mathcal{E}_{l,l,c_l,c_l} + \frac{\sigma_v^2T_f}{N_p}. \tag{13}$$

Two cases will be discussed for detection of input next: downlink and uplink.

4.1.1. UWB downlink

After subtraction of estimated reference signal, those N_f generic signals $\widetilde{r}_{k,m}(t)$ can be obtained. Signal $\bar{r}_{k,m}(t)$ given by (7) and used in (6) becomes

$$\bar{r}_{k,m}(t) = I_{k,n}h_k(t) + u_{k,m}(t) \tag{14}$$

where $u_{k,m}(t)$ represents interference plus noise

$$\begin{aligned}u_{k,m}(t) \quad &= \quad -A_{k,m}B_{k,m}\delta h_k(t + c_kT_c) + v_m(t + c_kT_c)B_{k,m}\\ &+ \quad \sum_{l,l\neq k}[A_{l,m}B_{k,m}h_l(t + c_kT_c)\\ &+ \quad B_{l,m}B_{k,m}I_{l,n}h_l(t + c_kT_c - c_lT_c)].\end{aligned} \tag{15}$$

Expressing estimated waveform by $h_k(t) + \delta h_k(t)$, the signal component in (6) is identified as $I_{k,n}\mathcal{E}_{k,k,0,0}$ whose power is $\epsilon_s = \mathcal{E}_{k,k,0,0}^2$, and the interference plus noise component as

$$\begin{aligned}z_n \quad &= \quad I_{k,n}\int_0^T \delta h_k(t)h_k(t)dt + \frac{1}{N_f}\sum_m\int_0^T h_k(t)u_{k,m}(t)dt\\ &+\frac{1}{N_f}\sum_m\int_0^T \delta h_k(t)u_{k,m}(t)dt.\end{aligned} \tag{16}$$

The power $\epsilon_n = E\{z_n^2\}$ depends on statistics of both $\delta h_k(t)$ and $u_{k,m}(t)$. Reasonably approximate $\delta h_k(t)$ as independent of all terms in $u_{k,m}(t)$ except the first term, $E\{u_{k,m}(t)u_{k,m}(\tau)\}$ can be simplified based on (15) and (12). Applying (12) once more, ϵ_n can be derived. It turns out that ϵ_n involves cross-correlations of different waveforms at different offsets and is a function of N_f, N_p, time hopping codes as

$$\begin{aligned}\epsilon_n &= (\frac{\sigma_v^2}{N_f} + \frac{\sigma_v^2}{N_p} + \frac{\sigma_v^2}{N_f N_p})\mathcal{H}_{k,0} + \frac{\sigma_v^4}{N_f N_p}\mathcal{Y} \\ &+ \sum_l (\frac{\mathcal{E}^2_{k,l,0,c_l}}{N_p} + \frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f N_p} + \frac{\sigma_v^2 \mathcal{H}_{l,c_l}}{N_f N_p}) \\ &+ \sum_{l,l\neq k} \big[(\frac{1}{N_f} + \frac{1}{N_f N_p})\mathcal{E}^2_{k,l,0,-c_k} + \frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f} \\ &+ \frac{\mathcal{E}^2_{k,l,0,0}}{N_p} + \frac{\sigma_v^2 \mathcal{H}_{l,0}}{N_f N_p} + \frac{\sigma_v^2 \mathcal{H}_{l,-c_k}}{N_f N_p} + \frac{\sigma_v^2 \mathcal{H}_{l,c_l-c_k}}{N_f N_p}\big] \\ &+ \sum_{l_1,l_2,l_1\neq k,l_2\neq k} (\frac{\mathcal{E}^2_{l_1,l_2,0,-c_k}}{N_f N_p} + \frac{\mathcal{E}^2_{l_1,l_2,0,c_{l_2}-c_k}}{N_f N_p}) \\ &+ \sum_{l_1,l_2,l_2\neq k} (\frac{\mathcal{E}^2_{l_1,l_2,c_{l_1},-c_k}}{N_f N_p} + \frac{\mathcal{E}^2_{l_1,l_2,c_{l_1},c_{l_2}-c_k}}{N_f N_p}), \quad (17)\end{aligned}$$

where

$$\mathcal{H}_{k,d} \triangleq \iint_0^{T_f} \phi(t-\tau)h_k(t-dT_c)h_k(\tau-dT_c)dt\,d\tau,$$

$$\mathcal{Y} \triangleq \iint_0^{T_f} [\phi(t-\tau)]^2 dt\,d\tau.$$

Due to lack of space, we omit its derivation and also its expression for other cases in subsequent discussions. The result is based on our synchronism assumption, thus provides the worst-case performance for an asynchronous system since MAI may be significantly reduced from pulsed asynchronous transmissions. If $N_p \gg 1$, then the waveform estimation error plays little role. It is then simplified to

$$\epsilon_n = \frac{\sigma_v^2 \mathcal{H}_{k,0}}{N_f} + \sum_{l,l\neq k} (\frac{\mathcal{E}^2_{k,l,0,-c_k}}{N_f} + \frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f}), \quad (18)$$

mainly contributed by interference plus noise in the current symbol interval. The BER of the detector can be evaluated as $Q(\sqrt{\frac{\epsilon_s}{\epsilon_n}})$ where $Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-\frac{x^2}{2}} dx$.

4.1.2. UWB uplink

In this case, each user's waveform is estimated and estimated reference signal is subtracted. The signal component in $\bar{r}_{k,m}(t)$ is still $I_{k,n}h_k(t)$, while $u_{k,m}(t)$ becomes

$$\begin{aligned}u_{k,m}(t) &= -\sum_l A_{l,m}B_{k,m}\delta h_l(t+c_kT_c) + v_m(t+c_kT_c)B_{k,m} \\ &+ \sum_{l,l\neq k} B_{l,m}B_{k,m}I_{l,n}h_l(t+c_kT_c-c_lT_c). \quad (19)\end{aligned}$$

The power of z_n can be derived according to (16), (19) and (12), similarly as before. No cross-correlation of $\delta h_{l_1}(t)$ and $\delta h_{l_2}(t)$ for $l_1 \neq l_2$ is required since $A_{l_1,m}$ and $A_{l_2,m}$ are assumed independent. In a case of $N_p \gg 1$, it has a simple form

$$\epsilon_n = \frac{\sigma_v^2 \mathcal{H}_{k,0}}{N_f} + \sum_{l,l\neq k} \frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f}. \quad (20)$$

Compared with (18), this power is reasonably smaller since reference signals from interfering users are subtracted.

4.2. PPM signaling

Considering (3) and (5), $\delta h_k(t)$ satisfies

$$\begin{aligned}\delta h_k(t)N_p &= \sum_{l,m',l\neq k} A_{l,m'}A_{k,m'}h_l(t) + \sum_{m'} v_{m'}(t)A_{k,m'} \\ &+ \sum_{l,m'} B_{l,m'}A_{k,m'}h_l(t-c_lT_c-\tau_{I_{l,\lfloor m'/N_f\rfloor}}). \quad (21)\end{aligned}$$

Then invoking assumptions on PN codes, inputs and noise and considering PPM modulation, we obtain

$$\begin{aligned}E\{\delta h_k(t)\delta h_k(\tau)\} &= \frac{1}{N_p}\sum_{l,l\neq k} h_l(t)h_l(\tau) + \frac{\sigma_v^2}{N_p}\phi(t-\tau) \\ &+ \frac{1}{2N_p^2}\sum_{i,l,m'} h_l(t-c_lT_c-i\alpha T_c)h_l(\tau-c_lT_c-i\alpha T_c). \quad (22)\end{aligned}$$

Applying (22) to evaluate (10), the MSE becomes

$$\xi_k = \sum_{l,l\neq k} \frac{\mathcal{E}_{l,l,0,0}}{N_p} + \sum_{i,l} \frac{\mathcal{E}_{l,l,c_l+i\alpha,c_l+i\alpha}}{2N_p} + \frac{\sigma_v^2 T_f}{N_p}. \quad (23)$$

Similarly, two cases are discussed for detection next.

4.2.1. UWB downlink

In this case, signal $\bar{r}_{k,m}(t)$ for detection is given by

$$\bar{r}_{k,m}(t) = h_k(t - I_{k,n}\alpha T_c) + u_{k,m}(t) \quad (24)$$

where $u_{k,m}(t)$ represents interference plus noise

$$\begin{aligned}u_{k,m}(t) &= -A_{k,m}B_{k,m}\delta h_k(t+c_kT_c) + v_m(t+c_kT_c)B_{k,m} \\ &+ \sum_{l,l\neq k} [A_{l,m}B_{k,m}h_l(t+c_kT_c) \\ &+ B_{l,m}B_{k,m}h_l(t+c_kT_c-c_lT_c-I_{l,n}\alpha T_c)]. \quad (25)\end{aligned}$$

Detection is based on $y_{k,m}$ in (9). Then interference plus noise power again requires statistics $E\{u_{k,m}(t)u_{k,m}(\tau)\}$. Based on (25) and applying (22), it can be derived. In a case of $N_p \gg 1$, it has the following form

$$\epsilon_n = \frac{\sigma_v^2 \mathcal{Q}_{k,0}}{N_f} + \sum_{l,l\neq k} \frac{\mathcal{F}^2_{l,k,-c_k,0}}{N_f} + \sum_{i,l,l\neq k} \frac{\mathcal{F}^2_{l,k,c_l-c_k+i\alpha,0}}{2N_f}, \quad (26)$$

where

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

$$\Psi_{k,d}(t) \triangleq h_k(t-dT_c) - h_k(t-dT_c-\alpha T_c).$$

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

The BER of this detector can be evaluated from the signal to noise ratio (SNR) as before.

4.2.2. UWB uplink

Subtraction of K estimated reference signals yields $u_{k,m}(t)$ as

$$\begin{aligned} u_{k,m}(t) &= -\sum_{l} A_{l,m}B_{k,m}\delta h_l(t+c_kT_c) + v_m(t+c_kT_c)B_{k,m} \\ &+ \sum_{l,l\neq k} B_{l,m}B_{k,m}h_l(t+c_kT_c-c_lT_c-I_{k,n}\alpha T_c). \quad (27) \end{aligned}$$

The signal power is same as before, and the noise power ϵ_n can be re-derived. If $N_p \gg 1$, it becomes

$$\epsilon_n = \frac{\sigma_v^2 \mathcal{Q}_{k,0}}{N_f} + \sum_{i,l,l\neq k} \frac{\mathcal{F}^2_{l,k,c_l-c_k+i\alpha,0}}{2N_f}. \quad (28)$$

Clearly, it is smaller than (26).

5. NUMERICAL STUDY

We adopted the second derivative of Gaussian pulse of duration $D_g = 0.7ns$ as the transmitted monocycle [13]. Binary PN sequences are generated randomly. $N_f = 2$, $T_c = 1ns$, $K = 4$. Each user's TH code is chosen randomly from a set $\{3, 4, 5, 6\}$ in each of 100 channel realizations according to the IEEE UWB CM1 channel model [15]. T_f is set to be slightly larger than the maximum channel delay spread on the order of tens of nanoseconds because of long tails, yielding severe IPI at the receiver. For PPM signaling, modulation delay is set $\sigma_d = 0.156ns$ [13]. Figure 1 shows effects of N_s on BER of the conventional correlators and proposed uplink and downlink detectors for both PAM and PPM based systems with 10dB bit energy to noise ratio. "Bound" curves are based on true channels, for comparison with analytical and experimental results. Our method improves significantly with N_s because of improved template. Convergence to each bound at large N_s is observed. PAM modulation is better than PPM. Noisy template (for conventional detector) fails to recover inputs. Figure 2 presents effects of E_b/N_o with waveforms estimated from $N_s = 500$ received signals. Convergence of experimental results to analytical ones is observed. Also, BERs of the proposed detectors approach their bounds.

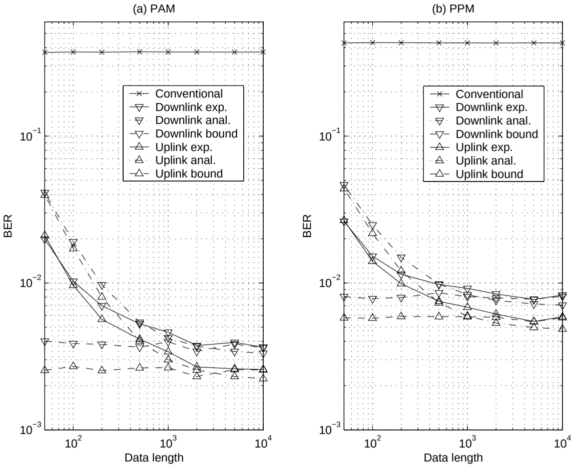

Fig. 1. BER versus data length.

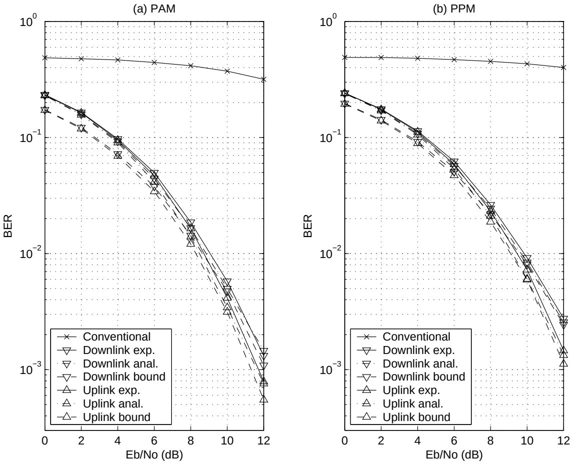

Fig. 2. BER versus SNR.

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