Multiuser Transmitted Reference Ultra-Wideband Communication Systems

Zhengyuan Xu, *Senior Member, IEEE,* and Brian M. Sadler, *Senior Member, IEEE*

***Abstract*—A conventional transmitted reference (TR) modulation scheme is an effective means to combat severe multipath distortion in an ultra-wideband (UWB) system, significantly relaxing the equalization requirements. However, it suffers from multiple-access interference and a data rate loss. In this paper, we propose a multiuser TR scheme to extend TR modulation to the multiuser case, while almost doubling the data rate by allowing arbitrarily small spacing of pulse pairs. The proposed scheme assigns a pair of pseudorandom sequences to each user to enable multiple access, modulating the amplitude of data and reference pulses, respectively. A waveform template is first estimated, followed by data demodulation. A time-hopping code can be employed to further minimize the effect of signal collisions. The method is appropriate for both pulse amplitude modulation and pulse position modulation. Waveform estimation and bit-error rate analysis are provided, and confirmed with simulations. Substantial detection improvements over conventional TR detectors are observed.**

***Index Terms*—Interpulse interference, multiple access, multiuser transmitted reference (MTR), ultra-wideband (UWB).**

I. Introduction

THE APPROVAL of ultra-wideband (UWB) transmission in the United States [1] and elsewhere has sparked significant research interest [2]–[4]. Potential UWB applications include not only short-range data and multimedia communications, but also sensing, localization, and tracking, as well as vehicle collision avoidance and other radar-like scenarios.

UWB offers unique features such as high resolvability of multiple paths, fine timing resolution [5], and coexistence via overlay with existing wireless systems [6]. However, UWB communication systems must somehow accommodate the significant channel distortion, and a full accounting requires very high receiver complexity. Generally, UWB receivers sacrifice performance for lowered complexity [7], e.g., the RAKE [8] and autocorrelation receivers [9]. A practical RAKE receiver consists of multiple correlators. It must select a moderate yet limited number of strong paths to combine from dozens to hundreds of possible paths. Despite medium complexity, the captured energy may be relatively low. The RAKE also suffers from channel (time of arrival and attenuation) mismatch, although high rate sampling helps to estimate channel coefficients in the design of linear receivers [10], [11].

Transmitted reference (TR) modulation is effective in mitigating multipath distortion [12]–[14]. It was proposed for narrowband systems a few decades ago [15]. The first pulse of each doublet is information-free, and the second (delayed) pulse carries the user's information via binary phase-shift keying (BPSK), pulse amplitude modulation (PAM), or pulse position modulation (PPM). The delay of the data pulse is ideally designed to be larger than the channel spread such that no interpulse interference (IPI) occurs after multipath propagation. The received waveform resulting from the reference pulse can then serve as a template to demodulate the latter data pulse using a low-complexity correlation receiver [13], [16]. However, minimum spacing of the two pulses inevitably sacrifices data rate, especially when the channel delay spread is large [17]. As the channel is used only half the time for data at best, there is a 50% rate penalty. In addition, the template may be very noisy, limiting the conventional TR performance.

In order to improve template estimation with PAM modulation, [13] and [16] propose to average signals from multiple frames within one symbol interval to minimize the noise effect; see also [18] where a channel estimation method in the presence of IPI has been proposed, based on a well-formulated discrete-time system model. Consequently, better detection performance is achieved than a conventional receiver built upon an instantaneous estimate of the template. For either PAM or PPM modulation-based systems, the noise effect can be further alleviated by statistically averaging signals over multiple symbol intervals [19]. The signal waveform estimator utilizes the first-order statistic of the received signal. As in [18], no requirement on large pulse spacing is imposed, resulting in increased data rate. At high signal-to-noise ratio (SNR), the waveform estimation mean-square error (MSE) decreases in proportion to the observation window size. Consequently, detection based on the improved template shows significant performance gain. This gain is slightly better for PAM than for PPM, in terms of both MSE and bit-error rate (BER) [19].

In this paper, we build on the above ideas, expanding to a multiuser scenario. The proposed multiuser TR (MTR) scheme incorporates psuedo-random spreading codes, similar to that in [20], [21] borrowed from an overlaying code division multiple-access (CDMA) system [22]. Since both reference and data pulses need to be differentiated across users, two pseudorandom (PN) sequences are assigned to each user at the frame rate. They modulate the amplitude of the pulse pair respectively. A

Manuscript received March 1, 2005; revised October 24, 2005. This work was supported in part by the Army Research Laboratory CTA on Communications and Networks under Grant DAAD19-01-2-0011. This paper was presented in part at the IEEE International Workshop on Signal Processing Advances for Wireless Communications, New York, June 2005. The U.S. Government is authorized to reproduce and distribute reprints for Government purposes notwithstanding any copyright notation thereon.

Z. Xu is with the Department of Electrical Engineering, University of California, Riverside, CA 92521 USA (e-mail: dxu@ee.ucr.edu).

B. M. Sadler is with the Army Research Laboratory, AMSRD-ARL-CI-CN, Adelphi, MD 20783 USA (e-mail: bsadler@arl.army.mil).

Digital Object Identifier 10.1109/JSAC.2005.863826

mean-based estimation algorithm is proposed to obtain an improved signal waveform template, applicable for both PAM and PPM-based UWB systems. Arbitrarily small pulse separation leads to increased data rate, a similar goal that can be achieved by a differential UWB system [23]. Assisted by the waveform template for the desired user, and the PN sequence modulating its data pulse, the interfering contribution from its reference pulse is subtracted before demodulation. Reference signals from all users can be suppressed if their PN sequences are known and signal waveforms estimated, such as in communications from different users/nodes to an access point. For brevity, this scenario is termed a uplink, and the opposite scenario a downlink. The proposed scheme is able to mitigate both IPI and multiple-access interference (MAI), mainly due to the PN sequence properties. In order to enhance the interference mitigation capability, a time hopping sequence may be applied to the data pulse of each user to further avoid pulse collision. Waveform estimation MSE and BER detection performance are developed analytically, and studied via simulation using the IEEE UWB channel model [17]. Substantial detection improvements over conventional detectors are observed.

The paper proceeds as follows. First, in Section II, the proposed MTR transmission schemes and data models are developed for both PAM and PPM. The corresponding waveform estimation and detection methods are described in Section III. Due to length constraints, performance is analyzed in detail in Section IV only for downlink with PAM. Numerical examples are shown in Section V, and key contributions of the work are then summarized.

II. MTR-UWB Systems

Motivated by multiple access, we propose an MTR UWB scheme to associate a unique covering PN sequence with the reference pulse at the frame rate, and a second distinct PN sequence is employed to randomize the data modulated pulse. Together, these two PN sequences uniquely specify a user. They help to reduce waveform estimation error, MAI, and increase the system capacity. The data pulse delay can be controlled by another user-dependent time hopping sequence to further minimize MAI. Denote the pulse by $w(t)$ with duration T_w. We assume each symbol repeats N_f frames, and each frame has duration T_f, so the symbol period is $T_s = N_f T_f$. For easy illustration of the proposed transmission, estimation, and detection schemes, binary PAM or PPM modulation is assumed. It is straightforward to generalize the models and methods to higher order modulations.

For PAM signaling, denote the nth PAM symbol of user k in a K-user system by $I_{k,n} \in \{\pm 1\}$. Its transmitted signal with power $\mathcal{P}_k$ can be described by

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

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 prespecified probabilities, providing more flexibility to MAI rejection and multipath mitigation [24]. Notation $\lfloor \cdot \rfloor$ is an integer floor operator. Delay $\tau_k = c_k T_c$ of the data pulse is adjusted by a hopping code $c_k \in \{D, D+1, \ldots, D_{\max}\}$, where T_c is the chip duration, $T_f = N_c T_c$. For simplicity, a constant τ_k is assumed here, although it may vary from frame to frame [25]. The minimum spacing $T_d \triangleq DT_c$ of two pulses can be arbitrarily small, eliminating a 50% rate penalty, similar to a single-user case [19]. 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 the front-end bandpass filter by $g(t)$, the received signal becomes

$$r(t) = \sum_{k=1}^{K} \sum_{n=-\infty}^{\infty} \Big[A_{k,n} h_k(t - nT_f) + I_{k,\lfloor n/N_f \rfloor} B_{k,n} h_k(t - nT_f - c_k T_c) \Big] + v(t) \quad (1)$$

where $h_k(t) = \sqrt{\mathcal{P}_k/2}\, w(t) \star \theta_k(t) \star g(t)$ 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 $N_0/2$. Suppose all $h_k(t)$ have support restricted in $(0, T_h)$. Since each pulse pair propagates 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 τ_k. Though technically and analytically unnecessary, propagation delay for each user has been ignored and $T_h + \tau_k < T_f$ is assumed for simplicity. Our approach can be easily generalized to other situations. If the reference signal is directly used as a template for a correlation receiver as in a conventional TR system, it leads to a large data demodulation error due to interference contamination. Therefore, a mean-based estimation technique will be proposed to clean the "dirty" template based on an observation window spanning multiple symbol intervals.

Clearly, our model subsumes a superimposed training transmission scheme if $\tau_k = 0$. It reduces to a conventional TR system if $K = 1$, $A_{k,n}$ and $B_{k,n}$ take values 1, and τ_k is set as T_d. Together with subsequent development, it can be adapted to the case of preamble training [26].

Similarly for PPM signaling, the received signal has the following form:

$$r(t) = \sum_{k=1}^{K} \sum_{n=-\infty}^{\infty} \Big[A_{k,n} h_k(t - nT_f) + B_{k,n} h_k(t - nT_f - c_k T_c - \tau_{I_{k,\lfloor n/N_f \rfloor}}) \Big] + v(t) \quad (2)$$

where $h_k(t)$ is defined as before, $\tau_{I_{k,n}} = I_{k,n}\sigma_d$ is the modulation delay controlled by an information sequence $I_{k,n} \in \{0, 1\}$ and modulation parameter σ_d. We assume $\sigma_d = \alpha T_c$, where α can be designed to optimize the detection performance [7].

Our goal is to detect information sequence $I_{k,n}$ in the unknown channel according to (1) or (2). First, the signal waveform $h_k(t)$ will be estimated from received signal $r(t)$. This will serve as a template to detect the PAM symbol (or used to construct a template to detect the PPM symbol) via a correlation detector [7], [13].

III. Template Acquisition and Symbol Detection

In a multiuser system, $h_k(t)$ is corrupted by reference signals of other users, all users' data pulses, and background noise. Therefore, a clean template needs to be obtained. Exploiting the PN sequences and zero-mean property of the noise, statistical averaging of segments of $r(t)$ (normalized by $A_{k,n}$) can significantly reduce interference, even for the nonzero-mean PPM symbol case. Because of repetitive transmission of a reference pulse, each user's waveform $h_k(t)$ repeats from frame to frame. Assume perfect frame synchronization (sensitivity to synchronization will be considered later). Take $r(t)$ in N_s symbol intervals. Partition it into segments, each of frame duration T_f, in order to estimate $h_k(t)$. There are a total of $N_p \triangleq N_f N_s$ segments. The m'th ($m' = 1, \ldots, N_p$) segment is defined as $r_{m'}(t) \triangleq r(t + m'T_f)$ for $t \in [0, T_f)$, and $r_{m'}(t) \triangleq 0$ elsewhere. Similarly, define $v_{m'}(t)$ for the noise. For PAM signaling, according to (1) and assisted by the first PN sequence of this user that takes values ± 1, one can find that the following expected value contains the desired waveform:

$$E\{A_{k,m'} r_{m'}(t)\} = h_k(t) + \sum_{l \neq k, l=1}^{K} A_{k,m'} A_{l,m'} h_l(t) \quad (3)$$

because $E\{I_{l,\lfloor m'/N_f \rfloor}\} = 0$ and $E\{v_{m'}(t)\} = 0$. So, in the mean, interference is attributed to reference signals only. It can be further reduced after considering the PN property, as discussed below. For PPM signaling, using (2) and noting modulation delay $\tau_{I_{k,\lfloor m'/N_f \rfloor}} = I_{k,\lfloor m'/N_f \rfloor} \alpha T_c$, where $I_{k,\lfloor m'/N_f \rfloor}$ takes equally probable values in $\{0, 1\}$, we similarly obtain

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

where $i = 0, 1, l = 1, \ldots, K$ in the first summation. If higher order PPM modulation is employed, then we can adapt the upper limit in the summation for i (and change probability 1/2) in the above equation. Now, interference stems not only from reference signals, but also data signals due to nonzero mean of all inputs, and these depend on the PN sequences. 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, according to (3) and (4), an estimate of the waveform for a multiuser system (either PAM or PPM signaling) can be described along the lines of a single-user waveform estimator in [19] as follows:

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

Detection of a PAM symbol is performed by employing the estimated waveform. Consider detection of the nth symbol $I_{k,n}$ of user k. Correspondingly, there are N_f segments $r_m(t)$ for $m = nN_f, \ldots, (n+1)N_f - 1$. If we assume all $A_{l,m}$ are known to the receiver such as in the uplink, then the contribution of reference signals $A_{l,m} h_l(t)$ from all users can be subtracted from $r_m(t)$ after their signal waveforms are estimated. In a case when only the desired user's PN sequence $A_{k,m}$ is known, such as in the downlink, only the desired user's reference signal is subtracted. Denote the generic signal after subtraction by $\widetilde{r}_{k,m}(t)$. It contains the signal part $B_{k,m} I_{k,n} h_k(t - c_k T_c)$, and interference plus noise part. Then, assisted by time hopping code c_k and the other user specific PN sequence $B_{k,m}$ modulating its data pulse, we can obtain the following signal that carries its data:

$$\bar{r}_{k,m}(t) \triangleq B_{k,m} \widetilde{r}_{k,m}(t + c_k T_c). \quad (6)$$

Afterwards, PAM modulated input $I_{k,n}$ can be estimated based on outputs of N_f correlators in the nth symbol interval via

$$\widehat{I}_{k,n} = \text{sign}\left(\frac{1}{N_f} \sum_{m=nN_f}^{(n+1)N_f - 1} \int_0^{T_f} \widehat{h}_k(t) \bar{r}_{k,m}(t) dt \right). \quad (7)$$

Various summations will be involved, and upper and lower limits for corresponding indices need to be clearly stated. For notational convenience, we will hereafter omit limits 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 modulation index i (possibly additional ones as i_1, i_2) from 0 to 1.

For detection of the PPM modulated input, a template from the estimated waveform is constructed as $\widehat{h}_k(t) - \widehat{h}_k(t - \sigma_d)$, which replaces $\widehat{h}_k(t)$ in (7), and also a simple mapping from $\{\pm 1\}$ to $\{0, 1\}$ is performed [7].

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

IV. Waveform Estimator and Detector Performance

Given N_p received signal segments, our waveform estimator depends on the received signal statistics. Subsequently, the detector performance also depends on these statistics. To quantify the 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

$$\text{MSE}_k = \int_0^{T_f} E\{[\delta h_k(t)]^2\} dt \quad (8)$$

where T_f is the maximum channel delay spread. If it is beyond T_f, then the upper limit of the integral needs to be increased together with duration of each segment $r_{m'}(t)$ used for waveform estimation. The BER of each detector with imperfect template can be evaluated. For tractable analysis, we approximate binary PN sequences 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 [27]. Due to space limitation, we only detail analysis for downlink detector with PAM signaling. Other cases can be similarly studied, with corresponding analytical results provided in [25].

Evaluation of (8) requires $E\{[\delta h_k(t)]^2\}$. After substituting the received data model (1) into the estimator (5), waveform estimation error $\delta h_k(t)$ can be expressed as

$$\delta h_k(t) = \frac{1}{N_p}\sum_{l,m',l\neq k} A_{k,m'}A_{l,m'}h_l(t) + \frac{1}{N_p}\sum_{l,m'} A_{k,m'}B_{l,m'}I_{l,\lfloor m'/N_f\rfloor}h_l(t-c_lT_c) + \frac{1}{N_p}\sum_{m'} A_{k,m'}v_{m'}(t). \quad (9)$$

It consists of noise, reference signals, and data signals. For easy later derivation of BER, consider a general term $R_k \triangleq E\{\delta h_k(t+a)\delta h_k(\tau+b)\}$, which encompasses the special case required in (8) by setting $a=b=0$ and $t=\tau$.

A noise statistic is first derived. For ideal bandpass filter $g(t)$ with unit frequency response over $f\in[-\mathcal{B}/2,\mathcal{B}/2]$, then

$$g(t) = \frac{\sin(\pi\mathcal{B}t)}{\pi t} = \mathcal{B}\,\mathrm{sinc}(\pi\mathcal{B}t).$$

Noting $v(t)=n(t)\star g(t)$ and $E\{n(t)n(\tau)\}=(N_0/2)\delta(t-\tau)$, the autocorrelation of $v(t)$ and $v(\tau)$ can be found to be

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

where $\sigma_v^2 \triangleq (N_0/2)\mathcal{B}$, $\phi(t)\triangleq \mathrm{sinc}(\pi\mathcal{B}t)$. According to (9) and invoking assumptions on PN codes, inputs and noise, we obtain statistics of the waveform estimation error

$$R_k = \frac{1}{N_p}\sum_{l,l\neq k} h_l(t+a)h_l(\tau+b) + \frac{\sigma_v^2}{N_p}\phi(t+a-\tau-b) + \frac{1}{N_p}\sum_l h_l(t+a-c_lT_c)h_l(\tau+b-c_lT_c). \quad (10)$$

To evaluate MSE, define a deterministic cross correlation of PAM templates of users l_1 and l_2 at offsets d_1T_c and d_2T_c as

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

For convenience in subsequent discussions, also define

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

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

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

$$\mathrm{MSE}_k = \sum_{l,l\neq k}\frac{\mathcal{E}_{l,l,0,0}}{N_p} + \sum_l \frac{\mathcal{E}_{l,l,c_l,c_l}}{N_p} + \frac{\sigma_v^2 T_f}{N_p}. \quad (11)$$

The first summation has autocorrelations of interfering users waveforms without offset, the second summation contains autocorrelations of all users' waveforms at offsets equal to delays of their data pulses (determined by hopping codes), and the last term results from noise. The MSE is inversely proportional to the sample size (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. If N_f segments from one symbol interval ($N_s=1$) are used, then MSE decreases [13], [16]. Our windowed smoothing of received signals across multiple symbol intervals significantly reduces waveform estimation error and improves detection quality. The degree of improvement depends on the window size.

Based upon the above result, analysis of the PAM-based detector can be continued. For downlink, the desired user's waveform is estimated first, based on its PN sequence $A_{k,m'}$. Then, its reference signals are subtracted to obtain N_f segments $\tilde{r}_{k,m}(t)$ in the nth symbol interval ($m=nN_f,\ldots,nN_f+N_f-1$). Subsequently, (6) becomes

$$\bar{r}_{k,m}(t) = I_{k,n}h_k(t) + u_{k,m}(t) \quad (12)$$

where $u_{k,m}(t)$ consists of waveform estimation error, MAI, and noise

$$u_{k,m}(t) = B_{k,m}[-A_{k,m}\delta h_k(t+c_kT_c) + v_m(t+c_kT_c)] + \sum_{l,l\neq k}[A_{l,m}B_{k,m}h_l(t+c_kT_c) + B_{l,m}B_{k,m}I_{l,n}h_l(t+c_kT_c-c_lT_c)]. \quad (13)$$

Expressing the estimated template as $h_k(t)+\delta h_k(t)$, and substituting (12) into the detector (7), signal and noise components can be identified as

$$z_s = I_{k,n}\int_0^{T_f}[h_k(t)+\delta h_k(t)]h_k(t)dt$$

$$z_n = \frac{1}{N_f}\sum_m\int_0^{T_f}[h_k(t)+\delta h_k(t)]u_{k,m}(t)dt.$$

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. Applying (10), the signal power is easily found to be

$$\epsilon_s = \mathcal{E}_{k,k,0,0}^2 + \frac{\sigma_v^2}{N_p}\mathcal{H}_{k,0} + \sum_l \frac{\mathcal{E}_{k,l,0,c_l}^2}{N_p} + \sum_{l,l\neq k}\frac{\mathcal{E}_{k,l,0,0}^2}{N_p}. \quad (14)$$

To evaluate the power of z_n, statistics of $\delta h_k(t)$ and $u_{k,m}(t)$ are required. If those N_p segments used for waveform estimation exclude N_f segments in the current (nth) symbol interval, clearly, all terms in the expression of $u_{k,m}(t)$ in (13), except the first, are independent of $\delta h_k(t)$. Even if those N_f segments are used, it is still plausible to assume that they are independent of $\delta h_k(t)$ for simplified expressions, 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

$$\begin{aligned}\epsilon_n =& \frac{1}{N_f}\iint_0^{T_f} E\{\delta h_k(t)\delta h_k(\tau)\}E\{u_{k,m}(t)u_{k,m}(\tau)\}dt\,d\tau \\ &+ \frac{1}{N_f}\iint_0^{T_f} h_k(t)h_k(\tau)E\{u_{k,m}(t)u_{k,m}(\tau)\}dt\,d\tau.\end{aligned} \tag{15}$$

Noting (13) and applying (10), statistics of $u_{k,m}(t)$ can be easily simplified, and subsequently, (15) becomes

$$\begin{aligned}\epsilon_n \approx& \left(1+\frac{1}{N_p}\right)\frac{\sigma_v^2}{N_f}\mathcal{H}_{k,0} + \frac{\sigma_v^4}{N_f N_p}\mathcal{Y} \\ &+ \sum_l \left(\frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f N_p} + \frac{\sigma_v^2 \mathcal{H}_{l,c_l}}{N_f N_p}\right) \\ &+ \sum_{l,l\neq k}\Bigg[\left(1+\frac{1}{N_p}\right)\frac{\mathcal{E}^2_{k,l,0,-c_k}}{N_f} + \frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f} \\ &+ \frac{\sigma_v^2\mathcal{H}_{l,0}}{N_f N_p} + \frac{\sigma_v^2\mathcal{H}_{l,-c_k}}{N_f N_p} + \frac{\sigma_v^2\mathcal{H}_{l,c_l-c_k}}{N_f N_p}\Bigg] \\ &+ \sum_{l_1,l_2,l_1\neq k,l_2\neq k}\left(\frac{\mathcal{E}^2_{l_1,l_2,0,-c_k}}{N_f N_p} + \frac{\mathcal{E}^2_{l_1,l_2,0,c_{l_2}-c_k}}{N_f N_p}\right) \\ &+ \sum_{l_1}\sum_{l_2,l_2\neq k}\left(\frac{\mathcal{E}^2_{l_1,l_2,c_{l_1},-c_k}}{N_f N_p} + \frac{\mathcal{E}^2_{l_1,l_2,c_{l_1},c_{l_2}-c_k}}{N_f N_p}\right)\end{aligned} \tag{16}$$

where all terms of order $1/N_p^2$ have been ignored.

Once again, the power is observed to depend on autocorrelations, as well as cross correlations of all users waveforms at different offsets, where offsets depend on the hopping codes. Clearly, the larger difference in codes, the smaller their correlations. Thus introduction of different hopping codes helps to reduce interference power. Noise contributions are reflected by terms $\mathcal{H}_{k,d}$ and $\mathcal{Y}$. Most terms in (16) are inversely proportional to sample size N_p, so increasing N_p will decrease ϵ_n as well. However, there is a lower bound dominated by those terms dependent only on N_f, which corresponds to the limiting case $N_p \to \infty$. The result in this case is given by

$$\epsilon_n \approx \frac{\sigma_v^2}{N_f}\mathcal{H}_{k,0} + \sum_{l,l\neq k}\frac{\mathcal{E}^2_{k,l,0,-c_k}}{N_f} + \sum_{l,l\neq k}\frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f}. \tag{17}$$

Thus, even in the absence of waveform estimation error ($\mathrm{MSE}_k \to 0$ as $N_p \to \infty$ according to (11)), contributions from noise (first term), other users reference (first summation) and data pulses (second summation) in the current symbol interval are nontrivial. The spreading factor N_f plays an important role. Except for the first term, all other terms in ϵ_n are inversely proportional to N_f. So, increasing N_f is desirable to minimize interference while meeting the data rate requirement. However, the noise term will not decrease as N_f increases because σ_v^2 (or $N_0/2$) that characterizes the noise after spreading includes the effect of N_f, resulting in noise-limited power. The signal power becomes $\epsilon_s = \mathcal{E}^2_{k,k,0,0}$ based on (14). The BER of the detector depends on the signal-to-interference-plus-noise ratio. As discussed before, the interference plus noise can be well modeled as a Gaussian process. Then, the BER can be evaluated as $Q(\sqrt{\epsilon_s/\epsilon_n})$, where the Q-function is given by $Q(x) = \int_x^\infty (1/\sqrt{2\pi})e^{-t^2/2}dt$.

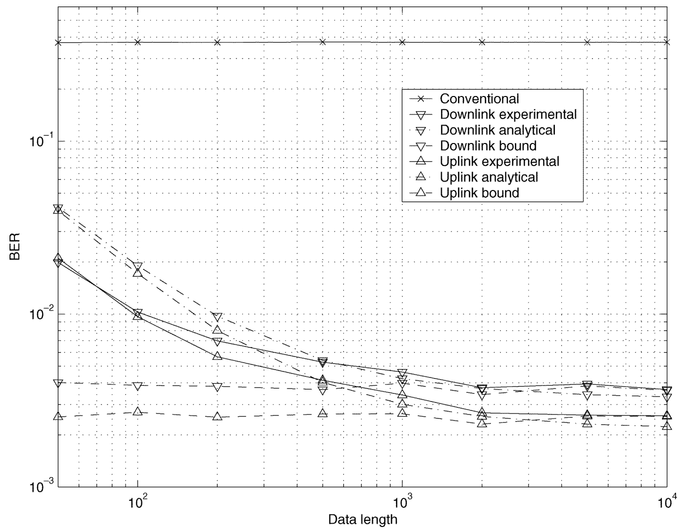

Fig. 1. Effect of data length on BER, comparing theory and simulation.

Reference signal subtraction helps to reduce interference. Without subtraction of the desired user's estimated reference signal, one more term from the desired user should be added to the first summation in (17) as

$$\epsilon_n \approx \frac{\sigma_v^2}{N_f}\mathcal{H}_{k,0} + \sum_{l}\frac{\mathcal{E}^2_{k,l,0,-c_k}}{N_f} + \sum_{l,l\neq k}\frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f}. \tag{18}$$

If all users reference signals are subtracted such as in the uplink, then the first summation in (17) disappears, yielding

$$\epsilon_n \approx \frac{\sigma_v^2}{N_f}\mathcal{H}_{k,0} + \sum_{l,l\neq k}\frac{\mathcal{E}^2_{k,l,0,c_l-c_k}}{N_f}. \tag{19}$$

The system with PPM signaling has been comprehensively treated in [25], and results are omitted due to space limitation.

V. Numerical Examples

The second derivative of Gaussian pulse $w(t) = [1 - 4\pi((t-D_g/2)/\tau_m)^2]\exp[-2\pi((t-D_g/2)/\tau_m)^2]$ is adopted as the transmitted pulse with $D_g = 0.7$ ns and $\tau_m = 0.2877$ ns [7]. Except when stated otherwise, the following typical parameters are set: $N_s = 500$, $N_f = 2$, $T_c = 1$ ns, $K = 4$, $E_b/N_0 = 10$ dB, $D = 3$, $D_{\max} = D + K$, $\sigma_d = 0.156$ ns. One hundred total channels were generated according to the IEEE UWB CM1 channel model [17], and truncated at 99% total energy capture. The bandwidth of the front-end bandpass filter is twice the higher 3 dB cutoff frequency of the monopulse. T_f is slightly larger than the maximum channel delay spread, on the order of tens of nanoseconds. Effects of N_s, E_b/N_0, K, and spreading factor N_f are studied, with focus on PAM signaling.

First, the effect of sample size is examined. Theoretically, the MSE is inversely proportional to sample size and the BER is lower bounded when sample size is sufficiently large. Simulation and MSE analytical results are observed to perfectly match. The MSE level is favorably low. BER performance is demonstrated in Fig. 1. Curves with upward triangles are for uplink, and those with downward triangles are for downlink,

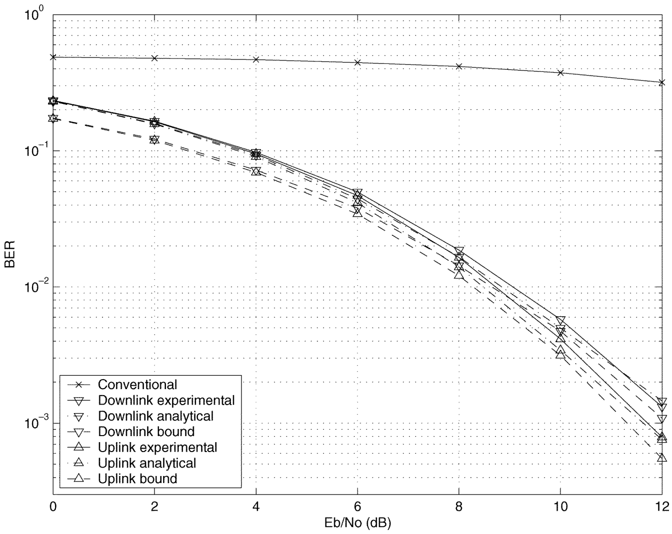

Fig. 2. BER versus SNR (E_b/N_o), comparing theory and simulation.

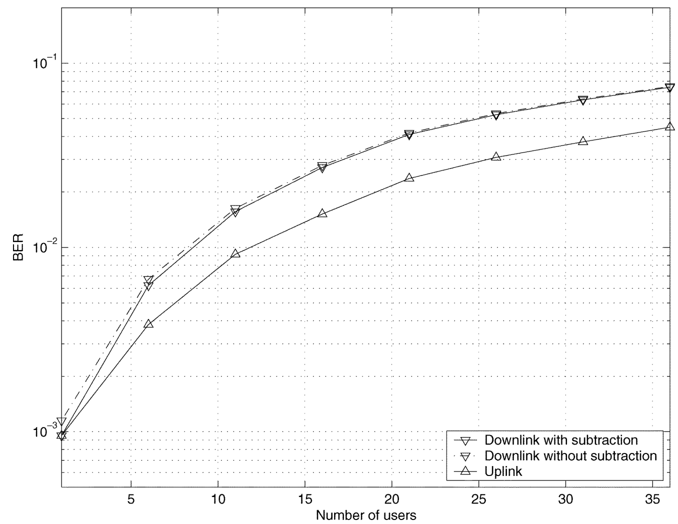

Fig. 4. BER versus number of users K, comparing uplink and downlink receivers.

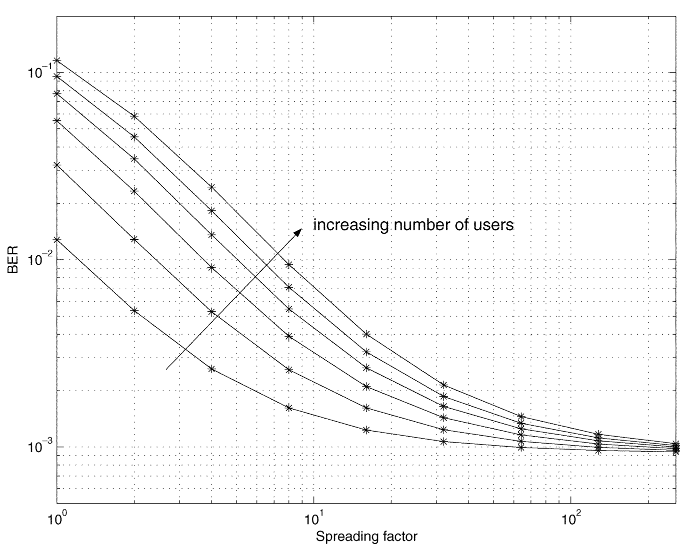

Fig. 3. BER versus spreading factor N_f. Curves are parameterized by the number of users K ($K = \{5, 10, 15, 20, 25, 30\}$ from bottom to top).

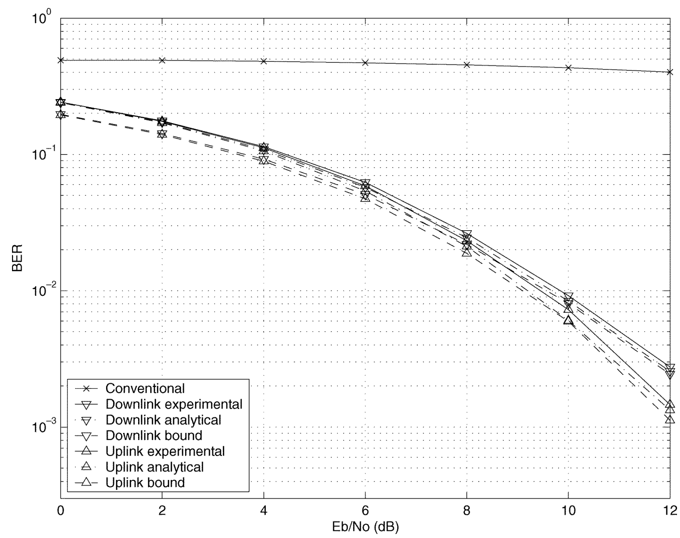

Fig. 5. BER versus SNR for a PPM system.

with stars for the conventional TR receiver. Solid lines are simulation results, while dashed lines represent bounds, where the true noise-free waveforms are used in the detector in simulation. Dashed-dotted lines are based on analysis, e.g., using (14) and (16) for downlink. Simulation results converge to both analytical ones and bounds as sample size increases to about 1000. The uplink detector is better than the downlink one for large sample size. Both detectors significantly outperform the conventional one that uses a very noisy template.

Fig. 2 shows SNR effect on the BER, comparing theory and simulation once more. The E_b/N_0 ranges from 0 to 12 dB. Reliable detection is achieved, and the proposed detectors substantially outperform the conventional receiver. The uplink detector is slightly better than the downlink detector at high SNR, with analysis and simulation agreeing closely. Small gaps from associated bounds are due to the finite sample effect on the proposed detectors.

The lower bound of interference power depends on spreading factor N_f and noise. Fig. 3 shows analytical BER versus N_f equal to 2^j for $j = 0, 1, \ldots, 8$ based on (17). The curves are parameterized by number of users K, taking values 5, 10, 15, 20, 25, 30 from the bottom to top curve. The BER lower bound of about 10^{-3} is due to 10 dB noise (the same as the single-user bound). To maintain a particular BER level, N_f needs to increase with increasing K. If $N_f = K$, the BER is about 2×10^{-3}. If the BER requirement is relaxed, to say 10^{-2}, the system can support many more users. For example, K can be 15 when N_f is as small as 4, and can be more than 30 when $N_f = 10$.

The effect of interference is demonstrated in Fig. 4. Analytical BERs of downlink detectors with and without reference signal subtraction, and uplink detector with all reference signals subtracted are shown for different K varying from 1 up to 36 with a step size of 5. They are evaluated based on (17)–(19), respectively. $D_{\max}$ is fixed at 8 to make a fair comparison. Uplink detector performs the best. When $K = 1$, downlink with subtraction reduces to uplink, providing a slight gain by subtraction of user's reference signal. As K increases, effect of downlink subtraction becomes marginal due to dominant MAI, leading to convergence of two downlink curves. Downlink detectors degrade faster with increasing K than uplink. All detectors are able to provide raw BERs lower than 5×10^{-3} with five users.

Finally, we present both analytical and simulation results for PPM signaling in Fig. 5. Compared with results for PAM signaling in Fig. 2, the BERs for PPM are slightly larger.

VI. Conclusion

Incorporating PN sequences, multiuser transmitted reference schemes are proposed for both PAM and PPM UWB systems. To obtain a satisfactory template for each correlation detector, a mean-based waveform estimation method is derived. Performance of waveform estimator and detector BER are analyzed. Simulation results demonstrate that the proposed detectors substantially outperform conventional TR detectors. PAM signaling is slightly advantageous compared to PPM. Aided by PN sequences, multiuser communication is enabled, leading to large system capacity; the proposed systems can support many users at a reasonable raw BER level, while almost doubling the data rate of a conventional TR system.

References

[1] Federal Communications Commission News Release, "Revision of Part 15 of the commission's rules regarding ultra-wideband transmission systems," ET Docket 98-153, Washington, DC, 2002.

[2] R. C. Qiu, H. Liu, and X. Shen, "Ultra-wideband for multiple access communications," *IEEE Commun. Mag.*, vol. 43, no. 2, pp. 80–87, Feb. 2005.

[3] S. Roy, J. R. Foerester, V. S. Somayazulu, and D. G. Leeper, "Ultrawideband radio design: The promise of high-speed, short-range wireless connectivity," *Proc. IEEE*, vol. 92, pp. 295–311, Feb. 2004.

[4] L. Yang and G. B. Giannakis, "Ultra-wideband communications: An idea whose time has come," *IEEE Signal Process. Mag.*, vol. 21, no. 6, pp. 26–54, Nov. 2004.

[5] R. A. Scholtz, "Multiple access with time-hopping impulse modulation," in *Proc. MILCOM*, Oct. 1993, pp. 447–450.

[6] M. Hamalainen, V. Hovinen, R. Tesi, J. H. Iinatti, and M. Latva-aho, "On the UWB system coexistence with GSM900, UMTS/WCDMA, and GPS," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 9, pp. 1712–1721, Dec. 2002.

[7] M. Win and R. Scholtz, "Ultra-wide bandwidth time-hopping spread-spectrum impulse radio for wireless multiple-access communications," *IEEE Trans. Commun.*, vol. 48, pp. 679–689, Apr. 2000.

[8] M. Z. Win and R. A. Scholtz, "On the energy capture of ultrawide bandwidth signals in dense multipath environments," *IEEE Commun. Lett.*, vol. 2, no. 9, pp. 245–247, Sep. 1998.

[9] N. Guo and R. Qiu, "Improved autocorrelation receivers based on multiple symbol differential detection for UWB communications," *IEEE Trans. Wireless Commun.*, 2005, to be published.

[10] Z. Xu, J. Tang, and P. Liu, "Multiuser channel estimation for ultra-wideband systems using up to the second order statistics," *EURASIP J. Appl. Signal Process.*, vol. 2005, no. 3, pp. 273–286, Mar. 2004.

[11] P. Liu and Z. Xu, "POR-based channel estimation for UWB communications," *IEEE Trans. Wireless Commun.*, vol. 4, no. 6, pp. 2968–2982, Nov. 2005.

[12] R. T. Hoctor and H. W. Tomlinson, "An overview of delay-hopped transmitted reference RF communications," General Electronic Tech. Rep., 2001CRD198, Class 1, 2002.

[13] J. D. Choi and W. E. Stark, "Performance of ultra-wideband communications with suboptimal receivers in multipath channels," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 9, pp. 1754–1766, Dec. 2002.

[14] S. Franz and U. Mitra, "On optimal data detection for UWB transmitted reference systems," in *Proc. IEEE GLOBECOM*, vol. 2, San Francisco, CA, Dec. 2003, pp. 744–748.

[15] C. K. Rushforth, "Transmitted-reference techniques for random or unknown channels," *IEEE Trans. Inf. Theory*, vol. 10, no. 1, pp. 39–42, Jan. 1964.

[16] Y. Chao and R. Scholtz, "Optimal and suboptimal receivers for ultra-wideband transmitted reference systems," in *Proc. IEEE GLOBECOM*, vol. 2, San Francisco, CA, Dec. 2003, pp. 759–763.

[17] (2002) Channel Modeling Subcommittee Report Final. IEEE P802.15-02/368r5-SG3a. [Online]. Available: http://grouper.ieee.org/groups/802/15

[18] A. Trindade, Q. H. Dang, and A.-J. van der Veen, "Signal processing model for a transmit-reference UWB wireless communication system," in *Proc. IEEE UWBST*, Nov. 2003, pp. 270–274.

[19] Z. Xu, B. M. Sadler, and J. Tang, "Data detection for UWB transmitted reference systems with inter-pulse interference," in *Proc. ICASSP*, vol. 3, Philadelphia, PA, Mar. 2005, pp. 601–604.

[20] J. Tang and Z. Xu, "Multidimensional orthogonal design for ultra-wideband downlink," in *Proc. ICASSP*, Quebec, Canada, May 2004, pp. 541–544.

[21] C. J. Le Martret and G. B. Giannakis, "All-digital impulse radio with multiuser detection for wireless cellular systems," *IEEE Trans. Commun.*, vol. 50, pp. 1440–1450, Sep. 2002.

[22] H. Sari, F. Vanhaverbeke, and M. Moeneclaey, "Extending the capacity of multiple access channels," *IEEE Commun. Mag.*, vol. 38, no. 1, pp. 74–82, Jan. 2000.

[23] K. Witrisal and M. Pausini, "Equivalent system model of ISI in a frame-differential IR-UWB receiver," in *Proc. IEEE GLOBECOM*, vol. 6, Nov. 2004, pp. 3505–3510.

[24] B. M. Sadler and A. Swami, "On the performance of episodic UWB and direct-sequence communication systems," *IEEE Trans. Wireless Commun.*, vol. 3, no. 6, pp. 2246–2255, Nov. 2004.

[25] Z. Xu and B. M. Sadler, "Transmitted reference ultra wideband transceivers for multiuser communication," Army Research Laboratory, Tech. Rep. ARL-TR-3708, Adelphi, MD, 2005.

[26] L. Yang and G. B. Giannakis, "Optimal pilot waveform assisted modulation for ultra-wideband communications," in *Proc. 36th Asilomar Conf.*, vol. 1, Nov. 2002, pp. 733–737.

[27] Z. Xu, "Low complexity multiuser channel estimation with aperiodic spreading codes," *IEEE Trans. Signal Process.*, vol. 49, pp. 2813–2822, Nov. 2001.

Zhengyuan Xu (S'97–M'99–SM'02) received the B.S. and M.S. degrees in electronic engineering from Tsinghua University, Beijing, China, in 1989 and 1991, respectively, and the Ph.D. degree in electrical engineering from Stevens Institute of Technology, Hoboken, NJ, in 1999.

From 1991 to 1996, he worked as an Engineer and Department Manager at the Tsinghua Unisplendour Group Corporation of Tsinghua University. Since 1999, he has been with the Department of Electrical Engineering, University of California, Riverside, as an Assistant Professor, and then an Associate Professor. His current research interests include detection and estimation theory, spread spectrum and ultra-wideband wireless technology, wireless optical communication, and ad hoc and wireless sensor networks.

Dr. Xu received the Outstanding Student Award and the Motorola Scholarship from Tsinghua University, and the Peskin Award from Stevens Institute of Technology. He also received the Academic Senate Research Award and the Regents' Faculty Award from University of California, Riverside. He has served as an Associate Editor for the IEEE Transactions on Vehicular Technology and the IEEE Communications Letters, a session chair and technical program committee member for international conferences. He is an elected member of the IEEE Technical Committee on Signal Processing for Communications.

Brian M. Sadler (S'81–M'81–SM'02) received the B.S. and M.S. degrees from the University of Maryland, College Park, and the Ph.D. degree from the University of Virginia, Charlottesville, all in electrical engineering.

He is a Senior Research Scientist at the Army Research Laboratory (ARL), Adelphi, MD. He was a Lecturer at the University of Maryland, and has been lecturing at Johns Hopkins University since 1994 on statistical signal processing and communications. His research interests include signal processing for mobile wireless and ultra-wideband systems, and sensor signal processing and networking.

Dr. Sadler is an Associate Editor for the IEEE Signal Processing Letters, was an Associate Editor for the IEEE Transactions on Signal Processing, is on the Editorial Board for the *EURASIP Journal on Wireless Communications and Networking*, and was a Guest Editor for the IEEE Journal on Selected Areas in Communications (Special Issue on Military Communications). He is a member of the IEEE Technical Committee on Signal Processing for Communications, and Co-Chaired the 2nd IEEE Workshop on Signal Processing Advances in Wireless Communications (SPAWC-99).