Original Research Article

Effective timing synchronization scheme for DHTR UWB receivers

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Abstract: Ultrawideband (UWB) wireless communication systems are ideally suited to short-distance and high-data-rate wireless communications. UWB systems have a low average transmission power, and therefore ensure a fixed synchronization error between the transmitted and received ends. In order to reduce multipathinterference, UWB systems are generally implemented using Rake receivers, for which the delay time and attenuated amount of each transmitted route must be known in advance. In the present study, the complexity of Rake receivers with a large number (e.g., >10) of"fingers" is reduced by using the delay-hopped transmitted-reference (DHTR) method, which correlates the delayed signal with the original signal,and therefore avoids the need for a template signal. The synchronization performance of the proposed DHTR receiver is analyzed both theoretically and by means of numerical simulations. An effective timing synchronization scheme, designatedas "parallel signal acquisition with shared looped delay-line" (PS-SLD), is then proposed. The simulation results show that the proposed synchronization scheme achieves a higher detection probability and a lower normalized MSE than the traditional timing dirty template (TDT) algorithm in multipath environments with a single user.

Keywords: Synchronization; Multipath; Delay-Hopped Transmitted-Reference(DHTR); Template signal

1. Introduction

The term *ultra wide band* (UWB) refers to electromagnetic waveforms, which are characterized by an instantaneous fractional energy bandwidthgreater than around 0.20 – 0.25. Note that the fractional bandwidth is defined here as B / f_c, where B is the bandwidth and is give by $B = f_h - f_l$, and fc is the center frequency and is given by $f_c = \left(f_h - f_l\right) / 2$, in which f_h and f_l are the upper and lower frequencies of the -10dB emission point, respectively[1].For example, given a signal with an energy bandwidth of 1MHz and a center frequency of 2MHz, the fractional bandwidth has a value of 0.5, and thus the signal is a UWB signal. Note that the definition of a UWB signal can be given only relative to the center frequency. For example, the 500 MHz minimum bandwidth limit sets a threshold of 2.5 GHz. Below thisthreshold,signals are UWB if their fractional bandwidth exceeds 0.20. By contrast, above this threshold, signals are UWB if their bandwidth exceeds 500 MHz.

Current UWB radios use frequencies ranging from 3.1 GHz to 10.6 GHz. This frequency range satisfies the bandwidth, cost and power consumption requirements of next-generation consumer electronic (CE) devices, including DVD players, HDTVs, MP3 players and stereos, and so forth.In time hopping (TH) UWB systems, data are transmitted in trains of pulses on the order of nanoseconds[2]. Since the pulse duration is very short, TH-UWB systems achieve a fine multipath resolution, and are therefore viable candidates for communications in dense multipath environments, such as short-range or indoor wireless communications.

One of the most critical challenges in implementing TH-UWB systems is that of clock synchronization since the information-bearing waveforms are pulse-like. Moreover, the multipath channel through which these low-power narrow pulses propagate is unknown at the receiver at the synchronization stage. As a result, peak-picking the output of a sliding correlator between the received signal and the transmitted waveform template is not only sub-optimum in the presence of dense multipaths, but also results in an unacceptably slow acquisition speed[3]. Consequently, differential detection (DD)[4]or transmitted reference (TR)[5-7] schemes are generally preferred since they do not require channel estimation.

TR receivers have the major advantage of capturing energy from all of the multipath components with a far lower degree of complexity than coherent receivers. Unfortunately, however, the reference pulse is corrupted by noise and interference. In frame-differential (FD) receivers, the data are transmitted by applying differential binary modulation at the frame-level, and thus there is no need to transmit reference pulses. As a result, FD receivers have a lower inter-frame interference (IFI) than TR receivers, and are therefore better suited for higher data rate applications[8]. Although Witrisal *et al.*[8] developed an equivalent system modelfor FD impulse-radio (IR) UWB systems, the synchronization problem was not addressed. The literature contains various synchronization schemes for TR receivers[9-12]. For example, Yang and Giannakis[12] presented a novel synchronization criterion referred to as "timing with dirty templates" (TDT).The proposed criterion exploited the fact that the cross-correlation of these "dirty templates" exhibits a unique maximum at the correct symbol time. Although TDT algorithmsare blind in the sense that they do not require timing-hopping (TH) code, they are inoperable in multi-user environments in which other users employ the same training pattern.

This paper proposes an effectivesynchronization scheme for multi-user and dense multipath environments designated as "parallel signal acquisition with shared looped delay-line" (PS-SLD). The receiver architecture of PS-SLD resembles that of the FD IR-UWB system proposed in [8], but incorporates an additional shared loop delay-line (SLD). Furthermore, the synchronization criterion of PS-SLD is similar to that of the TDT algorithms in that it also relies on determining the maximumintegration output by testing all of the candidate time offsets. However, the candidate time offsets considered by TDT algorithms belong to the interval $[0, T_s)$, whereas those of the PS-SLD algorithm belong to theinterval $[0, T_f)$. Note that this interval is referred to hereafter as the "uncertainty region" due to its parallel search mechanism. The simulation results confirm that PS-SLD achieves a higher probability of detection (PD) and a lower normalized mean square error (MSE) than traditional TDT algorithms for UWB systems with a single user.

2. Signal model

This paper considers a BPSK random time-hopping (TH) impulse-radio (IR)UWB system. In other words, each information bit is first modulated by BPSK and the data symbol, $d_i \in \{+1,-1\}$, where i is the data symbol index, is then transmitted via N_f frames. (Note that each frame conveys just one pulse waveform.) Traditionally, the polarities of the N_f pulses representing an information symbol are always the same. However, more recently, pulse-based polarity randomization schemes have been proposed, in which each modulated pulse has a random polarity code, $b_j \in \{+1,-1\}$. The use of random polarity codes not only helps to fit the spectral shape according to the Federal Communications Commission (FCC) constraint by eliminating the power spectral lines in UWB IR systems, but also provides an improved robustness toward multiple access interference (MAI).

In pulse-based polarity randomization schemes, the known random polarity sequence $b_j \in \{+1,-1\}$ is differentially modulated withthe transmitted pulses. Note that $j \in \{0,1,2,...,N_f-1\}$ is the pulse index within a symbol. By superimposing BPSK and the random polarity code, each pulse is differentially modulated by both BPSK and the random polarity sequence. Thus, the differentially modulated pulsepolarities are obtained as

$a_{i,j+1} = a_{i,j} d_i b_j$ and $a_{i+1,0} = a_{i,N_f-1} d_i b_{N_f-1}$. The transmitted signal from the transmitter k is then given by [2]

$$s_k(t) = \sum_{i=\infty}^{\infty} \sum_{j=0}^{N_f-1} a_{i,j} w(t - iT_s - jT_f - c_{k,j} T_c), \tag{1),}$$

where $w(t)$ represents the transmitted pulse waveform and is non-zero only for $t \in (0, T_p)$, i is the symbol index, j is the frame index, and T_p is the pulse duration. Furthermore, $a_{i,j}$ is the pulse polarity code, and $c_{k,j}$ is the time-hopping code of the *kth* user in the *jth* frame. Finally, T_f is the frame duration, T_s is the symbol duration, T_c is the chip duration, and N_c is the number of chips in a frame.

An important feature of UWB IR systems is their use of time-hopping code, which avoids collisions in multiple accesses. This time-hopping code is a pseudorandom code with a period N_p, i.e., $c_{j+iN_p} = c_j$ for all integers i and j. Generally $N_p \geq N_f$. In the present study, N_p is assumed to be equal to N_f.

Within the k^{th} transmitter, let $D_{k,p}$ be the time duration between the p^{th} and $(p+1)^{th}$ pulses. Thus, the absolute time interval, $D_{k,p}$, can be expressed as [8]

$$D_{k,p} = T_f + (c_{k,\text{mod}(p+1)} - c_{k,\text{mod}(p,N_f)}) \times T_c \quad \text{for } p \in [0, N_f - 1], \tag{2}$$

where $(c_{k,\text{mod}(p+1,N_f)} - c_{k,\text{mod}(p,N_f)}) \times T_c$ is the relative time interval. As a result, $s_k(t)$ can be obtained as

$$s_k(t) = \sum_{i=-\infty}^{\infty} \sum_{j=0}^{N_f-1} a_{i,j} w(t - iT_s - c_{k,0} T_c - \sum_{p=0}^{j-1} D_{k,p}), \tag{3}$$

where $\sum_{p=0}^{-1} D_{k,p} = 0$. If the channel comprises L_k multipath routes, it can be modeled as an L_k tapped-delay-line(TDL) equalizer. Moreover, the channel impulse response can be expressed as

$$h_k(t) = \sum_{l=1}^{L_k} \alpha_{k,l} \delta(t - \tau_{k,l}), \tag{4}$$

where l is the path index, τ is the delay time, $\tau_{k.l} = (l-1)T_c$, $\delta(t)$ is the delta function, and $\alpha_{k,l}$ is the amplitude of the l^{th} path for the k^{th} transmitter. Given the assumption that $Y \sim N(\mu, \sigma^2)$ is normally distributed, $\alpha_{k,l}$ becomes a log-normal distribution, which satisfiesthe following probability density function (pdf):

$$f_\alpha(\alpha) = \frac{1}{\sqrt{2\pi}\sigma\alpha} e^{-\frac{(\ln\alpha - \mu)^2}{2\sigma^2}}, \quad 0 < \alpha < \infty. \tag{5}$$

3. SLD receiver

After the transmissions of all (N_u) users are transmitted over the channel, the received signal, $r(t)$, has the form

$$\begin{aligned} r(t) &= \sum_{k=0}^{N_u-1} s_k(t) * h_k(t) + n(t) \\ &= \sum_{k=0}^{N_u-1} \sum_{i=-\infty}^{\infty} \sum_{j=0}^{N_f-1} \sum_{l=1}^{L_k} a_{i,j} \alpha_{k,l} w(t - iT_s - c_{k,0} T_c - \sum_{p=0}^{j-1} D_{k,p} - \tau_{k,l}) + n(t) \end{aligned} \tag{6}$$

where $n(t)$ is additive white Gaussian noise(AWGN). For analytical convenience, let $r(t)$ be simplified as

$$r(t) = \sum_{k=0}^{N_u-1} \sum_{i=\infty}^{\infty} \sum_{j=0}^{N_f-1} a_{i,j} v_{k,i,0}(t - \sum_{p=0}^{j-1} D_{k,p}) + n(t). \tag{7}$$

Let $v_{k,i,j}(t)$ be the pulse corresponding to the *ith* symbol of the *kth* user in the *jth* frame. Having passed through the multipath channel, the pulse is gathered in the form of a pulse train at the receiver end. Figure 1 presents adiagram of the UWB SLD receiver proposed in the present study. As shown, the receiver consists of N_f parallel branches, with each branch containing one time-shift element and one integratorformed using a loop circuit arranged as $D_0, D_1, \ldots, D_{N_f-1}$, respectively. The output value of each branch is the integration result obtained after accumulating the output values of each of the other branches. Each value is equal to the pulse cross-correlation value multiplied by the relative random polarity codes, $b_0, b_1, \ldots, b_{N_f-1}$.

Every symbol in the SLD receiver is composed of N_f frames, and the integration time in each frame is denoted as the sub-integration-window (SIW). For all frames, the SIW has the same width T_w, which is usually less than T_f and depends on the maximum delay spread of the channel. When the maximum delay spread of the channel is large, the SIW should be sufficiently large to capture most of the signal energy.

In the decision part of **Figure 1**, the value of every search and every branch is observed. The N_f output values from every search are then compared to find the largest value, and the synchronization process then comes to an end. For each search, the maximum value will occur in only one branch, and thus the symbol boundary position can be reliably estimated.

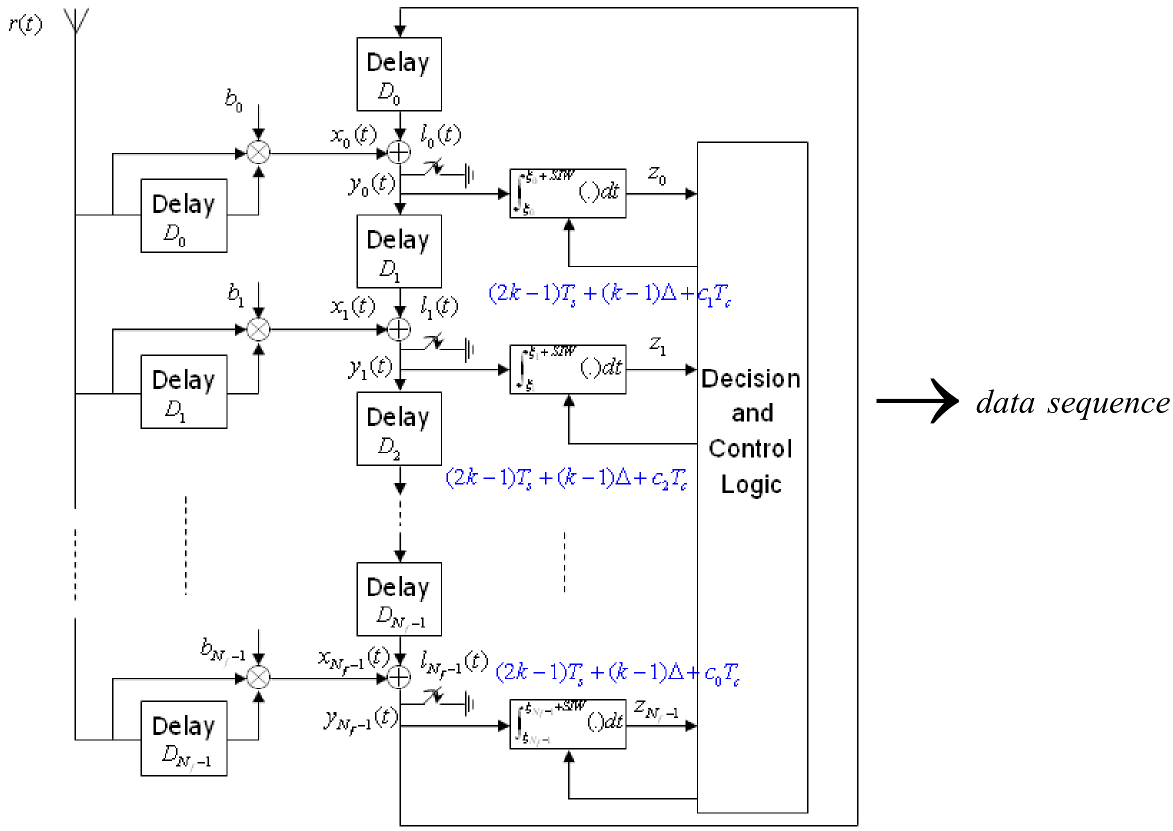

Figure 1. Architecture of SLD receiver.

For every branch, the output value can be expressed as

$$z_j = \sum_{m=\phi(j+1)}^{N_f-1+\phi(j+1)} b_{\phi(m)} \int_{\varsigma_j}^{\varsigma_j+SIW} r(t) r(t - D_{\phi(m)}) \delta(t - \sum_{n=m+1}^{N_f-1+\phi(j+1)} D_{\varphi(n)}) dt, \tag{8}$$

where $\phi(\cdot)$ denotes the modulo function of Nf, i.e., $\phi(\cdot) = \mathrm{mod}(\cdot, N_f)$.Note that $j \in [0, N_f - 1]$, $\varsigma_j = sp + (2k-1)T_s + (k-1)\Delta$

is the initial time of the integrator, *sp* is the initial searching time at the receiver end, Δ is the step shift , and *k* is the total number of searches performed (i.e., $k \in [1, {}^{2T_f}\!/\!_{\Delta}])$ [13].

4. Performance analyses

Manyprevious DHTR related studies have shown that the integration output noise term of TR-UWB systems has a Gaussian distribution. Thus, in the present study, the output of the SLD receiver is also assumed to beGaussian distributed, i.e.,

$$f_{z_j}(z) = \frac{1}{\sqrt{2\pi}\sqrt{Var[z_j]}} e^{-\frac{(z-E[z_j])^2}{2Var[z_j]}}, -\infty < z < \infty, \tag{9}$$

where $E\left[z_j\right]$ is the expected value of the output, $Var\left[z_j\right]$ is the variance of the output, and $f_{z_j}(z)$ is the pdf of the output end.

Assume that the distance from the target step position to the symbol boundary is denoted as ε, and has a uniform distribution.Let

$$E_b(\varepsilon) = (N_f)\sum_{l=1}^{L_k}\int_0^{SIW} E\{\alpha_{k,l}^2\} w^2(t-(l-1)T_c-\varepsilon)dt, \tag{10}$$

The mean and variance of the SLD receiver output can therefore be given respectively as

$$E[z_j] = d_{2k-2}E_b\left[(c_{k,0} - c_{\mathrm{mod}(j+1,N_f)})T_c - (k-1)\Delta + \sum_{p=0}^{\mathrm{mod}(j+1,N_f)} D_{k,p}\right], \tag{11}$$

$$\begin{aligned} Var[z_j] = &\left[\frac{1}{N_f}(e^{4\sigma^2}-1)E_b^2\left((c_{k,0}-c_{\phi(j+1)})T_c-(k-1)\Delta+\sum_{p=0}^{\phi(j+1)} D_{k,p}\right)\right]^2 \\ &+2\left[(1-e^{-\sigma^2})E_b\left((c_{k,0}-c_{\phi(j+1)})T_c-(k-1)\Delta+\sum_{p=0}^{\phi(j+1)} D_{k,p}\right)\right]^2 \\ &+2\left[\sqrt{N_f}\times e^{-\frac{\sigma^2}{2}}\sqrt{E_b\left((c_{k,0}-c_{\phi(j+1)})T_c-(k-1)\Delta+\sum_{p=0}^{\phi(j+1)} D_{k,p}\right)}\right]^2 \\ &+(N_f\times SIW)^2. \end{aligned} \tag{12}$$

The previous section has discussed the determination of the propagation delay, $\hat{\tau}_{k,0}$, from the candidate offsets. The following discussions analyze the probability of detection (PD) and the mean acquisition time (MAT) at the SLD receiver. The estimation criterion is defined as $[\hat{n},\hat{q}] = \underset{n\in[0,N_t-1]\, q\in[0,N_f-1]}{\arg\max} z_{k,q}[n]$. Note that the estimators $\hat{n}$ and $\hat{q}$ aim to find the values of n^* and q^* such that $\hat{t}_{symbol}^{(i)} = iT - (q^*T_f + n^*T_{shift})$, where T_{shift} is generally smaller than the pulse duration (T_p) in order to improve the timing precision. PD is then given by

$\hat{n}$

$$P_d(n^*,q^*) = \Pr\left\{\hat{n} = n^* \ and \ \hat{q} = q^*\right\} = \Pr\left\{z_{k,q^*}[n^*] = \max_{n,q} z_{k,q}[n]\right\}. \tag{13}$$

It is noted that for any given SNR, $P_d(n^*,q^*)$ generally depends on n^* since the noise terms of $z_{k,q}[n]$ are correlatedacross n due to the overlap of the integration windows[12].

For analytical tractability, the following discussions consider the coarse timing acquisition issue rather than the problem of estimating the true $\tau_{k,0}$. The aim of the coarse timing acquisition problem is to find the values of n^* and q^*which satisfy

$$-\frac{T_{shift}}{2} \le \left[iT - (q^*T_f + n^*T_{shift})\right]_{T_s} < \frac{T_{shift}}{2}, \tag{14}$$

whereT_{shift} is larger than the integration window size (T_w). For the SLD receiver considered in the present study, it is reasonable to assume that $z_{k,q}[n]$ are independent across n and q since the integration windows are non-overlapping. Assumethat the distance from the target step position to the symbol boundary is denoted as ε_τ (and is uniformly distributed over the interval $[-T_{shift}/2, T_{shift}/2]$). Assume also that each output based on this ε_τ is denoted as $z_{k,q}^{(\varepsilon_\tau)}[n]$. Denoting the pdf of $z_{k,q}^{(\varepsilon_\tau)}[n]$ as $f_{n,q}(z\,|\,\varepsilon_\tau)$, the PD for each n^*, q^* and ε_τ is given as

$$p(n^*,q^*\,|\,\varepsilon_\tau) = \int_{-\infty}^{\infty} f_{n^*,q^*}(z\,|\,\varepsilon_\tau) \prod_{n\ne n^*, q\ne q^*} \left(\int_{-\infty}^{z} f_{n,q}(x\,|\,\varepsilon_\tau)dx\right)dz. \tag{15}$$

Then, the PD for coarse timing acquisition is given by

$$P_d = \bigcup p(n^*,q^*\,|\,\varepsilon_\tau). \tag{16}$$

Sinceε_τ is uniformly distributed over $[-T_{shift}/2, T_{shift}/2]$, Eq. (15) can be integrated to obtain

$$P_d = \frac{1}{T_{shift}} \int_{-\frac{T_{shift}}{2}}^{\frac{T_{shift}}{2}} \left(\int_{-\infty}^{\infty} f_{n^*,q^*}(z\,|\,\varepsilon_\tau) \prod_{n\ne n^*} \left(\int_{-\infty}^{z} f_{n,q}(x\,|\,\varepsilon_\tau)dx\right)dz\right)d\varepsilon_\tau. \tag{17}$$

Note that the lower bound of PD can be derived with a similar form to that of Eq. (30) in [12]. Thus, the PD is given as

$$P_d \ge \underline{P_d} := \frac{1}{T_{shift}} \int_{-\frac{T_{shift}}{2}}^{\frac{T_{shift}}{2}} \left(\prod_{n\ne n^*} \int_{-\infty}^{\infty} f_{n^*,q^*}(z\,|\,\varepsilon_\tau) \int_{-\infty}^{z} f_{n,q}(x\,|\,\varepsilon_\tau)dxdz\right)d\varepsilon_\tau. \tag{18}$$

In accordance with the central limit theorem, $z_{k,q}^{(\varepsilon_\tau)}[n]$ can be considered to be Gaussian distributed. As a result, the lower bound in (18) can be expressed as

$$\underline{P_d} = \frac{1}{T_{shift}} \int_{-\frac{T_{shift}}{2}}^{\frac{T_{shift}}{2}} \left(\prod_{n\ne n^*} Q\left(\frac{E\left\{z_{k,q}^{(\varepsilon_\tau)}[n]\right\} - E\left\{z_{k,q}^{(\varepsilon_\tau)}\left[n^*\right]\right\}}{\sqrt{Var\left\{z_{k,q}^{(\varepsilon_\tau)}\left[n\right]\right\} + Var\left\{z_{k,q}^{(\varepsilon_\tau)}\left[n^*\right]\right\}}}\right)\right)d\varepsilon_\tau, \tag{19}$$

where$Q(\cdot)$ is the complementary cumulative distribution function (cdf) of a standard Gaussian random variable with zero mean and unit variance.

5. Simulation results

Figure 2 presents a histogram of the SLD output values.Note that in obtaining the results, the simulation parameters were specified as: CM4, *Nu*=1, SNR=0, and total outcomes=20000. It is seen that the output values have a Gaussian distribution, with the majority ofthe valuesbeing centralized around 0.5 ~ 1.0. In other words, the simulation results are consistent with the assumption in Section IV that the SLD receiver output is Gaussian distributed.

 ISSN: 2424-8460(O) 2251-2608(P)

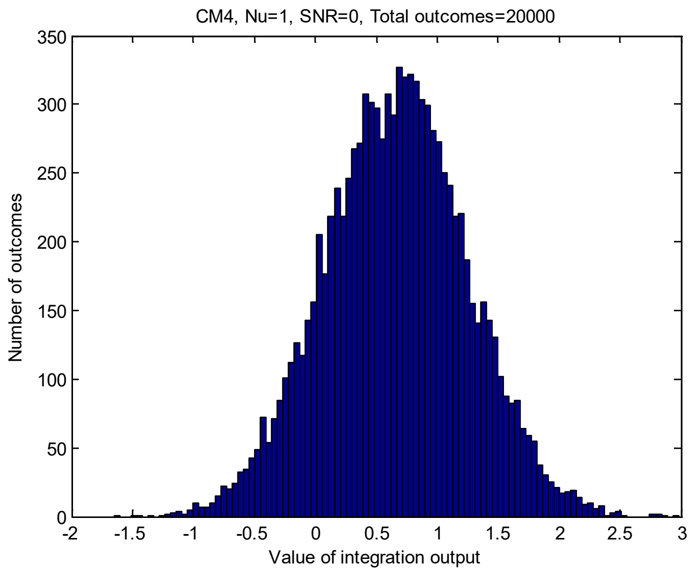

Figure 2. Histogram of SLD output value.

Figure 3 shows the mean value of the SLD receiver output. Note that the X-axis represents the shift value of the symbol boundary, while the Y-axis represents themean value of the integrator output.Note also that thesimulation settings were given as follows: CM4 and total number of outcomes=20000. The main peak in the mean value curve corresponds to the point when the correlated value reaches its maximum value, i.e., the two symbol boundaries in the original signal and the delayed signal, respectively, are perfectly matched. Meanwhile, the exponentially increasing and decreasing regions of the curve correspond to a partial correlation condition. Finally, the flat regions of the curve correspond to average output values of approximately zero; indicating that correlation of the two signals does not occur. The variance of the SLDoutput is shown in Fig.4.It is noted that the simulation results presented in Figure 3 and 4 are consistent with Eqs. (11) and (12). Thus, the validity of the theoretical model proposed in Section IV is confirmed.

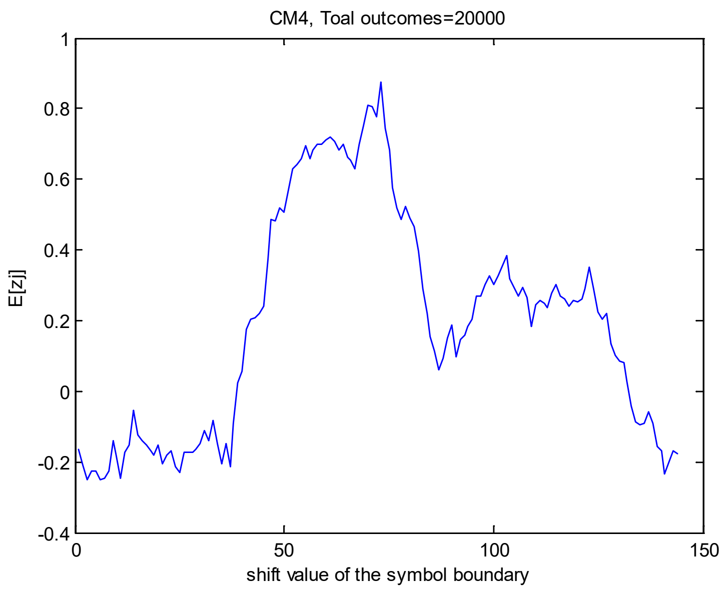

Figure 3. Simulation results for mean value of SLD output.

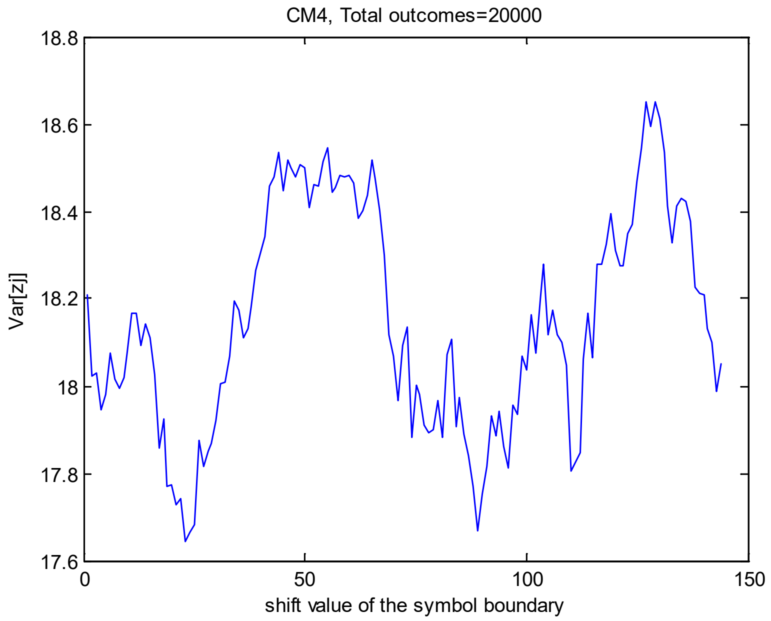

Figure 4. Simulation results for variance of SLD output.

The performance of the proposed PS-SLD algorithm was evaluated through a further series of simulations. The multipath channels were generated using the UWB channel model proposed in the IEEE 802.15.3a standard[5] with parameters of (1/Λ, 1/λ, Γ, γ)=(42.9, 0.4, 7.1, 4.3) ns. The channel impulse response was assumed to be invariant over one symbol duration. Furthermore, for each user *k*, an assumption was made that the propagation delay of the first arrival signal, $\tau_{k,1}$,was uniformly distributed over theinterval $[0, T_s)$ ns, where $T_s = N_f \times T_f$. The frame duration was specified as $T_f = 35$ ns, and each symbol contained $N_f = 32$ frames. The desired user(indexed as 0) was assigned a TH code $\{c_{0,j}\}_{j=0}^{N_f-1}$, which satisfies $D_{0,p} \neq D_{0,q}$ for each $p \neq q$. The remaining users were assigned random TH codes uniformly distributed over the interval $[0, N_c)$ with $N_c = 35$ and $T_c = 1$ ns. The training sequence for the data-aided (DA) TDT comprised a repeated pattern (1, 1, -1, -1) for all users, while for the PS-SLD algorithm, the transmitted data, $d_{0,i}$, were all1's. Finally, the integration window, T_w, was specified as 20 ns in all of thePS-SLD simulations.

Figure 5 presents the probabilities of detection (PD) ofthe PS-SLD algorithm obtained by computer simulations and Monte Carlo simulations, respectively. It is seen that the computer simulation results are in good agreement with the Monte Carlo simulation results obtained using Eq. (17). In other words, the validity of Eq. (17) is confirmed. Moreover, the Monte Carlo results obtained using Eqs. (18) and (19) are identical, and thus it is confirmed that the SLD receiver output, $z_{k,q}^{(\varepsilon_\tau)}[n]$, is indeed Gaussian distributed.

Figure 6 compares the PD performance of the PS-SLD algorithm with that of the TDT algorithm (Prop 4, *K*=1; and Prop 4, *K*=8). It is observed that the PS-SLD algorithm achieves a higher PD than the TDT algorithm with *K*=1for values of the SNR less than 6 dB. This result is reasonable since the SLD receiver uses the unique time interval between two successive pulses, i.e., $D_{k,p}$, as the time delay when correlating the received signal with the time-delayed signal. As a result, it is more robust to noise and interference than the dirty templates receiver. However, for all values of the SNR, the PD of the SLD receiver is worse than (or at best, equal to) that of the dirty templates receiver with *K* = 8.

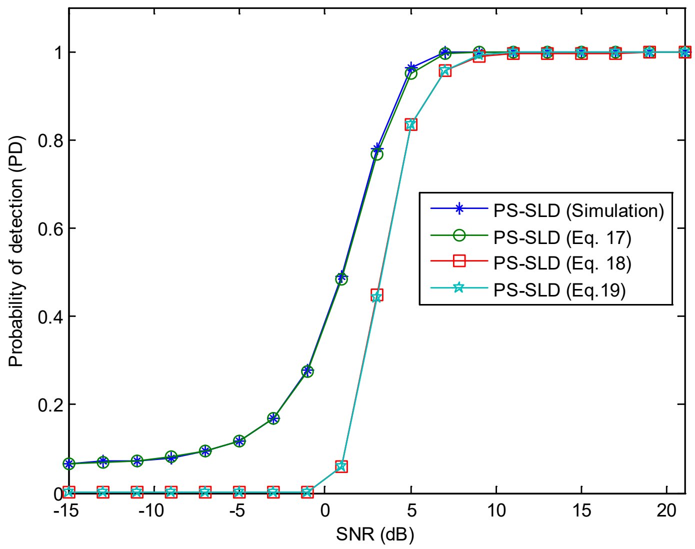

Figure 5. Theoretical and simulation results for probability of detection (PD) of PS-SLD receiver, T_{shift}=T_f.

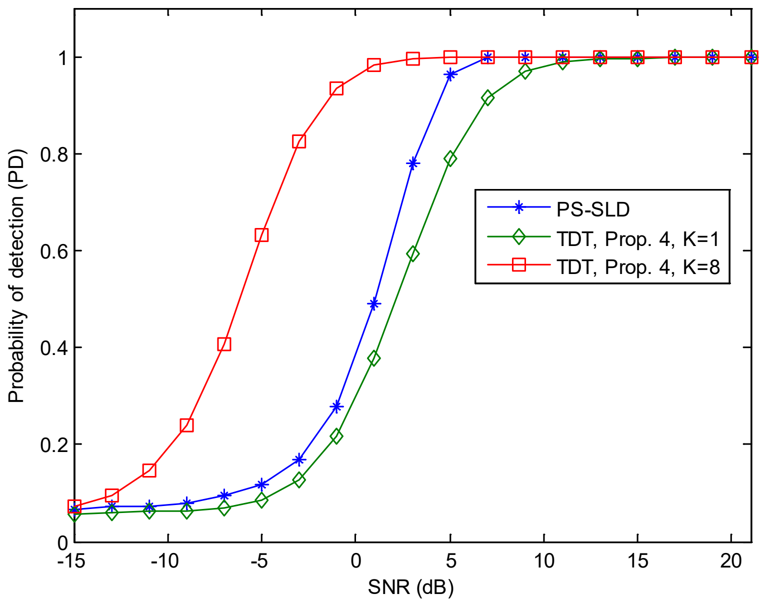

Figure 6: Probability of detection (PD) for PS-SLD receiver and TDT scheme, T_{shift}=T_f.

Figure 7 and 8compare the MSE performance of the PS-SLD and TDT algorithms for $T_{shift} = T_f$ and $T_{shift} = T_p$, respectively.It is seen that the MSE of the two algorithms is similar for $T_{shift} = T_f$ given K = 1.This result is to be expected. Moreover, the MSE of the PS-SLD algorithm is better than that of the TDT algorithms for $T_{shift} = T_p$ given a SNR value of more than ~ 6 dB.

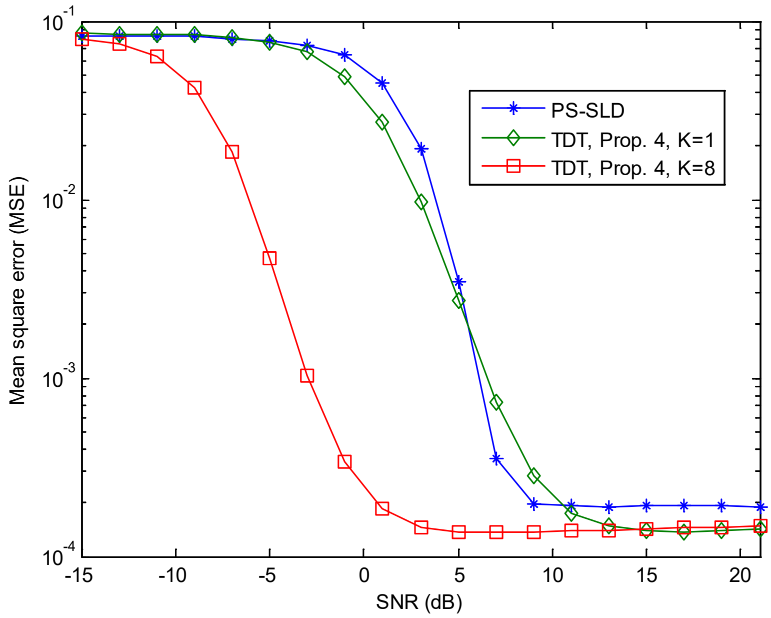

Figure 7. Normalized MSE of PS-SLD receiver and TDT scheme for T_{shift}=T_f.

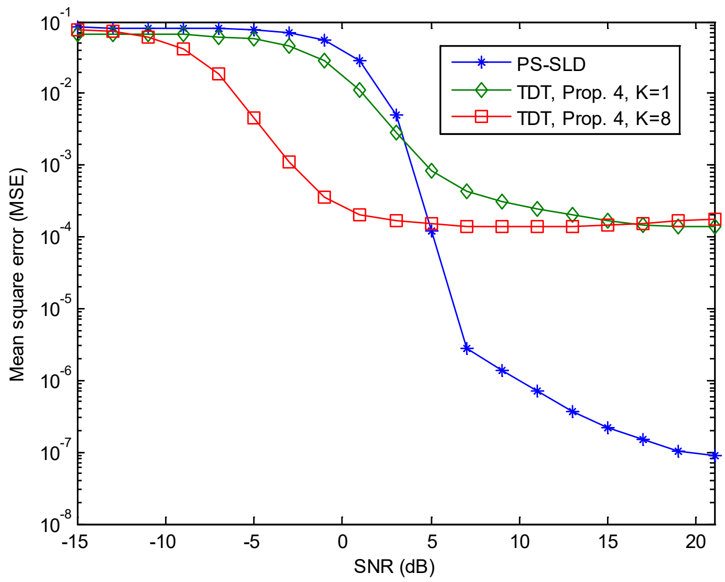

Figure 8. Normalized MSE of PS-SLD receiver and TDT scheme for T_{shift}=T_p.

6. Conclusions

This paper hasexamined the theoretical synchronization performance of DHTR receivers. The theoretical results for the mean and variance of the SLD integration output have been verified by simulations. The results have confirmed that the pdf of the SLD receiver output has a Gaussian distribution. This paper has also presented an effective timing synchronization scheme, designated as "parallel signal acquisition with shared looped delay-line" (PS-SLD). The hardware complexity of the PS-SLD scheme is higher than that of the TDT scheme. However, the uncertainty region of the PS-SLD algorithm is bounded within one T_f duration, which is only $1/N_f$ of that of the TDT algorithms. The simulation results have demonstrated that the PS-SLD algorithm achieves a higher PD and a lower normalized MSE than TDT algorithms in multipath environments given only one user (i.e.,

$K = 1$). Future studieswill evaluate the false alarm probability P_{fa} and mean acquisition time(MAT) of the PS-SLD receiver using the Markov chain model.

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