Cramér-Rao Lower Bounds for the Synchronization of UWB Signals

Jian Zhang
National ICT Australia (NICTA) and the Department of Telecommunications Engineering, Research School of Information Sciences and Engineering, The Australian National University, Canberra ACT 0200, Australia
Email: andrew.zhang@nicta.com.au

Rodney A. Kennedy
National ICT Australia (NICTA) and the Department of Telecommunications Engineering, Research School of Information Sciences and Engineering, The Australian National University, Canberra ACT 0200, Australia
Email: rodney.kennedy@anu.edu.au

Thushara D. Abhayapala
National ICT Australia (NICTA) and the Department of Telecommunications Engineering, Research School of Information Sciences and Engineering, The Australian National University, Canberra ACT 0200, Australia
Email: thushara.abhayapala@anu.edu.au

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We present Cramér-Rao lower bounds (CRLBs) for the synchronization of UWB signals which should be tight lower bounds for the theoretical performance limits of UWB synchronizers. The CRLBs are investigated for both single-pulse systems and time-hopping systems in AWGN and multipath channels. Insights are given into the relationship between CRLBs for different Gaussian monocycles. An approximation method of the CRLBs is discussed when nuisance parameters exist. CRLBs in multipath channels are studied and formulated for three scenarios depending on the way multipath interference is treated. We find that a larger number of multipaths implies higher CRLBs and inferior performance of the synchronizers, and multipath interference on CRLBs cannot be eliminated completely except in very special cases. As every estimate of time delay could not be perfect, the least influence of the synchronization error on the performance of receivers is quantified.

Keywords and phrases: ultra-wideband, synchronization, Cramér-Rao lower bounds.

1. INTRODUCTION

Ultra-wideband (UWB) is a promising technique in the application of short-range high-speed wireless communication and precise location tracking. Typically, ultranarrow pulses, such as Gaussian monocycles [1], are modulated to transmit information. These pulses could be narrower than 1 nanosecond. This brings very stringent synchronization requirements.

A UWB signal is basically a baseband signal without phase and carrier information, hence time delay estimation is the main task of a synchronizer. This synchronizer could be one in a simple single-pulse UWB system; however, due to the power limitation imposed by FCC [2], UWB pulses are generally combined with spread spectrum techniques, especially time hopping (TH), to achieve multiuser access, to ensure sufficient received energy, and to mitigate interference in existing wireless systems. Similar to traditional spread spectrum systems, the synchronization of a TH-UWB system is accomplished in two steps: code acquisition followed by code tracking. The former, involving the optimization of search strategies, tries to determine the phase of the incoming pseudonoise (PN) sequence within a fraction of chip width. The latter refers to the process of achieving and maintaining fine alignment of the chip boundaries of the incoming and locally generated PN sequences.

As UWB pulses are very narrow, very stringent synchronization requirements are incurred, and timing errors usually imply marked degradation of receiver performance [3]. Abundant research on the design and performance of synchronization systems have been reported in the literature, for example, [4, 5, 6]. These techniques can be transplanted into UWB systems with some modifications to meet the stringent timing requirement, as discussed in [7, 8, 9, 10]. Different to them, in this paper, we try to find some general performance limitations for UWB synchronizers, and provide guidelines for the system design within acceptable performance region.
It is known that in the presence of noise, perfect synchronization cannot be achieved. For UWB systems with stringent timing requirements, it is of special interest to characterize this synchronization error and its influence on the performance of detectors. This task becomes even more urgent when we realize that the radiated power of UWB signals is so low that the channel estimates could contain large errors and the performance of synchronizers could be largely deteriorated. Under these conditions, is it still possible for UWB synchronizers to reach a satisfying accuracy of timing locking? Some common performance parameters to evaluate synchronizers are tracking time, S-curve behavior, and probability of success. However, in order to provide benchmarks for actual UWB synchronizers, we are more interested in understanding their theoretical performance limits. In the theory of parameter estimation, Cramér-Rao lower bound (CRLB) is most widely used in evaluating the performance of estimates.

The CRLB [11] is a fundamental lower bound on the variance of any unbiased estimator. The analysis of CRLB for traditional systems is well founded [5, 12, 13, 14, 15, 16, 17, 18, 19], but for UWB, there is no systematic work reported yet to our knowledge. The evaluation of the CRLB is generally mathematically quite difficult when the observed signal contains, besides the parameter to be estimated, some nuisance parameters that are unknown [14, 19]. These nuisance parameters could be the transmitted data and, sometimes, multipath gains and delays which arise in fading channels. When the nuisance parameters are present, the modified CRLB (MCRB) [13, 14, 15], and the asymptotic CRLB (ACRLB) [14], are good approximations to the true CRLB at higher signal-to-noise ratio (SNR), and the lower-SNR limit of the CRLB is approximated in [18] by applying a Taylor Series expansion.

This paper is concerned with evaluating the CRLB for UWB signals. Both single-pulse systems and TH systems are considered. For TH, the CRLB should be a lower bound for the performance of code tracking. The structure of this paper is as follows. In Section 2, the problem is formulated. In Section 3, considering AWGN channels, the CRLBs for single-pulse systems are investigated in both cases of known and unknown transmitted data. Some insights into the relationship between CRLBs for different Gaussian monocytes are given explicitly. We also highlight an oversight in the lower-SNR approximation method [18] and provide a possible solution to remedy this problem by tightly locating the range of SNR $\gamma_s$. These results can be readily extended to a TH-UWB system in AWGN channels with minor modifications. In Section 4, we extend this work to more practical multipath channels while considering an unmodulated TH system. Depending on the way multipath interference is treated in a practical synchronizer, three scenarios are analyzed when multipath interference contributes as an increase of noise variance or multiple synchronization parameters. In Section 5, the influence of synchronization error on the performance of receiver is quantified, which may be the least influence a UWB correlator receiver can expect. Finally, numerical results are given in Section 6 to verify some analytical results and illustrate the effect of pulse truncation on CRLBs.

2. PROBLEM FORMULATION

Binary pulse position modulation (BPPM) and binary phase modulation (BPSK, or antipodal modulation) are considered here. Let $s(t)$ be the transmitted UWB signal. In a single-pulse system, $s(t) = \sum_i b_i \omega(t - iT_s)$ for BPSK, and $s(t) = \sum_i \omega(t - iT_s - b_i T_d)$ for BPPM, where $\omega(t)$ is a UWB pulse, $b_i \in \{-1, +1\}$ is the $i$th transmitted data, $T_s$ is the symbol period, and $T_d$ is the time offset of BPPM. In an unmodulated TH system, $s(t) = \sum_i s_i(t) = \sum_i \sum_j a_j(t - iT_s - jT_f - c_i T_c)$ where $s_i(t)$ is the $i$th transmitted symbol, $T_f$ is the frame width, $N_f$ is the number of frames in a symbol, $T_c$ is the chip width, and $c_i$ are the TH codes.

The UWB pulses considered are series of Gaussian monocycles $\omega(t; n, t_p)$, which are scaled and/or differentiated versions of the basic Gaussian waveform $\omega_0(t) = \exp(-2\pi t^2)$, that is, $\omega(t; n, t_p) = \omega_0^{(n)}(t/t_p)$, where the superscript $(n)$ stands for $n$-order differentiation with respect to $t$, and $t_p$ parameterizes the width of the pulse.

To ensure equal energy of monocycles, a coefficient $\epsilon(n, t_p)$ is introduced, and let $\omega(t) = \epsilon(n, t_p)\omega(t; n, t_p)$. Denote the energy of $\omega(t)$ as $E_p$ and symbol SNR as $\gamma_s$, then $\epsilon(n, t_p)$, depending on $n$ and $t_p$, satisfies

$$
\epsilon^2(n, t_p) = \frac{E_p}{\int_{-\infty}^{\infty} \omega^2(t; n, t_p) dt}.\tag{1}
$$

When passing through a pure AWGN channel $n(t)$, the received signal $r(t)$ becomes

$$
r(t) = s(t - \tau) + n(t),\tag{2}
$$

where every sample of $n(t)$ is Gaussian distributed with zero mean and variance $\sigma_n^2$, and $\tau$ is the timing delay to be estimated.

When passing through a frequency-selective fading channel, $h(t) = \sum_{\ell=1}^{\mathcal{L}} a_{\ell}\delta(t - \tau_{\ell})$, the received signal is given by

$$
r(t) = \sum_{\ell=1}^{\mathcal{L}} a_\ell s(t - \tau_{\ell}) + n(t),\tag{3}
$$

where $a_\ell$ and $\tau_{\ell}$ are real multipath gains and delays, respectively. Note that the time delay $\tau$ between the transmitter and the receiver is merged into $\tau_{\ell}$.

Due to the low-duty cycle of UWB signals, we assume the received signal is free of intersymbol interference (ISI) unless indicated otherwise. For the effect of ISI and the design of training sequence accordingly, the readers can refer to [20, 21].

For the AWGN model in (2), for the purpose of forming estimates based on $K$ independent observations, the received signal can be represented as a vector model

$$
r = s(b, \tau) + n,\tag{4}
$$
where \( \mathbf{r} = [r_1, \ldots, r_K], \mathbf{s} = [s_1, \ldots, s_K], \) and \( \mathbf{n} = [n_1, \ldots, n_K] \) are the sample vectors of the received signal \( r(t) \), the transmitted signal \( s(t - \tau) \), and the noise \( n(t) \), respectively, and \( \mathbf{b} = \{b_i\} \) is the transmitted data sequence.

Suppose an unbiased estimate \( \hat{\tau} \) of the time delay \( \tau \) can be generated from (4). Then the estimation error variance is lower bounded by the CRLB \( E_r[(\hat{\tau} - \tau)^2] \geq \text{CRLB}(\tau) \), where

\[
\text{CRLB}(\tau) = \left( E_{\mathbf{r}|\tau} \left[ -\frac{d^2}{d\tau^2} \ln \left( p(\mathbf{r}|\tau) \right) \right] \right)^{-1}.
\]

In (5), the conditional pdf \( p(\mathbf{r}|\tau) \) is the likelihood function of \( \tau \), and the expectation \( E_{\mathbf{r}|\tau}[^{}\cdot^{}] \) is with respect to \( p(\mathbf{r}|\tau) \).

The likelihood function \( p(\mathbf{r}|\tau) \) can be obtained by averaging the joint likelihood function \( p(\mathbf{r}|\mathbf{b}, \tau) \) over the a priori distribution of the data \( \mathbf{b} : p(\mathbf{r}|\tau) = E_{\mathbf{b}}[p(\mathbf{r}|\mathbf{b}, \tau)]. \) When \( \mathbf{b} \) is known, \( p(\mathbf{r}|\tau) = p(\mathbf{r}|\mathbf{b}, \tau). \)

Since the additive noise \( n(t) \) is white and zero mean, the joint conditional pdf \( p(\mathbf{r}|\mathbf{b}, \tau) \) can be expressed as

\[
p(\mathbf{r}|\mathbf{b}, \tau) = \prod_{k=1}^{K} \frac{1}{\sqrt{2\pi\sigma^2_0}} \exp \left( -\frac{1}{2\sigma^2_0} (r_k - s_k)^2 \right) \]
\[
= \left( \frac{1}{\sqrt{2\pi\sigma^2_0}} \right)^K \exp \left( -\frac{1}{2\sigma^2_0} \sum_{k=1}^{K} (r_k - s_k)^2 \right), \quad (6)
\]

Applying the signal orthogonal expressions \([6, \text{page 335}]\) or letting the number of samples \( K \) go to infinity \([11, \text{page 274}]\) (or from the standpoint of generating sufficient statistics), we have

\[
\sum_{k=1}^{K} (r_k - s_k)^2 = \int_{T_o} \left[ r(t) - s(t - \tau) \right]^2 dt, \quad (7)
\]

where \( T_o \) is the observation period.

Now, a continuous-time equivalent of \( p(\mathbf{r}|\mathbf{b}, \tau) \) can be developed. Considering the subsequent operations of logarithm and differentiation, only terms related to \( \mathbf{b} \) and \( \tau \) will be retained. Then the evaluation of \( p(\mathbf{r}|\mathbf{b}, \tau) \) is equivalent to evaluating the likelihood function

\[
\Lambda(\mathbf{b}, \tau) = \exp \left( \frac{1}{2\sigma^2_0} \left( 2 \int_{T_o} r(t) s(t - \tau) dt - \int_{T_o} s^2(t - \tau) dt \right) \right), \quad (8)
\]

The process from (4) to (8) can be applied to the multi-path model (3) with minor modifications.

### 3. CRLB FOR SINGLE-PULSE SYSTEMS IN AWGN CHANNELS

#### 3.1. CRLB with known transmitted data

The CRLB with known \( \mathbf{b} \), further derived from (8) or directly from [15], has the form

\[
\text{CRLB}(\tau; \mathbf{b}) = \frac{1}{\int_{T_o} s^2(t - \tau) dt}, \quad (9)
\]

where \( s(t - \tau) \) denotes first partial differentiation with respect to \( \tau \).

Assuming that the pulse is strictly restricted within a symbol period, and \( T_o = N T_s \), where \( N \) is the number of symbols contained in the observation period (one pulse per symbol in this case), then for both BPSK and BPPM, the denominator in (9) equals \( N \int_{T_o} \omega^2(t - \tau) dt \). For a specific monocycle, the lower variance bound becomes

\[
\text{CRLB}(\tau; b) = \frac{1}{N Y_s} \int_{T_o} \omega^2(t - \tau; n, t_p) dt, \quad (10)
\]

where the symbol SNR is \( y_s = E_p/\sigma^2_0 \).

If the symbol period \( T_s \) is large enough so that most of the energy of the pulse concentrates within \( T_o \), we can express (10) in the frequency domain:

\[
\text{CRLB}(\tau; b) = \frac{1}{N Y_s} \int_{-\infty}^{+\infty} \left| W(f; n, t_p) \right|^2 df, \quad (11)
\]

where \( W(f; n, t_p) \) is the Fourier transform of \( \omega(t; n, t_p) \).

According to the properties of the Fourier transform of derivatives of functions, we find explicit relationships existing between the CRLBs of monocycles with different \( n \) but the same \( t_p \), that is,

\[
\frac{\text{CRLB}(\tau; b)_n}{\text{CRLB}(\tau; b)_{n+1}} = \frac{\int_{-\infty}^{+\infty} \left| W(f; n, t_p) \right|^2 df \cdot \int_{-\infty}^{+\infty} f^4 \left| W(f; n, t_p) \right|^2 df}{\left( \int_{-\infty}^{+\infty} f^2 \left| W(f; n, t_p) \right|^2 df \right)^2} > 1, \quad (12)
\]

where the inequality follows from an application of Schwarz’s inequality. This inequality implies that monocycles with higher-order differentiation have the potential for better performance in the sense of lower synchronization error variance.

For monocycles with different \( t_p \) but with the same \( n \), the ratio between their CRLBs can be found as

\[
\frac{\text{CRLB}(\tau; b)_{t_p_1}}{\text{CRLB}(\tau; b)_{t_p_2}} = \left( \frac{t_p_1}{t_p_2} \right)^2, \quad (13)
\]

which implies that monocycles with smaller \( t_p \) (narrower effective pulse width) have the potential for better synchronization performance.

#### 3.2. CRLB with unknown randomly transmitted data

For PPM, the uncertainty of time jitter introduced by modulation will cause large synchronization error when the transmitted data is random and unknown. When further methods are adopted to solve this problem, the CRLB analysis in these cases will usually be based on a system model similar to the one with known data. Hence we only consider BPSK-UWB signals in this subsection.
For BPSK, the likelihood function in (8) becomes

\[ \Lambda(b, \tau) = \exp \left( \sum_{i=1}^{N} \frac{1}{\sigma_0^2} (b_i y(\tau) - b_i^2 y_i) \right), \]  

where \( y(\tau) = \int r(t) \omega(t-\tau) dt \).

Dropping the constant term \( y_i, \sum_{i=1}^{N} (b_i^2) = N y_i \), we obtain the log-likelihood function of \( p(r|\tau) \) as

\[ \mathcal{L}(r; \tau) = \ln p(r|\tau) = \ln \mathbb{E}_b [\Lambda(b, \tau)] = \sum_{i=1}^{N} \ln \mathbb{E}_b \left[ \exp \left( \frac{1}{\sigma_0^2} b_i y(\tau) \right) \right] = N \ln \cosh \left( \frac{1}{\sigma_0} y(\tau) \right). \]  

By differentiating \( \mathcal{L}(r; \tau) \) twice with respect to \( \tau \), we get

\[ \frac{\partial^2 \mathcal{L}(r; \tau)}{\partial \tau^2} = \frac{N}{\sigma_0^2} \tanh \left( \frac{y(\tau)}{\sigma_0} \right) y(\tau) \]
\[ + \frac{N}{\sigma_0^2} \left( 1 - \tanh^2 \left( \frac{y(\tau)}{\sigma_0} \right) \right) \dot{y}(\tau), \]  

where \( \dot{y}(\tau) \) and \( \ddot{y}(\tau) \) denote first and second derivatives of \( y(\tau) \) with respect to \( \tau \).

Due to the nonlinear function \( \tanh(\cdot) \) in (16), an analytical solution for \( F_{\tau|\bar{r}}[\partial^2 \mathcal{L}(r; \tau)/\partial \tau^2] \) is infeasible.

Since the pulse energy is restricted to be very low by the FCC [2] (the maximum power of a transmitted pulse with bandwidth 7 GHz is only 0.5 mW), one can refer to the lower-SNR limit of CRLB in [18], applying a Taylor extension of the likelihood function \( p(r|b, \tau) \), to obtain a similar result for UWB. One thing we wish to emphasize here is that, in [18], the statistical property of the likelihood function \( \mathcal{L}(u, \tau) \) (original notation in [18]) is somewhat ignored.

Due to \( \mathcal{L}(u, \tau) \) containing Gaussian variables with variance comparable to the reciprocal of symbol SNR, more care is needed when dropping the higher-order terms in Taylor extension according to the lower symbol SNR assumption. A similar problem arises in an alternative method we introduce below, where this ambiguity is revealed further, and resolved by tightly locating the value of the symbol SNR.

The alternative method we suggest is also based on approximation. The basic idea is to find best-fitting functions for \( \ln(\cosh(\cdot)) \) in a piecewise fashion. To make analysis tractable, these functions are polynomials with order smaller than 3. But they should not be constructed by only considering the goodness of fit due to the succeeding expectation operation. This is because \( y(\tau) \) is a random variable and when we deal with the expectation operation, we have to make sure that all the possible samples of \( y(\tau) \) are involved. Although integrating these polynomials in segments is feasible, it cannot produce a closed form result and is still a numerical method. Instead, we try to construct each polynomial in which the variable space supports the sampling space.

It seems impossible as the pdf of \( y(\tau) \) distributes in the entire one-dimension real space. We overcome this obstacle by assuring that most of the samples (say, 99%) are located in the interval of interest.

With this criterion in mind, we find that a three-segment approximation is a good choice by studying the shape of the waveform \( \ln(\cosh(x)) \). A detailed discussion is shown in Appendix A. Examples of such three lower-order (\( \leq 2 \)) polynomials are

\[ \ln(\cosh(x)) \approx \begin{cases} 0.3x^2 + 0.14x - 0.018, & |x| < 1.5, \\ 0.000034x^2 + x - 0.69, & 1.5 \leq |x| \leq 2.5, \\ x - 0.69, & |x| > 2.5. \end{cases} \]

(17)

The root mean squared approximation errors are 0.0081, 0.0091, 0.0031 for the three pieces, respectively. The ranges of corresponding SNR \( \gamma \) are \([-\infty, -6.25] \) dB, \([10.3, 10.8] \) dB, and \([10.8, +\infty] \) dB, respectively, which can be determined according to the way addressed in Appendix A.

Due to nonexistence of a polynomial with goodness of fit and a fully covered sampling space simultaneously, an appropriate interval corresponding to SNR range of \([-6.25, 10.3] \) dB cannot be found.

Representing a general 2-order polynomial function as \( \ln(\cosh(x)) \approx f(x) = ax^2 + b|x| + c \), \( \gamma \in [x_0, x_2] \), where \( 0 \leq x_1 < x_2 \), we derive the CRLB based on it below.

The reciprocal of the CRLB can be calculated as

\[ -E_{\tau|\bar{r}} \left[ \frac{\partial^2 \mathcal{L}(r; \tau)}{\partial \tau^2} \right] = -N E_{\tau|\bar{r}}[2ax\dot{x} + 2ax^2 + bx^2], \]  

(18)

where we utilize

\[ \frac{d^2|\chi|}{d\tau^2} = \frac{d^2}{d\tau^2}(\sqrt{x^2}) = \frac{d}{d\tau}\ddot{x} = \ddot{x}. \]

(19)

As shown in Appendix B, these expectations are given by

\[ E_{\tau|\bar{r}}[\dot{x}^2] = \frac{1}{\sigma_0^2} \int_{\mathcal{T}_r} \dot{\omega}^2(t-\tau) dt, \]
\[ E_{\tau|\bar{r}}[\ddot{x}^2] = \frac{\gamma + 1}{\sigma_0^2} \int_{\mathcal{T}_r} \omega^2(t-\tau) dt, \]
\[ E_{\tau|\bar{r}}[x^2] = -\frac{1}{\sigma_0^2} \int_{\mathcal{T}_r} \ddot{\omega}^2(t-\tau) dt. \]

(20)

Then for a specific monocyte \( \omega(t; n, t_p) \), the CRLB is

\[ \text{CRLB} = \frac{1}{N(2\gamma + b)} \frac{1}{\mathcal{T}_r} \int_{\mathcal{T}_r} \omega^2(t-\tau; n, t_p) dt. \]

(21)

By substituting \( a \) and \( b \) with the coefficients of polynomials in (17), the CRLBs for different \( \gamma \) are readily obtainable. It is clear that the relationship between CRLBs for monocytes with different \( n \) or \( t_p \) is identical to that when the transmitted data is known.
By comparing (10) and (21), we find that CRLB(\(\tau; b\))/CRLB(\(\tau\)) = 2\(\alpha\gamma_s + b\). Referring to (17), it is obvious that the CRLB with unknown data is always larger than that with known data for the lower SNR case, and converges to CRLB(\(\tau; b\)) in the higher-SNR case, which coincides with the attributes of ACRB given in [14].

4. CRLB FOR TH-UWB SYSTEMS IN SELECTIVE FADEING CHANNELS

When the channel is AWGN, the analysis and results in Section 3 can be applied to TH-UWB systems with minor modifications. The change can be merged into the symbol SNR \(y_i\), that is, \(y_i\) equals the ratio between the energy of \(N\) pulses and the noise variance \(\sigma_n^2\) for TH-UWB systems. In this section, we will focus on selective fading channels.

Synchronization in selective fading channels is a challenging task. The performance largely depends on the schemes and algorithms. Based on the way multipath signals are treated, these systems can be divided into three categories. Accordingly, we consider the CRLB for each of them. Since CRLB with unknown data is straightforward but computationally complex as derived in Section 3, we only consider the case of known data \(b\) here.

4.1. Passive methods: regarding multipath signals as interference

This refers to methods that apply general synchronizers, such as early/late gates, while treating multipath signals as interference [22, 23], or partly utilizing multipath energy [24], or using a whitening filter before a synchronizer [25]. The effect of multipath interference on synchronizers has been studied in [23, 25, 26, 27, 28, 29]. From the viewpoint of CRLB, all these methods can be generalized to a model in which only a specific multipath is of interest. Mathematically, we can represent this model as

\[
\tau(t) = a_m s(t - \tau_m) + \sum_{\ell=1,\ell\neq m}^L a_{\ell}s(t - \tau_{\ell}) + n(t),
\]

where \(a_m\) and \(\tau_m\) are the parameters to be estimated.

Generally, the received signal \(\tau(t)\) first passes through a PN code correlator \(s(t - \hat{\tau}_m)\), where \(\hat{\tau}_m\) is the preestimated delay, so that the energy of all pulses in a symbol are collected to make an estimation. Then the model in (22) can be further written as

\[
\begin{align*}
\tau_f(\hat{\tau}_m) = a_m \sum_{i=1}^N \phi(\hat{\tau}_m + iT_s - \tau_{m}) \\
+ \sum_{i=1}^N \sum_{\ell=1,\ell\neq m} a_{\ell}\phi(\hat{\tau}_m + iT_s - \tau_{\ell}) + n_f(\hat{\tau}_m),
\end{align*}
\]

where

\[
\begin{align*}
\tau_f(\hat{\tau}_m) &= \sum_{i=1}^N \int_{iT_s} r(t)s_i(t - \hat{\tau}_m)dt, \\
\phi(\nu) &= \int_{iT_s} s(t - \nu)s(t)\,d\nu, \\
n_f(\hat{\tau}_m) &= \sum_{i=1}^N n(t)s_i(t - \hat{\tau}_m)dt.
\end{align*}
\]

Successful detection requires sampling \(\tau_f(\hat{\tau}_m)\) at \(\hat{\tau}_m = iT_s + \tau_m\) accurately.

Each sample of \(n_f(\hat{\tau}_m)\) is Gaussian distributed with zero mean and variance \(\sigma_n^2\). The component \(\tau_m\), containing interchip interference and ISI, is hard to model and evaluate without prior knowledge of TH codes and multipath delay. To make the analysis mathematically tractable, here we assume that \(\tau_m\) is Gaussian distributed with mean \(m_{\nu}\) and variance \(\sigma_n^2\). In Appendix C, more information is given on the parameters of this distribution.

Recall that when considering the CRLB for TH-UWB synchronizers in the phase of code tracking, it is reasonable to assume that \(\phi(t - \tau_m)\) is restricted in an interval \([-T_\phi, T_\phi]\), where \(T_\phi\) is smaller than half of the frame period \((T_f < T_f/2)\). Then the sum of \(\phi(t - \tau_m)\) for \(N\) symbols, \(\sum_{i=1}^N \phi(t - iT_s - \tau_m)\), equals \(N\phi(t - \tau_m)\). Now, the estimation problem can be reformed as

\[
r_f(t) = a_mN\phi(t - \tau_m) + n_a + n_f(t),
\]

which is a problem of multiple parameters estimation in a Gaussian interference.

Although \(a_m\) and \(\tau_m\) are correlated via the mean power profile of fading, they are usually treated as unknown and deterministic parameters, and nonrandom parameter estimation techniques are applied, as the statistical relationship between them is hardly predictable. This means they are not a function of each other any more. Strictly speaking, \(\tau_m\) is the only synchronization parameter, and CRLB(\(\tau_m\)) can be obtained when regarding \(a_m\) as a nuisance parameter. However, it is known that joint estimation of \(\tau_m\) and \(a_m\) usually gives lower CRLB for \(\tau_m\) than that in a separate-estimation case [11, 14, 19]. Hence we will focus on joint estimation and generate CRLB(\(a_m\)) as a byproduct.

Representing (25) as a vector form \(\tau_f = a_m \Phi + n_a + n_f\) and applying the similar process from (6) to (8), the joint log-likelihood function \(\mathcal{L}(\tau_f; a_m, \tau_m)\) can be obtained as

\[
\begin{align*}
\mathcal{L}(\tau_f; a_m, \tau_m) &= -\frac{N}{2(\sigma_n^2 + \sigma_f^2)} \int_{2T_f} (Na_m^2\phi^2(t - \tau_m) - 2a_m r_f(t)\phi(t - \tau_m) \\
&\quad + 2m_n a_m \phi(t - \tau_m))dt.
\end{align*}
\]

1In [11, page 309], a general equation is provided for the CRLB of any unbiased estimate in colored noise. But a closed-form solution is not readily available.
Lower bounds on the variances of estimates for the components of $a_m$ and $\tau_m$ are given in terms of the diagonal elements of the inverse of the Fisher information matrix $J^{-1}$ [11]. In this example,

$$
J = \begin{pmatrix}
-\mathbb{E}\left[ \frac{\partial^2 \mathcal{L}(\mathbf{r}_i; a_m, \tau_m)}{\partial a_m^2} \right] & -\mathbb{E}\left[ \frac{\partial^2 \mathcal{L}(\mathbf{r}_i; a_m, \tau_m)}{\partial a_m \partial \tau_m} \right] \\
-\mathbb{E}\left[ \frac{\partial^2 \mathcal{L}(\mathbf{r}_i; a_m, \tau_m)}{\partial a_m \partial \tau_m} \right] & -\mathbb{E}\left[ \frac{\partial^2 \mathcal{L}(\mathbf{r}_i; a_m, \tau_m)}{\partial \tau_m^2} \right]
\end{pmatrix},
$$

(27)

where the expectation $\mathbb{E}[\cdot]$ is with respect to $p(\mathbf{r}_i; a_m, \tau_m)$.

Noting $\phi(t - \tau_m)$ and $a_m$ are mutually independent, the elements of $J$ can be calculated as

$$
J_{11} = C \int_{2T_\phi} \phi(t - \tau_m) dt,
$$

$$
J_{12} = J_{21} = C a_m \int_{2T_\phi} \phi(t - \tau_m) \phi(t - \tau_m) dt,
$$

$$
J_{22} = C a_m^2 \int_{2T_\phi} \phi^2(t - \tau_m) dt,
$$

where $C$ is a constant defined as $C \equiv N^2/(\sigma_\phi^2 + \sigma_n^2)$.

The crossterms $J_{12}$ and $J_{21}$ will vanish if we extend the observation period $T_\phi$ till $\phi(T_\phi) \approx 0$. Then the CRLBs for $\tau_m$ and $a_m$ are

$$
\text{CRLB}(\tau_m) = \frac{1}{J_{11}} = \left( C \int_{2T_\phi} \phi^2(t - \tau_m) dt \right)^{-1},
$$

$$
\text{CRLB}(a_m) = \frac{1}{J_{22}} = \left( C \int_{2T_\phi} \phi^2(t - \tau_m) dt \right)^{-1}.
$$

(29)

It is clear that the estimation of the time delay $\tau_m$ depends on the amplitude of the multipath given that $C$ is the same for all multipath signals, while the estimation of $a_m$ could be independent of $\tau_m$ supposing we extend the observation period appropriately. For the performance of synchronizer, the multipath interference contributes as an increased estimate variance.\(^2\)

Depending on the Gaussian approximation for the multipath interference $n_m$, $C$ may be related to a specific monocyte, hence the relationship between CRLB for different monocytes cannot be claimed directly.

Finally, we wish to say a little more on the relationship between our model in this section and practical systems. In the literature on synchronizers for spread spectrum systems such as CDMA, we can always find the terms fading bandwidth, tracking loop bandwidth, and predetection bandwidth and discussions on how the relationship between them affects the performance of synchronizers in a multipath channel (e.g., [26, 27, 28]). Briefly, the relationship between these bandwidths determines the degree of multipath interference entering the final decision part of the synchronizer. Considering our model, the effect can be regarded as a reduction of noise variance $\sigma_n^2$.

### 4.2. Positive joint detection of multipath signals

We refer to the method of jointly detecting fading amplitude and delay of all the multipath signals as a positive one. For CDMA, this method has been well studied in literature such as [16, 17, 25]. And the derivation of CRLB for CDMA systems can be found in [16, 17, 30]. Here, following the process in Section 3, we study the CRLB using joint detection for a UWB system where the parameters $a = [a_1, \ldots, a_L, \ldots, a_L]_{1 \times L}$ and $\mathbf{r} = [r_1, \ldots, r_\ell, \ldots, r_L]_{1 \times L}$ to be estimated are treated as unknown but deterministic.

Beginning with (3), similar to the derivation from (4) to (8), we can obtain the log-likelihood function $\mathcal{L}(\mathbf{r}; \mathbf{r}, \mathbf{a})$ as

$$
\mathcal{L}(\mathbf{r}; \mathbf{r}, \mathbf{a}) = \frac{1}{\sigma_0^2} \int_{T_\phi} r(t) \sum_\ell a_\ell s(t - \tau_\ell) dt
$$

$$
- \frac{1}{2\sigma_0^2} \int_{T_\phi} \left[ \sum_\ell a_\ell s(t - \tau_\ell) \right]^2 dt.
$$

(30)

After some manipulation, the Fisher information matrix $J$ has the structure

$$
J = \begin{pmatrix}
J_{rr}[\ell, m] & J_{ra}[\ell, m] \\
J_{ar}[\ell, m] & J_{aa}[\ell, m]
\end{pmatrix},
$$

(31)

where $J_{rr}$, $J_{ra}$, $J_{ar}$, and $J_{aa}$ are all $L \times L$ matrices with $[\ell, m]$th elements:

$$
J_{rr}[\ell, m] = \frac{1}{\sigma_0^2} \int_{T_\phi} a_\ell a_m s(t - \tau_\ell) s(t - \tau_m) dt,
$$

$$
J_{ra}[\ell, m] = \frac{1}{\sigma_0^2} \int_{T_\phi} s(t - \tau_\ell) s(t - \tau_m) dt,
$$

$$
J_{ar}[\ell, m] = J_{ra}[m, \ell] = \frac{1}{\sigma_0^2} \int_{T_\phi} a_\ell s(t - \tau_\ell) s(t - \tau_m) dt,
$$

(32)

respectively.

The CRLB for $\tau_m$ is just the $m$th diagonal element of the inverse of $J$. Use $m = 1$ as an example and rewrite the matrix $J$ as

$$
J = \begin{pmatrix}
J_{11} & \mathbf{B} \\
\mathbf{C} & \mathbf{D}
\end{pmatrix};
$$

(33)

we have

$$
\text{CRLB}(\tau_1) = J_{11}^{-1} + J_{11}^{-1} \mathbf{B} (\mathbf{D} - \mathbf{C} J_{11}^{-1} \mathbf{B})^{-1} \mathbf{C} J_{11}^{-1}
$$

$$
= J_{11}^{-1} + J_{11}^{-1} \tilde{\mathbf{B}}_{11}^{-1} \mathbf{C}
$$

$$
\geq J_{11}^{-1},
$$

(34)

(35)

(36)

where $\tilde{\mathbf{B}}_{11}$ is called the Schur complement of $J_{11}$ [31, page 175]. Since $J$ is nonnegative definite, the Schur complement matrix $\tilde{\mathbf{B}}_{11}$ is also nonnegative definite, and so is $\tilde{\mathbf{B}}_{11}^{-1}$. At the same
time, $B$ is the transpose of $C$ since $J$ is a symmetric matrix in this case. Thus we get $BJ_{11}C \succeq 0$ and the inequality in (36) follows immediately. Whenever $J_{11} > 0$, we can get the inequality in (36) more readily according to

$$\text{CRLB} \left( \tau_1 \right) = \left( J_{11} - BD^{-1} C \right)^{-1} > J^{-1}.$$  \hfill (37)

As $J_{11}^{-1}$ can be regarded as the CRLB in an AWGN channel with a known scalar of amplitude, this inequality implies that the CRLB in joint detection is always larger than that in the single-parameter estimation in an AWGN channel. Then an interesting question arises, whether more multipath means higher CRLB and inferior performance of synchronizers accordingly.

We consider a channel with $L - 1$ multipath signals. The Fisher information matrix $J'$ can be written as

$$J' = \begin{pmatrix} J_{11} & B \\ C & D' \end{pmatrix},$$  \hfill (38)

with

$$D' = \begin{pmatrix} D_{1} & 0 \\ 0 & 0 \end{pmatrix},$$  \hfill (39)

where $0$ is an $(L-2) \times 1$ zero vector and $\dagger$ stands for transpose operation. Then the CRLB with $L - 1$ multipath is

$$\text{CRLB} \left( \tau_1 \right)_{L-1} = \left( J_{11} - BD^{-1} C \right)^{-1}.$$  \hfill (40)

Comparing $BD^{-1} C$ and $BD'^{-1} C$ gives

$$BD^{-1} C - BD'^{-1} C = B \left( D^{-1} - \begin{pmatrix} D_{1}^{-1} & 0 \\ 0 & 0 \end{pmatrix} \right) C \geq 0,$$  \hfill (41)

where the inequality in (42) yields from the fact that $D^{-1} - D'^{-1}$ is a nonnegative definite matrix as can be proven according to the property of partitioned nonnegative definite matrices (see, e.g., [31, page 178] and let $D^{-1} = A$ in equation (6.10)).

Since $J_{11} > 0$, we have

$$\text{CRLB} \left( \tau_1 \right)_{L} > \text{CRLB} \left( \tau_1 \right)_{L-1},$$  \hfill (43)

which shows that more multipath does lead to higher CRLB and inferior performance of synchronizers. Since the number of multipaths is closely relevant to the bandwidth of monocycles, we conclude that narrower monocycles will very likely cause larger CRLBs. We did not say “absolutely” because all other variables besides $D$ during this derivation are assumed unchanged, but it could be unrealistic when different monocycles are applied.

Another key factor with influence on CRLB is the choice of TH codes. When the autocorrelation of TH codes is ideal, both the CRLBs in this subsection and Section 4.2 will be the same and similar to the one in an AWGN channel.

### 4.3. Active methods: cancellation of interference?

From Sections 4.1 and 4.2, we have seen that the performance of synchronizers is deteriorated by the multipath interference. It is natural to ask whether the multipath interference can be mitigated or fully eliminated before entering the decision part of a synchronizer.

As shown for CDMA systems in [27], it is possible to remove part of multipath interference in UWB systems. However, unless the correlation of TH codes is ideal, the total removal of multipath interference is impossible due to the existence of $n(t)$. This is because, from Section 4.2, we see that any estimate of parameters, including amplitude and delay, even though unbiased, may still have a nonzero variance in the present of noise. The CRLB can generally be achieved by maximum likelihood estimation asymptotically (when the number of observation samples goes to infinity), and the estimation error becomes Gaussian distributed with zero mean and variance equivalent to the CRLB [5, 11]. Therefore, the final signal with a pair of synchronization parameters of interest contains the sum of $2(L - 1)$ Gaussian variables, which has a variance larger than the variance of $n(t)$. Since CRLB is proportional to the variance of (interference and) noise, the CRLBs for this pair of parameters will be larger than those in a single-path channel. So no matter how perfect the structure and algorithm to remove multipath signal are, the effect of multipath interference can only be mitigated but cannot be canceled thoroughly. This result also partly explains why more multipath generally leads to higher CRLBs.

However, there are some special cases when multipath interference becomes negligible. For example, when the maximal multipath delay is smaller than the frame period in a single-pulse system, multipath signals do not interfere with each other due to the low duty cycle of UWB signal structure.

### 5. Influence of synchronization error on BER

As every estimate of time delay could not be perfect, we use an example to show the influence of synchronization error on the performance of receivers in UWB systems.

We consider a BPSK modulated single-pulse signal in an AWGN channel like that in Section 3. A correlator receiver [32, 33] is used to detect the signal.

The conditional bit error ratio (BER), depending on the synchronization error $\varepsilon$, is given by

$$P_c(\varepsilon) = Q \left( \frac{\rho(\varepsilon)}{\sqrt{E_p \sigma_i}} \right),$$  \hfill (44)

where we have assumed that the observation period equals a symbol period such that $N = 1$, $Q(x) = \int_x^\infty \exp(-t^2/2)/\sqrt{2\pi} dt$, and $\rho(\varepsilon) = \int_0^T \omega(t) \omega(t - \varepsilon) dt$.

Recall that the best theoretically achievable $\varepsilon$ is Gaussian distributed with zero mean and variance equivalent to the CRLB (denoted by $\sigma^2_\varepsilon$). In the best case, $\sigma^2_\varepsilon = \sigma^2_0/N \int_T \omega^2(t - \tau) dt$ from (9) is the smallest. Averaging $P_c(\varepsilon)$ over $\varepsilon$,
we get the mean BER

$$P_e = \mathbb{E}\left[ P_e(e_t) \right]$$

$$= \int_{-\infty}^{+\infty} \frac{1}{\sqrt{2\pi}\sigma_e}Q \left( \sqrt{\frac{\rho^2(e_t)}{E_p\sigma_0}} \right) \exp \left( -\frac{e_t^2}{2\sigma_e^2} \right) de_t. \quad (45)$$

Statistically, this is the best achievable performance under certain SNR. This equation can be evaluated numerically by Monte Carlo simulation which requires high computational complexity. Alternatively, we invoke the Hermite-Gaussian quadrature [34], and $P_e$ can be accurately approximated by

$$P_e \approx \frac{1}{\sqrt{\pi}} \sum_{n=1}^{N_h} H_{x_n} Q \left( \frac{\rho \sqrt{2\sigma_c x_n}}{\sqrt{E_c} \sigma_0} \right), \quad (46)$$

where $N_h$ is the order of the Hermite polynomial $H_{N_h} (\cdot)$, and $x_n$ and $H_{x_n}$ are the zeros (abscissas) and weight factors of $N_h$-order Hermite polynomial, respectively. These values are tabulated in many mathematical handbooks (e.g., [35]). In experiments, we find that 16 coefficients ($N_h = 16$) are enough to generate accurate approximation results.

Further define a variable $\eta$ as the degrading ratio between $P_e$ and $P_e(0) = Q(\sqrt{\gamma_t})$, which is the BER in the case of perfect synchronization. We show the values of $\eta$ for different monocycles in Section 6 to compare the synchronization error robustness of monocycles.

6. NUMERICAL RESULTS

Since for multipath channels, the CRLBs depend on the TH codes and detailed fading channel models, we only show numerical results on the CRLBs in pure AWGN channels in this paper.

In Figures 1, 2, and 3, the CRLBs for different monocycles in the case of known data $b$ are demonstrated. Since in practice, a transmitted monocycle is usually the truncated portion of a whole pulse $w(t; n, t_p)$, this effect of truncation is considered by varying the actual width of pulse in (10).

From Figure 1, we can see that CRLBs are inversely proportional to the symbol SNR and the observation period $N_T$. The relationship between CRLBs for monocycles with different $n$ order coincides with the analytical results in (12). This can be further observed in Figure 2, which also depicts the effect of truncated pulses on CRLB. The CRLBs change little even when the truncated portion narrows to 1.6 $t_p$ (symmetric with respect to $t = 0$). However, with the width of truncated pulse decreasing further, the CRLBs become orderless. Figure 3 shows the effect of $t_p$ on the CRLBs and is a direct verification of (13).
Figure 2: CRLB versus n-order for monocycles with \( t_p = 1 \) nanoseconds; different lines correspond to different widths of truncation.

Figure 4 demonstrates the influence of synchronization error on the performance of receivers. It is plotted from (46) using Hermite-Gaussian approximation. The influence is notable when the observation window in the stage of synchronization has small width \((NT_s)\) and weakens with \(N\) increasing (CRLBs decreasing). The figure also indicates that synchronization errors of different monocycles have very close influence on BER, although the data in experiments shows that the influence of monocycles with larger \(n\) is a little worse when \(\text{SNR} \ y_s\) is small and changes toward the opposite with \(\text{SNR}\) increasing.

7. CONCLUSIONS

We have derived the Cramér-Rao lower bounds (CRLBs) for the synchronization of UWB signals for both single-pulse systems and time-hopping systems in AWGN and multipath channels. Insights are given on the relationship between CRLBs for different Gaussian monocycles. The CRLBs in AWGN channels are discussed in both cases of known and unknown transmitted data, exist. An oversight in the lower-SNR approximation method [18] is highlighted, and a possible solution is provided by tightly locating the range of SNR \(y_s\). The CRLBs in multipath channels are studied for three scenarios depending on the way multipath interference is treated in a practical synchronizer, where multipath interference contributes as an increase of noise variance or multiple synchronization parameters. It is found that larger number of multipaths implies higher CRLBs and inferior performance of synchronizers, and multipath interference on CRLBs cannot be eliminated completely except in very limited cases.

The least influence of synchronization error on the performance of receivers is quantified. The influence is notable when observation window \((NT_s)\) in the stage of synchronization is small, and weakens with \(N\) increasing (CRLBs decreasing). Synchronization errors of different monocycles have very close influence on BER.

APPENDICES

A. APPROXIMATION OF \(\ln(\cosh(y(\tau)/\sigma_0^2))\)

Here we show how to approximate \(\ln(\cosh(y(\tau)/\sigma_0^2))\) as low-order polynomials in a piecewise fashion and determine the corresponding range of symbol SNR \(y_s\).

From \(y(\tau) = \int_{\tau} r(t) \omega(t - \tau) \, dt\), \(y(\tau)\) has Gaussian distribution \(\mathcal{N}(E_1, \gamma, \sigma_0^2)\), where \(|E_1| = E_p\) in the case of perfect synchronization, otherwise \(|E_1| < E_p\). As the estimate is usually clustered tightly around the true value in our case, and \(E_1\) changes smoothly for UWB monocycles, we assume that \(|E_1| = E_p\) (this can also be obtained from the assumption of unbiased estimation of \(\tau\)). Then \(y(\tau)/\sigma_0^2\) is also Gaussian distributed \(\mathcal{N}(\gamma, \gamma)\) or \(\mathcal{N}(-\gamma, \gamma)\). For Gaussian distribution, we know that when the distance between a sample and the mean is larger than about \(2.6\sqrt{\gamma}\), the probability of appearance can be assumed to be zero. Let the interval of interest be \([x_1 \leq y(\tau)/\sigma_0^2 \leq x_2]\); to ensure that most of the samples are in this interval, \(y_s\) should satisfy the following equations:

\[
-2.6\sqrt{\gamma} + y_s \geq |x_1|,
\]
\[
2.6\sqrt{\gamma} + y_s \leq |x_2|,
\]
\[
|x_1| + 5.2\sqrt{\gamma} \leq |x_2|.
\]

(A.1)
Briefly, two guidelines for determining the interval \([x_1, x_2]\) are (1) to ensure that this variable space be larger than the sampling space for a specific polynomial and SNR \(\gamma_s\), and cover the range of \(\gamma_s\) as widely as possible; (2) although two intervals can overlap, each interval should be fully covered by a single polynomial.

Considering the waveform of \(\ln(\cosh(x))\), from \(x = 0\) to a small \(x_2\), it has a very different shape with other segments and has to be approximated separately by a polynomial. This implies that there is only one interval covering the segment containing the point zero. For this interval, only \(x_2\) needs to be determined since \(\ln(\cosh(x))\) is an even function, and the distributions \(\mathcal{N}(\gamma_s, \gamma_s)\) and \(\mathcal{N}(-\gamma_s, \gamma_s)\) are symmetric with respect to 0. For the same reason, it is enough to consider the positive value of \(x_1\) and \(x_2\) for other segments hereafter. Note that \(\gamma_s\) should be at least 6.76 for \(x_1 > 0\), this implies that \(x_2 > 13.52\).

A well-known fact is that \(\ln(\cosh(x))\) can be accurately approximated by \(x^2/2\) when \(|x| \ll 1\), and by \(|x| - 0.69\) when \(|x| \gg 1\). But this simple scheme is not good enough to be realistic. For example, for a value \(x_2\) as large as 0.5, the approximation error is already 0.005, while the corresponding maximum SNR \(\gamma_s\) is only 0.0324 = \(-15\) dB which is of little interest in practice.

To summarize the description above, we find that a three-segment approximation is a good choice. Although the construction of these approximations is not unique, they can be represented as a general 2-order polynomial function \(f(x)\), which leads to a general CRLB expression as shown in (21).

**B. Derivation of \(E_{r_t|r|}\)**

First we derive the autocorrelation of \(r(t)\) which will be used in subsequent calculation:

\[
E_{r_t} \left[ r(t_1) r(t_2) \right] = E_{r_t} \left[ (s(t_1 - \tau) + n(t_1))(s(t_2 - \tau) + n(t_2)) \right] + \sigma_0^2 \delta(t_1 - t_2),
\]

where in the last equality, we utilize the assumption that signal and noise are mutually independent and \(n(t)\) is AWGN.
with zero mean and covariance $\sigma_r^2 \delta(t_1 - t_2)$. Note that the expectation with respect to $p(r|\tau)$ is equivalent to average over the data $b$ and noise $n(t)$. Recalling that the convolution between $r(t)$ and $\omega(t)$ in $y(\tau)$ is only within one symbol period $T_s$, in the case of ISI-free, we have

$$E_{\tau|r}[s(t_1 - \tau)s(t_2 - \tau)] = \omega(t_1 - \tau)\omega(t_2 - \tau),$$

$$E_{\tau|r}[r(t_1)r(t_2)] = \sigma_r^2 \delta(t_1 - t_2) + \omega(t_1 - \tau)\omega(t_2 - \tau). \quad \text{(B.2)}$$

Then expectations on $y(\tau)$ can be calculated as

$$E_{\tau|y}[y(\tau)] = \int_{T_s} E_{\tau|r}[r(t_1)] \omega(t_1 - \tau)\omega(t_2 - \tau) d\tau_1 d\tau_2$$

$$= \left[ \int_{T_s} \omega^2(t) d\tau + \sigma_r^2 \right] \cdot \int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau$$

$$= (\gamma_s + 1)\sigma_r^2 \int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau,$$

$$E_{\tau|y}[\gamma^2(\tau)] = \int_{T_s} E_{\tau|r}[r(t)] \omega^2(t) d\tau$$

$$= \int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau,$$

$$E_{\tau|y}[\gamma^2(\tau)] = \int_{T_s} E_{\tau|r}[r(t_1)r(t_2)] \omega(t_1 - \tau)\omega(t_2 - \tau) d\tau_1 d\tau_2$$

$$= \sigma_r^2 \int_{T_s} \omega^2(t - \tau) d\tau + \left[ \int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau \right]^2. \quad \text{(B.3)}$$

Assume that the energy of a pulse outside $T_s$ is negligible, these results can be further simplified due to

$$\int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau = 0,$$

$$\int_{T_s} \omega(t - \tau)\omega(t - \tau) d\tau = -\int_{T_s} \omega(t - \tau)d(\omega(t - \tau))$$

$$= -\omega(t - \tau)\omega(t - \tau)|_{T_s} - \int_{T_s} \omega^2(t - \tau) d\tau$$

$$= -\int_{T_s} \omega^2(t - \tau) d\tau.$$

According to the linear relationship between $x$ and $y(\tau)$, the expectations in terms of $x$ are straightforward.

### C. GAUSSIAN APPROXIMATION OF MULTIPATH INTERFERENCE

The key assumption we make is, for each multipath with index $\ell \neq m$, $\phi(t)_{\ell \neq 0}$ is identically independently distributed with mean $m_\phi$ and variance $\sigma_\phi^2$. As the number of multipaths $L$ in a dense UWB channel is very large, we invoke the central limit theorem so that every sample variable of $n_a(t)$ is Gaussian distributed with

$$\text{mean } m_n = \sum_{\ell = 1, \ell \neq m}^{L} a_\phi m_\phi, \quad \text{(C.1)}$$

$$\text{variance } \sigma_n^2 = \sum_{\ell = 1, \ell \neq m}^{L} a_\phi^2 \sigma_\phi^2.$$

The distribution of $\phi(t)_{\ell \neq 0}$ and the values of $m_\phi$ and $\sigma_\phi^2$ can be determined according to a general model describing each sample and probability in detail or some specifically chosen TH codes and multipath delays.

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Jian Zhang received the B.S. degree from Xi’an Jiaotong University, China, in 1996, the M.S. degree from Nanjing University of Posts and Telecommunications, China, in 1999, and the Ph.D. degree from The Australian National University, Australia, in 2004, all in telecommunications engineering. From 1999 to 2001, he was employed with Zhongxing Telecom Corporation (ZTE), China, as a Research and Development Engineer. His current research interests are in the areas of adaptive signal processing and wireless communications, with emphasis on ultra-wideband systems.

Rodney A. Kennedy received the B.E. (with honors) degree in electrical engineering from the University of New South Wales, Australia, in 1982, the M.E. degree in digital control theory from the University of Newcastle, Australia, in 1986, and the Ph.D. degree from the Department of Systems Engineering, The Australian National University, Canberra, in 1988. He has been the Head of the Department of Telecommunications Engineering, Research School of Information Sciences and Engineering, The Australian National University, since 1994. His research interests are in the fields of digital and wireless communications, digital signal processing, and acoustical signal processing. He is currently an interim Program Leader for the Wireless Signal Processing Program, part of the National ICT Australia (NICTA) Research Centre. He has shared responsibility in supervising about 25 research students and has coauthored approximately 180 papers. He is on a number of editorial boards for international journals and monograph series.
Thushara D. Abhayapala was born in Colombo, Sri Lanka, on March 1, 1967. He received the B.E. degree with honors in interdisciplinary systems engineering from the Faculty of Engineering and Information Technology, The Australian National University (ANU), Canberra, in 1994. He received the Ph.D. degree in telecommunications from the Research School of Information Sciences and Engineering, ANU, in 1999. From 1995 to 1997, he was employed as a Research Engineer at the Arthur C. Clarke Centre for Modern Technologies, Moratuwa, Sri Lanka. From December 1999 to July 2002, he worked as a Research Fellow at the Department of Telecommunications Engineering, and as a Lecturer at the Faculty of Engineering and Information Technology, both at ANU. Since July 2002, he has been a Fellow at the Research School of Information Sciences and Engineering, ANU. Currently he also holds an affiliation with the Wireless Signal Processing Program, Canberra, Research Lab, National ICT Australia (NICTA), Canberra, as a Senior Researcher. His research interests are in the areas of space-time signal processing for wireless communication systems, spatio-temporal channel modeling, MIMO capacity analysis, space-time receiver design, equalization, UWB systems, array signal processing, broadband beamforming, and acoustic signal processing. Dr. Abhayapala is currently an Associate Editor for the EURASIP Journal on Wireless Communications and Networking.