Performance analysis of impulse-radio ultra-wideband techniques for low-rate communications

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**PERFORMANCE ANALYSIS OF
IMPULSE-RADIO ULTRA-WIDEBAND
TECHNIQUES FOR LOW-RATE
COMMUNICATIONS**

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Summary

Impulse-radio (IR) ultra-wideband (UWB) has drawn much attention as a promising technology for low-data-rate applications such as ranging, identification and low-rate communication in sensor networks. In this research work, we focus on the communication perspective of the IR-UWB applications.

To begin with, we investigate various UWB transmission and reception schemes. With the criteria of low complexity, low power consumption and reliable data transmission, we select two systems for detailed study. They are transmitted reference (TR) signaling with auto-correlation receiver (AcR) and on-off keying (OOK) transmission with energy detection receiver.

After studying the systems' characteristics, we observe that the system performance suffers mainly from two types of interference. They are narrowband interference (NBI) from existing in-band wireless services, and multiple access interference (MAI). TR systems have another source of interference which is inter-pulse interference (IPI) from the overlapping of reference and data waveforms after passing through the highly dispersive propagation channel. Various suppression schemes have been investigated for all these types of interference. For NBI mitigation, notch filtering is adopted in this work. IPI can be mitigated by applying the statistic averaging concept. Our study shows that with these two techniques,

the overall performance of TR systems can be improved significantly. Theoretical analysis is provided to evaluate the lower-bound performance for the single-user case. It shows that the system performance is able to approach that under the additive-white-Gaussian-noise (AWGN) environment.

We also observe that OOK can be seen as a special case of TR when the delay between reference and amplitude-modulated data pulses is zero. Inspired by the AcR used for TR signals, we then propose a pseudo-coherent detector for OOK signals which is able to provide significant performance improvement as compared with the conventional energy detector. We also look into implementation issues of the proposed receiver. The high sampling rate required by the digital implementation motivates the investigation on techniques to reduce the sampling rate. We then develop a novel subsampling technique, namely bandpass downsampling (BPDS) technique from the standpoint of signal detection rather than signal reconstruction. Our proposed BPDS is capable of reducing the sampling rate to one tenth of the Nyquist rate with small performance degradation introduced to the receivers.

Our analysis work is subsequently extended to multiple access system or multi-user system under quasi-synchronous condition. Extended from OOK signaling for single user system, unipolar Walsh code (UWC) are adopted in our proposed multi-user system to distinguish different users. Over bipolar coding, unipolar coding has the advantages of providing better correlation template, and hence, lower detection error rate for the receiver. BPDS technique is also considered in our multi-user system for the ease of digital implementation with significant reduction in the required sampling rate. Theoretical analysis and numerical results support that our system can provide near-optimal performance under perfect timing acquisition

and satisfactory performance with certain amount of timing acquisition error.

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List of Acronyms

ABBREVIATIONS	FULL EXPRESSIONS
AcR	Auto-Correlation Receiver
ADC	Analog-to-Digital Convertor
AWGN	Additive White Gaussian Noise
BC	Block Coded
BER	Bit Error Rate
BO	Binary Othogonal
BOC	Binary Orthogonal Coded
BPDS	Bandpass Downsampling
BPF	Bandpass Filter
BSF	Band-Stop Filter
BUOGC	Balanced Unipolar Orthogonal Gold Code
BWC	Bipolar Walsh Code
CC	Cross-Correlation
CED	Conventional Energy Detector
CDMA	Code Division Multiple Access
DS	Direct Sequence
DTFT	Discrete-Time Fourier Transform

EGC	Equal Gain Combining
FCC	Federal Communications Commission
FIR	Finite Impulse Response
GLRT	Generalized Likelihood Ratio Test
GSM	Global System for Mobile Communication
HP	Hermite Polynomial
ICI	Inter-Chip Interference
IEEE	Institute of Electrical and Electronics Engineers
IFI	Inter Frame Interference
IIR	Infinite Impulse Response
IPI	Inter-Pulse Interference
IR	Impulse-Radio
ISI	Inter-Symbol Interference
MAI	Multiple Access Interference
MC	Multi-Carrier
MF	Matched Filter
MHP	Modified Hermite Orthogonal Pulse
MRC	Maximal-Ratio Combining
MSE	Mean Square Error
MUD	Multi-User Detection
MUTR	Multi-User Transmitted-Reference
NBI	Narrow Band Interference
NF	Notch Filter
NS	Nyquist Sampling
OCPAM	Orthogonal Coded Pulse Amplitude Modulation

OFDM	Orthogonal Frequency Division Multiplexing
OOK	On-Off Keying
OPMA	Orthogonal Pulse Multiple Access
PAM	Pulse Amplitude Modulation
PN	Pseudo Noise
PPM	Pulse Position Modulation
PS	Prolate Spheroidal
PSD	Power Spectral Density
PWAM	Pilot Waveform Assisted Modulation
SL	Square Law
SNR	Signal-to-Noise (Power) Ratio
TAE	Timing Acquisition Error
TCF	Template Cleaning Filter
TDL	Tapped Delay Line
TDMA	Time Division Multiple Access
TH	Time Hopping
TP	Template Purification
TR	Transmitted-Reference
UWC	Unipolar Walsh Code
UWB	Ultra-Wideband
WED	Weighted Energy Detection
WiMax	Worldwide Interoperability for Microwave Access
WLAN	Wireless Local Area Network
ZD	Zero Delay

List of Major Symbols

$a_{i,j}^k$	Amplitude of the i -th chip in the j -th symbol of the k -th user
$[a_0, a_1, a_2]$	Numerator coefficient set of a second-order digital notch filter
$[b_1, b_2]$	Denominator coefficient set of a second-order digital notch filter
B_{BPF}	Bandwidth of a bandpass filter
B_{NF}	Bandwidth of a notch filter
B_T	System bandwidth
c_i^k	The i -th chip in the unipolar Walsh codes for the k -th user
$\mathbf{C}_4$	The 4-by-4 unipolar Walsh matrix
$\mathbf{c}^k$	Unipolar Walsh code for the k -th user
$\hat{\mathbf{c}}^k$	Bipolar Walsh code for the k -th user
$d_i \in \{+1, -1\}$	Binary pulse-amplitude-modulation symbol
d_j^k	The j -th binary data symbol of the k -th user
E_c	Chip energy after bandpass downsampling
$E_{c,m}$	E_c of receiver unit m
$E_{c,in}$	In-phase chip energy
$E_{c,m,in}$	In-phase chip energy for receiver unit m
$E_{c,out}$	Out-of-phase chip energy
$E_{c,m,out}$	Out-of-phase chip energy for receiver unit m

E_p	Energy of the transmitted pulse $p(t)$
E_q	Energy of the received waveform $q(t)$
$E_x(f_s)$	Signal energy of $x[n]$ being sampled at f_s
E_x^{BPDS}	Signal energy of $x[n]$ with bandpass downsampling
E_x^{NS}	Signal energy of $x[n]$ with Nyquist sampling
$E_{\hat{\mathbf{n}}}$	Noise energy per chip in the correlation template
f_c	Centre frequency of a bandpass downsampling filter
f_s	Sampling frequency
f_{ss}	Sampling frequency used for digital notch filter design
Δf_v	Bandwidth of narrowband interference
f_v	Center frequency of narrowband interference
$H_{AcR}(f)$	Frequency response of an auto-correlation receiver
$H_{BSF}(f)$	Frequency response of a band stop filter
$H_{MF}(f)$	Frequency response of a matched filter
$h(t)$ and $h_{CH}(t)$	Channel impulse response
$h_{AcR}(t)$	Impulse response of an auto-correlation receiver
$h_{BPF}(t)$	Impulse response of a band pass filter
$h_{BSF}(t)$	Impulse response of a band stop filter
I_0	Spectrum amplitude constant of narrowband interference
I_j^k	Code-level interference to symbol j of user k due to t_e
$I_{j,\alpha}^k$	Value of I_j^k at interference level α
K	Number of users
k	Length of vector $\mathbf{q}$
L	Number of samples obtained within each chip duration
M	Number of parallel receiver units

N	Number of signal frame averaging used for template cleaning
N_s	Number of symbols used for template cleaning
N_0	Power spectral density of white Gaussian noise
N_c	Number of chips per symbol
N_Z	Residue noise component in the decision variable Z_i
$n(t)$	White Gaussian noise
$n_i(t)$	White noise in the i -th received symbol
$\hat{n}(t, N)$	Noise in $\hat{r}(t, N)$
$\tilde{n}_i(t)$	Residue white Gaussian noise in the i -th received data symbol
$\overline{\tilde{n}_i(t)}$	Mean value of $\tilde{n}_i(t)$
$n_Z(N)$	Total noise in decision variable
$n_{Z,1}(N)$	Product of noise in correlation template with clean data waveform
$n_{Z,2}(N)$	Product of clean correlation template with noise in data waveform
$n_{Z,3}(N)$	Product of noise in both correlation template and data waveform
$\hat{\mathbf{n}}(N)$	Noise in $\hat{\mathbf{q}}(N)$
$P_{e,best}$	Lower bound performance of bit-error-rate
$P_{e,N}$	N dependent probability error
$P_{e,N,M}^k$	N dependent probability error of user k with M receiver units
p_α^I	Probability of occurrence of each $I_{j,\alpha}^k$
$p(t)$	Transmitted ultra-wideband pulse
$Q(x)$	Q function
$q(t)$	Received ultra-wideband waveform
$\mathbf{q}$	Sample vector of received ultra-wideband waveform
$\hat{\mathbf{q}}(N)$	Estimation of $\mathbf{q}$ with N -symbol averaging
$R_j^{k,b}$	Interference from user b to user k in symbol j

$r(t)$	Received ultra-wideband signal
$r_i(t)$	Received i -th symbol
$\hat{r}(t, N)$	Output of the template purification filter
$\tilde{r}(t)$	Clean correlation template waveform
$S_{\hat{n}}(f)$	Power spectral density of N_Z
$S_v(f)$	Power spectral density of narrowband interference
$S_{\hat{v}}(f)$	Power spectral density of V_Z
$s(t)$	Transmitted ultra-Wideband signal
$s^k(t)$	Transmitted signal for the k -th user
T	Sampling interval
T_c	Chip duration
T_d	Time delay between reference and data pulses
T_f	Frame duration of the transmitted signal
T_p	Ultra-Wideband pulse duration
T_s	Symbol duration
T_{ss}	Sampling interval used for digital notch filter design
t_e	Timing acquisition error
t_i	Initial sampling instant
$t_{Nyquist}$	Nyquist sampling time interval
t_s	Sampling period of an analog to digital converter
$U_{i,j}$	Correlation output for the i -th chip of j -th symbol
V_Z	Residue interference component in the decision variable Z_i
$v(t)$	Narrowband interference
$\tilde{v}_i(t)$	Residue narrowband interference in the i -th received data symbol
$\overline{\tilde{v}_i(t)}$	Mean value of $\tilde{v}_i(t)$

W	Signal bandwidth
$X^+(f)$	Positive frequency spectrum of $x(t)$
$X^-(f)$	Negative frequency spectrum of $x(t)$
$X(\omega)$	Discrete-time Fourier transform of $x[n]$
$X_T(f)$	Poisson summation of Fourier transform of $x(t)$
$x(t)$	A continuous signal
$x[n]$	Sampled sequence of $x(t)$
Z_i	Decision variable of the i -th received data symbol
$Z_i(N d_i = 1)$	Decision variable for data symbol 1
$Z_{i,m}$	Decision variable of symbol i from receiver unit m
$Z_{j,m}^k$	Decision variable of symbol j from receiver unit m of user k
α	Levels of possible multi-user interference due to t_e
$\delta(f)$	Dirac delta function in frequency domain
η	Percentage loss of the symbol energy due to downsampling
Γ	The largest integer that satisfies $0 \leq t_i \leq t_s$
γ_j	Inter-pulse-interference at a sampling instant in j -th symbol
λ	Effective E_b/N_0 after downsampling
λ_m	Effective E_b/N_0 after downsampling of receiver unit m
σ_γ^2	Variance of γ_j
$\sigma_{N_Z}^2$	Variance of N_Z
σ_n^2	Variance of white Gaussian noise
$\sigma_{n(Z,3,N)}^2$	Variance of $n_{Z,3}(N)$
σ_v^2	Variance of narrowband interference
$\sigma_{V_Z}^2$	Variance of V_Z
μ_d	Mean value of the transmitted data

ω	Normalized radiant sampling frequency
$\Upsilon_{\mathbf{q}}$	Energy of the $\mathbf{q}$ vector

Chapter 1

Introduction

1.1 Motivation

Ultra-wideband (UWB) technology has been widely accepted as an emerging option for current and future communications [11]. Generally speaking, there are two types of UWB signals, namely, impulse-radio UWB (IR-UWB) and multicarrier UWB (MC-UWB). IR-UWB is characterized by extremely short-duration pulses with no carrier involved, whereas MC-UWB adopts multiple simultaneous carriers for transmission. One of the most popular scheme in MC-UWB is Orthogonal Frequency Division Multiplexing (OFDM) patented in the US in 1970 [12] and tremendous effort has been put on high-data-rate communication using OFDM technique. The other type of UWB signal, IR-UWB has been mainly studied for low-data-rate applications such as localization, ranging, and identification. The work in [13] has demonstrated that UWB-based 3-D localization is able to achieve 5 cm accuracy with potential applications in motion tracking for healthcare and entertainment industries. For these applications, some low-rate communication is

1.1 Motivation

desirable. However, they have stringent requirement on low system complexity and low power consumption. This requirement is very challenging to meet and also very important for battery operated devices.

In this work, we focus on developing low complexity IR-UWB communication system. The important design issues for wireless communication systems will apply to this work also, which can be summarized as: (i) interference suppression, (ii) signal detection scheme, and (iii) application to multi-user scenario.

1.1.1 Coexistence with In-Band Wireless Services

The US Federal Communications Commission (FCC) has assigned a 7.5 GHz band to UWB applications from 3.1 GHz to 10.6 GHz [14]. Some wireless services have been operating in this band, such as IEEE 802.11a WLAN and IEEE 802.16 WiMax. The stringent emission mask of UWB minimizes the interference from UWB to those existing wireless services. However, these relatively narrowband services form narrowband interference (NBI) to UWB systems and they have severe impact on UWB systems since they have much higher power spectral density (PSD). Figure 1.1 illustrates the basic principle of NBI. There is a need to investigate NBI mitigation techniques so that the UWB system performance can be improved.

1.1.2 Baseband Signal Processing

Since IR-UWB is fundamentally a baseband technique, the detection of IR-UWB signals can be seen as carrier-less baseband signal processing. The extremely short-duration pulses will experience highly dispersive propagation channel during

1.1 Motivation

transmission. The dense-multipath fading channel will make the detection difficult. To collect information from all the multipath components, RAKE receiver is optimal theoretically when the separation between two adjacent multipath components is at least the pulse duration. However, the complexity of RAKE receiver could be very high since a large number of fingers are needed and the performance greatly relies on accurate information of received pulse shape and channel impulse response, which could be very hard to obtain without complex estimation. IR-UWB has drawn a great amount of attention because of its potential to have low-complexity transmitters and receivers. The RAKE receiver, hence, may not be a proper choice to fully deliver the appealing features of IR-UWB.

Low-complexity systems are desired for low-rate UWB applications such as ranging, localization, identification and low-rate communications. Two of such systems are on-off keying (OOK) transmission with energy detection based non-coherent receiver and transmitted-reference (TR) signaling with autocorrelation-based differential receiver. The conventional non-coherent receiver and the dif-

Figure 1.1: Comparison of UWB and NBI spectra.

1.2 Objectives

ferential receiver provide poor bit-error-rate (BER) performance due to the high noise floor in the decision statistics. The huge performance gap between these low-complexity receivers and the RAKE receiver motivates us to study these systems and look for new or improved detection schemes to balance performance and complexity.

1.1.3 Multiple Access

To support multiple users in any communication system, multiple access techniques are essential. For IR-UWB systems, Time Division Multiple Access (TDMA) and Code Division Multiple Access (CDMA) techniques are considered as proper candidates. To apply TDMA, the duty cycle of the transmitted pulses must be very low, especially for IR-UWB systems of large channel delay spread. In other words, the transmission rate must be very low. When many users are using the system, multiple access interference (MAI) will occur when the duty cycle is not small enough to accommodate so many users. To avoid MAI, TDMA also requires synchronization among all users. For CDMA, code design is the key issue to ensure low cross correlation among different users for both synchronous and asynchronous transmission. It is necessary to study the MAI and to explore techniques or coding schemes to increase the system performance.

1.2 Objectives

This work aims to look into various issues faced by IR-UWB communications and thereafter to develop a multi-user IR-UWB communication system for low-data-rate applications with low system complexity and satisfactory performance.

Theoretical analysis on the performance of such low-complexity IR-UWB system will be provided and numerical results will be obtained to support theoretical derivations.

1.3 Major Contributions of the Thesis

There are three major contributions in this work. Firstly, we have studied and proposed filter designs for the suppression of NBI and provided the analysis framework for the receiver performance. Analog filter design and the corresponding digital finite-impulse-response (FIR) and infinite-impulse-response (IIR) versions are investigated. The receiver performance is compared across different filter designs and the performance dependence on filter design parameters is also studied. The considered filter design parameters include the center frequency, bandwidth, and the degree of interference suppression.

Secondly, through the investigation on base band signal processing, we have proposed a pseudo-coherent detection scheme for OOK and pulse position modulation (PPM) signals. The proposed scheme significantly removes noise contribution in the conventional energy detection receiver. Therefore, it is able to deliver a superior BER performance. To enable low-complexity implementation of the proposed scheme, a novel subsampling technique, namely, bandpass downsampling (BPDS) technique has been proposed. BPDS is developed with emphasis on signal detection rather than signal reconstruction. With BPDS, the proposed scheme is still able to deliver a near optimal performance. At the same time, the sampling rate can be reduced to one-tenth of the Nyquist rate. We also discovered that for the subsampling BPDS technique, the initial sampling instant is critical to optimize

1.4 Organization of the Thesis

the overall performance. Optimization on BPDS has been investigated based on this observation.

Last but not least, we have proposed a multiple-access IR-UWB communication system with BPDS and the pseudo-coherent detection scheme. The system design parameters including coding requirement have been discussed. Particularly, to minimize the estimation error in correlation template, balanced codes with equal number of 0s and 1s need to be used. System performance has been analyzed under quasi-synchronous scenario and it is well supported by numerical results. By considering quasi-synchronous scenario, we have investigated the effect of timing acquisition error to our system performance.

1.4 Organization of the Thesis

The thesis is organized according to our contributions. A literature review on UWB communication is provided in Chapter 2. Following that, we focus on our proposed schemes and results. In Chapter 3, we propose receiver designs with NBI suppression capability for our UWB communication system. The performance of these proposed receivers is studied with respect to different design parameters. Chapter 4 discloses our contribution in baseband signal processing. The relationship between TR and OOK has been discovered and a new detection scheme for OOK and even PPM signals is proposed based on our receiver design for TR signaling. Further more, a novel sampling technique BPDS is introduced in details. Their performance analysis supported by numerical results is given as well. After all the above study under single-user scenario, we move on to multi-user system in Chapter 5. A multiple-access system adopting unipolar coding is proposed with

1.4 Organization of the Thesis

our proposed receiver design discussed in the previous chapter. The performance of our multiple-access system is analyzed under quasi-synchronous condition. Finally, Chapter 6 concludes and discusses on future work.

Chapter 2

Literature Review

UWB, as a promising technology to support both high data rate, short-distance applications and low data rate, longer-distance applications, has become an attractive option for current and future wireless applications. Tremendous research efforts can be found in the areas of UWB communications and localizations, attributed to the emphasis on low power, low interference and low regulation. Some books can serve as good sources of reviews on fundamentals of UWB technologies. Such books include [11] by Nekoogar and [1] by Reed. Both provide a great introduction to UWB communications. In addition, the book by Benedetto, et al. [15] gives a comprehensive overview on UWB communication systems. In this chapter, we will give a more focused literature review on specific topics closely related to the motivation of our work on IR-UWB communications.

2.1 Coexistence Issues

In February 2002, the US Federal Communications Commission (FCC) allocated a 7.5 GHz band for UWB communications from 3.1 GHz to 10.6 GHz [14]. The FCC spectrum mask for UWB is shown in Fig. 2.1. If the entire band of 7.5 GHz is optimally used, the maximum power available to a transmitter is reported as approximately 0.5 mW [15]. The limited power should make UWB signals appear noise-like to existing in-band wireless services. Some studies on UWB interference to Bluetooth and Global System for Mobile Communication (GSM) systems have been carried out in [16] with a conclusion that UWB source gives interference very similar to Gaussian noise.

Figure 2.1: FCC spectrum mask for UWB communications in [1].

However, to UWB systems, the existing in-band services become a major inter-

2.1 Coexistence Issues

ference source in the form of NBI. Coexistence of UWB and other in-band wireless services becomes an issue. The problem of NBI suppression is not new to spread spectrum communications and has been extensively studied throughout the years such as the works in [17–24]. Different filtering techniques have been explored in those works to estimate the key parameters of NBI which subsequently are used for NBI suppression. These filtering techniques developed for spread spectrum communications can be great candidates to solve the NBI problem in UWB communications. For low-complexity filtering techniques, notch filtering is a great candidate and has been discussed in [21, 22] for spread spectrum communications. In the context of UWB communications, the work in [25] also suggested notch filtering techniques to handle the interference from WLAN. The work in [2] provided a thorough study on narrowband interference suppression in time-hopping IR-UWB communications. A transversal-type notch filter was applied to reject NBI and a closed-form expression of bit-error probability was derived in [2]. The notch filter is applied to every finger of the RAKE receiver in order to form the modified RAKE receiver used in [2]. The structure of the r -th finger of the modified RAKE receiver is shown in Fig. 2.2. The r -th finger aims to capture signal information of r -th multipath.

The receiver in Fig. 2.2 shows the notch filter structure. It is a Wiener filter used to predict and notch out the NBI. The selective maximal combination technique is used to combine the outputs from fingers with highest power for decision making. However, this proposed structure has very high complexity and is very hard to implement.

The work in [3] proposed an adaptive NBI suppression for a IR-UWB energy detector. The NBI centre frequency is estimated iteratively by comparing the

Figure 2.2: The r -th finger of the receiver in [2].

2.1 Coexistence Issues

outputs of the energy detector till the minimized output is achieved. Notch filtering is also applied to mitigate the NBI with the assumption that the notching bandwidth is no less than the NBI bandwidth. The proposed receiver with NBI mitigation capability is shown in Fig. 2.3. The corresponding bit-error-rate (BER) performance under CM3 environment of the IEEE 802.15.4a standard [26] is provided in Fig. 2.4. It demonstrates that a great performance improvement can be achieved by applying a notch filter to the energy detector for NBI mitigation. This low-complexity and low-consumption design of energy detector combined with adaptive notch filtering is a great candidate for the IR-UWB applications in Body Area Networks.

Figure 2.3: Energy detector with narrowband interference suppression in [3].

A master piece of work on NBI suppression by Milstein [27] for spread spectrum communications can be used as a guideline for filtering design to mitigate NBI in UWB communications. It summarized and compared different filtering techniques for NBI suppression in spread-spectrum systems. Three design criteria have been presented in [27] and the author concluded that for a strong interferer, the performance would converge to the same results under any of the three design criteria. The three design criteria are: (1) Whitening the entire received signal, (2) whitening the noise and interference only, and (3) placing a infinitely deep notch at the frequency location of the interfering tone. For UWB systems, the

2.1 Coexistence Issues

potential NBI will be much stronger than the UWB signals. Hence, any of these three criteria is applicable for the filter design.

Beside techniques to mitigate the NBI afterwards such as [28] using diversity selection of multiple-antennas and [29] adopting digital delay filters, other potential solutions for NBI problem could be NBI avoidance or prevention. Such techniques include multi-banding techniques and pulse shape design discussed in [30–32]. They are capable of reducing the power spectral density of the transmitted UWB signal in the NBI bands. In the next section, we are going to explore more on the pulse shape design for IR-UWB communications. It is useful not only for the NBI avoidance but also for the overall system performance, signal modulation schemes and multiple-access schemes. Hence, we are going to discuss these aspects in a separate section.

Figure 2.4: Energy detector performance in [3].

2.2 Pulse Shape Design for Impulse-Radio UWB

For UWB systems, different kinds of pulse shapes have been proposed in the literature, such as Gaussian [33], Rayleigh [34], Laplacian [1], cubic [35], and modified Hermitian monocycles [36]. The main objective is to obtain a nearly flat frequency spectrum over the UWB band and avoid any DC component. By far the most popular pulse shapes used are the Gaussian pulse and its derivatives. The Gaussian pulse is represented as [1]:

$$p(t) = \frac{1}{\sqrt{2\pi}\sigma^2} e^{-(t-\mu)^2/(2\sigma^2)} \quad (2.1)$$

where σ is the standard deviation of the Gaussian pulse in seconds, and μ is the location in time for the midpoint of the Gaussian pulse in seconds and the pulse width is defined as $\tau_p = 2\pi\sigma$ in [1].

More pulse shapes can be generated by a sort of high-pass filtering of this Gaussian pulse which is similar to taking the derivative of (2.1). Figure 2.5 illustrates some Gaussian pulses. The time axis is not specified because it depends on the values assumed for the parameters mentioned above. The pulses shown in (a), (b) and (c) of Fig 2.5 may not be useful for commercial UWB systems since all of them are not located in the UWB band under FCC regulations. The higher-order derivatives of Gaussian pulse, for example, the 6-th derivative shown in Fig. 2.5 (d), are proper candidates which fulfill the FCC regulations.

The work by Hu and Beaulieu [4] established a solid review on the performance of different pulse shapes that can fulfill FCC spectral mask, such as higher order derivatives of Gaussian pulses, prolate spheroidal (PS) function-based pulses, and frequency shifted modified Hermite polynomial (HP)-based pulses. The per-

Figure 2.5: Examples of Gaussian pulses for UWB communications.

2.2 Pulse Shape Design for Impulse-Radio UWB

formance examined is the averaged bit error rate of time-hopping UWB system under multiple-access scenario. The work has shown that higher-order Gaussian pulses are able to provide great performance consistently with relatively easy circuit implementation. One of the performance curve is included in Fig. 2.6 for illustration, where N_s in the figure represents the number of frames per symbol. The figure supports that the sixth-order Gaussian monocycle is a great candidate for UWB communications.

Figure 2.6: Performance comparison of different pulse shapes in TH-BPSK UWB system in [4].

Beaulieu and Hu went on to investigate other potential candidates for UWB pulse shape in [5]. The work proposed the truncated sinc pulse and suggested that it is in some senses a possible best pulse shape for multiple access UWB

2.2 Pulse Shape Design for Impulse-Radio UWB

communications [5]. Figure 2.7 from [5] illustrates the power spectral densities of the truncated sinc pulse compared with other pulses including the 6th-order Gaussian monocycle and $p_4^e(t)$, a cosine function time windowed by 4 convolved rectangular windows to meet the FCC mask. If properly designed, the truncated sinc pulse is able to achieve the maximum allowable pulse energy under FCC mask in the range of 3.1 GHz to 10.6 GHz, and hence, maximize the signal to interference and noise ratio (SINR).

Figure 2.7: Power spectral densities of the truncated sinc pulse and other pulses together with FCC spectrum mask in [5].

The family of modified Hermite pulses, although any single one of them may not be the best option for multiple-access UWB communications suggested in [4], the combined family can become a potential candidate for M-ary pulse shape modu-

lation [37] and orthogonal pulse multiple access [36].

Pulse shape design is the starting point to build an IR-UWB communication system. Moving forward, we are going to review works on the system level. As our objective is to build a low-complexity IR-UWB communication system, we are going to review the major low-complexity systems.

2.3 Low-Complexity Systems for IR-UWB Communications

For low-complexity realization of IR-UWB communication systems, there are two popular schemes. They are transmitted reference (TR) scheme and unipolar-pulsed scheme such as on-off keying (OOK). Systems under both schemes are closely related to each other. In this Section, we are going to review these two systems in details.

2.3.1 Transmitted Reference Systems

Proposed in 1964 to tackle the difficulties in signal detection caused by random or unknown channels, the TR technique [38] has become very popular for UWB systems. An auto-correlation receiver (AcR) with TR signaling offers a simple receiver structure with the capability of obtaining real-time channel information.

In a TR system, a pair of unmodulated and modulated signals is transmitted. The former serves as the reference and the later serves as the data signal. The reference signal is employed to demodulate the data signal at the receiver. The reference signal can be treated as a channel sounding signal which provides a

2.3 Low-Complexity Systems for IR-UWB Communications

rough estimate of the channel. The AcR used in TR systems is able to capture the entire signal energy for a slowly varying channel and it is relatively robust to synchronization problems [39].

Two modulation schemes, namely, pulse amplitude modulation (PAM) and pulse position modulation (PPM) are commonly applied to TR signals. The transmitted signal for a single user case can be generalized as

$$s(t) = \sqrt{E_p} \sum_i [p(t - iT_f) + a_i p(t - iT_f - T_d - b_i \delta)] \quad (2.2)$$

where $p(t)$ is the transmitted pulse shape, E_p is the pulse energy, T_f is the frame duration, T_d is the delay between the data pulse and the reference pulse, $a_i = \pm 1$ is the PAM bit, $b_i = \{0, 1\}$ is the information to be transmitted, and δ is the unit time delay for PPM. It shows that PAM and PPM can be applied separately or in a combined manner to TR signals. The receiver for TR signals is AcR, which delays the reference signal and correlates it with the data signal to get the decision statistics. Figure 2.8 shows the basic concept of an AcR.

Figure 2.8: AcR for TR signaling.

Equation (2.2) shows the definition for conventional TR signals. A lot of variations have been proposed to fully explore the potential of the TR schemes. A delay-hopped TR communication system has been proposed in [39] where the de-

2.3 Low-Complexity Systems for IR-UWB Communications

lay between the reference and data pulses is not fixed but with a hopping pattern. A general pilot waveform assisted modulation (PWAM) scheme was introduced by Giannakis, et al, in [40], which includes TR as a special case. The system parameters were derived to minimize the channel's mean square error (MSE) and to maximize the average capacity [40]. The circumstances under which the UWB autocorrelation-TR system is optimal were also analyzed [40]. It shows that when the burst size is equal to 2 symbols, one symbol needs to be reference symbol and the other is data symbol and the transmitted power needs to be split evenly between these two symbols for optimal reception. This resulting PWAM system turns out to be equivalent to the conventional TR autocorrelation system which was discussed in [41] in the context of UWB systems.

The work by Quek and Win in [42] has provided an analytical framework based on the sampling expansion approach for TR-UWB communication systems in dense multipath channels. It showed the signal-to-noise power ratio (SNR) penalty associated with TR signaling with AcR as compared to all-Rake and partial-Rake receivers. After that work, they also provided unified analysis of TR-UWB systems in the presence of NBI in [43]. Note that there was no NBI suppression scheme proposed to improve the system performance in [43].

Reference [44] provides the optimal and suboptimal receivers for TR-UWB systems in a single user multipath environment. The optimal receiver was derived using a generalized likelihood ratio test (GLRT). The derivation was based on the perfect knowledge of the channel impulse response and the assumption of ideal synchronization and resolvable multipath. The conventional AcR was proved to be a sub-optimal receiver in [44]. However, the optimal GLRT receiver needs to estimate the arrival time of each multipath and to implement a large number of

2.3 Low-Complexity Systems for IR-UWB Communications

correlators [44]. With such high complexity, GLRT receiver is only able to provide a 1.1 dB improvement from AcR at $BER = 10^{-5}$. Hence the GLRT receiver is not suitable for practical implementation and it primarily serves as a theoretical benchmark.

All the TR systems described above have a common major assumption/requirement on T_d , which is the delay between the reference and data pulses, that T_d must be no shorter than the maximum delay of the propagation channel. This is to avoid any inter-pulse interference (IPI) between the reference and data pulses. Conventional AcR is not able to handle IPI and there is huge performance degradation when IPI occurs. This requirement on T_d is one of the fundamental system weaknesses which limits the maximum transmission rate of TR system to half of that under other signaling schemes. To increase transmission rate, IPI is inevitable and the receiver needs to be modified.

The work in [45] proposed a modified AcR structure in digital implementation to tackle the IPI issue in TR-UWB communication systems. The proposed receiver was based on statistical averaging concept to recover the reference and data pulses from IPI. A pair of reference and data pulses is named a doublet. Taking PAM signals as an example, the average value of large number of PAM bits will approach zero. By averaging over a large number of received doublets, it is possible to clean the reference pulse from the interference due to the closely placed data pulse. Then the data pulses can be recovered from the IPI by subtracting the reference pulses. Note that the noise contained in the data pulses is not removed from this process. Correlation between the recovered data and reference pulses is carried out and decision is made based on the correlation output. It showed a 5 dB improvement at $BER = 10^{-4}$ compared with the conventional AcR.

2.3 Low-Complexity Systems for IR-UWB Communications

Another structure has been proposed in [6] and [46] to tackle the low data-rate issue of TR systems. A dual pulse transmission with conventional AcR was proposed. The dual pulse signaling means the reference and data pulses are transmitted with T_d just equal to the pulse duration T_p . Since the reference and data pulses are placed closely together, the transmission rate of the proposed system is nearly doubled compared to the conventional TR systems. The proposed system treats the reference plus data pulses as a new pulse shape for transmission. The new pulse shape is the dual pulse. After propagating through the channel, the dual pulse disperses as a whole unit and hence the dispersed reference and data pulses have no interference to each other. Figure 2.9 shows the dual pulse structure. Performance analysis for this proposed system was provided in [46]. The same group of researchers generalized the idea in [46] and proposed a pulse cluster concept for TR systems in [47]. Instead of single dual pulse, a group of identical dual pulses with uniform spacing is transmitted as a pulse cluster. The work in [46] and [47] has performance improvement compared to the conventional TR systems.

2.3.2 On-Off Keying Systems

Beside TR signaling with AcR, the other low-complexity system for UWB communications is OOK transmission with energy detection receiver. OOK with energy detector has the lowest complexity and the worst BER performance compared with other systems.

The detection of OOK signals can be either coherent or non-coherent. Under Gaussian noise environment, if the signal has a known form, even with unknown parameters, a matched filter or its correlator equivalent can be used as a coherent detector. When the signal has an unknown form, it is appropriate to consider

2.3 Low-Complexity Systems for IR-UWB Communications

an energy detector, which is non-coherent, to determine the presence of a signal. Urkowitz's work in [48] is a pioneer in energy detection of unknown deterministic signals. By applying the sampling expansion theory, Urkowitz represented the decision variables of the energy detector as the sum of squares of statistically independent Gaussian variates. It was shown that the decision variable is non-central chi-square distributed when signal is present and central chi-square distributed when signal is absent. This work greatly reduced the complexity in analyzing the energy detection scheme and it provided an analysis framework for the following research on energy detection related area.

Simon, et al. presented work on energy detection of unknown signals over fading channels [49]. This work is also based on sampling expansion approach used in [48]. This work quantified the improvement in detection capability when receiver

Figure 2.9: Dual pulse structure for transmission proposed in [6].

2.3 Low-Complexity Systems for IR-UWB Communications

diversity schemes are employed. It employed diversity combining schemes, such as square-law combining and square-law selection combining, to combine the output from the square-law device (square-and-integrate operation) for each multipath. Since this work is related to signal detection with fading channels, it is readily to be applied to the UWB systems with OOK transmission.

In the context of UWB communication, coherent detection for OOK signals needs perfect knowledge of the dispersive channel, which is rather difficult to obtain. Energy detection based non-coherent receiver is preferred because of its low complexity. Except the maximum channel delay, energy detector does not require any other information about the channel. A lot of research has been carried out to further improve the energy detection scheme. Optimization of energy detection receivers for UWB systems is presented in [7]. The receiver structure is shown in Fig. 2.10. The signal processing component of the receiver jointly estimates the optimal integration interval, the optimal detection threshold and the synchronization point in an adaptive manner. It is clearly shown in [7] that the expressions for bit error probabilities for bit 0 and bit 1 are quite different since they are of different distribution.

The weighted combining technique was applied to energy detection receiver in [8] proposed by Tian, et al. Two energy detection schemes were proposed for weighted combining: square law (SL) energy detection and cross-correlation (CC) energy detection. The latter one is similar to the AcR for TR signals. In the weighted energy detection (WED) receiver, each received symbol waveform is partitioned into M segments. A set of M integrators are employed to collect energy from these M segments. Then the outputs from these integrators are linearly combined with a weighting coefficient for each output. The optimal weighting

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coefficients are proportional to the SNR values of corresponding segments. Figure 2.11 illustrates the WED receiver structure.

Subsequently, Tian, et al, enhanced the SL based WED receiver with synchronization and channel estimation in [9]. Furthermore, feedback loops are proposed in the SL based WED to continuously improve the running estimation of channel information, the synchronization and the weighting coefficients for signal detection and demodulation. Figure 2.12 shows the improved structure of the WED.

Figure 2.10: UWB energy detection receiver with adaptive parameter estimation from [7].

Figure 2.11: Structure of weighted energy detection from [8].

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For practical communication systems, even error probability on detecting bit 0 and bit 1 is desirable. The decision threshold must be properly chosen to provide balanced error probability and the overall error probability must be low as well. However, the performance of the energy detection receiver is highly sensitive to the decision threshold. With a small shift from the optimal threshold, the performance degradation could be significant and the error probability for bit 0 and bit 1 could be unbalanced. This phenomenon has been discussed in [10]. Figure 2.13 shows the sensitivity of the conventional OOK to the decision threshold at certain SNR. The optimal threshold is very hard to obtain in practice. Although training symbols can be used for suboptimal threshold estimation [7], system throughput will decrease. The inevitable estimation error will degrade the performance.

Coding technique can be a solution to circumvent the non-trivial threshold determination required by conventional OOK schemes. A block-coded OOK (BC-OOK) modulation scheme was proposed in [10]. Each OOK symbol $s \in \{0, 1\}$ is encoded by a block code, $c_s = [c_{s,1}, c_{s,2}, \dots, c_{s,N_f}]^T$ with $c_{s,i} \in \{0, 1\}$ and the block codes for symbol 0, c_0 , and symbol 1, c_1 , are mutually orthogonal, i.e. $c_{0,i} + c_{1,i} = 1$, $\forall i$ and $\sum_{i=1}^{N_f} c_{s,i} = N_f/2$, $s = 0, 1$, assuming that there are N_f frames per symbol. This coding technique was adopted in [50] for PPM signals

Figure 2.12: Structure of the improved weighted energy detection from [9].

Figure 2.13: Sensitivity of conventional OOK energy detection to decision threshold from [10].

2.4 Multiple Access

with RAKE reception, whereas in [10] it was adopted to enable threshold-less energy detection for OOK signals. Exploiting the orthogonality between c_0 and c_1 , the detection for BC-OOK can avoid searching for the optimal threshold and the decision rule is designed as

$$\hat{s} = \begin{cases} 0, & z \geq 0 \\ 1, & z \leq 0 \end{cases}, \quad (2.3)$$

where $z = \frac{1}{N_0} (c_0^T - c_1^T) \underline{y}$ and $N_0/2$ is the power spectral density of the white Gaussian noise in the channel. Each element in vector $\underline{y}$ is the output of the square-law device corresponding to each frame duration.

Attributed to the symmetry, the error probability for symbol 0 and symbol 1 is balanced. It has been proved in [10] that the BC-OOK scheme is able to provide the same performance of conventional OOK with optimal decision threshold. However, the performance of all the works presented here for non-coherent OOK detection is still far worse than that of coherent detection, which suggests possibility of further improvement.

2.4 Multiple Access

Up to now, we have implicitly presumed a single-user scenario for UWB communication systems in this chapter. However, in practice, we must also think about the potential issues arising from the multi-user scenario. To separate multiple users effectively, traditional multiple access techniques can be considered. Particularly, we will emphasize on TDMA and CDMA schemes which are applicable to IR-UWB

2.4 Multiple Access

systems. We will also discuss a novel multiple access technique using orthogonal pulses, which is an interesting special case for UWB. Note that for signal detection of multi-user UWB systems, RAKE receivers are most widely used which are of high complexity and need accurate channel information. There is a need for low-complexity systems that can handle multiple access communications.

2.4.1 Time Division Multiple Access

In TDMA technique, a different time offset is applied for each user to avoid interference from other users. All the users in the system can use the same codes and the same bandwidth. TDMA is applicable to IR-UWB, since the pulses transmitted are of very short duration in time and they share exactly the same bandwidth. However, generally speaking, TDMA requires synchronized communication across all users, which gets difficult as the number of users increases. This technique would only be appropriate to be applied to the downlink (from a central base station to mobile users) [51].

2.4.2 Code Division Multiple Access

Another possible multiple access technique for IR-UWB is CDMA. Under this scheme, a unique spreading code will be allocated to each user. Two types of CDMA can be applied to IR-UWB, which are direct sequence (DS) and time hopping (TH). DS is normally paired with PAM and TH can be applied to both PAM and PPM. The work in [52] provided accurate performance analysis for DS-UWB and TH-UWB. Measured by the average probability of bit error, it was claimed in [52] that TH-PAM system outperforms the TH-PPM system for all

2.4 Multiple Access

values of SNR and in particular, the DS-UWB system is superior to the TH-UWB system in terms of the multiple access error rate performance, although TH-UWB was the original proposal for multiple access UWB systems.

The work in [53] provided an adaptive coding scheme to improve the performance of conventional TH-PPM UWB systems. An adaptive co-decoding scheme able to mitigate the MAI and to resist near-far effect was introduced in [53]. The basic idea is to jointly use Hamming distances (hard decisions) and Euclidean distances (soft decisions) of the candidate codewords to the received one. Since Hamming distances are in integer values, it is possible that multiple admissible codewords share the same Hamming distance from the received one. When this occurs, Euclidean distances are used instead of the Hamming ones. This proposed system is suitable in MAI-limited scenarios, where it outperforms the conventional TH-PPM UWB systems.

A flexible TH scheme combined with M -ary orthogonal PPM scheme and partial-RAKE receiver has been proposed in [54]. The TH slots and PPM slots have no fixed partition but rather flexible across the whole frame duration. This flexibility greatly increases the PPM modulation level and the multiple-access capacity. With this advantages, the flexible TH scheme is able to outperform both the conventional TH scheme and the conventional DS scheme [54].

2.4.3 Orthogonal Pulse Multiple Access

Orthogonal pulse multiple access (OPMA) scheme enables multi-user UWB systems based on mutually orthogonal pulse shape design. Since all the users in the system use orthogonal pulses, they are able to communicate simultaneously with very low interference to each other. The work in [36] developed the modified Her-

modified orthogonal pulse (MHP) system. MHP system is able to generate a family of pulses that are mutually orthogonal to each other. Since modified Hermite pulse shapes of lower order can be generated by integrating a pulse of higher order, a relatively low-complexity multiple pulse generator can be constructed. The OPMA technique is attractive because of its potential to provide nearly MAI-free UWB communication systems. However, this scheme limits the system capacity by the maximum number of orthogonal pulses that can be generated. Besides, due to the complex pulse generator, the system complexity of OPMA could be much higher compared with that of conventional multiple access schemes.

2.4.4 Synchronization and Timing Acquisition Errors

One practical issue concerning any communication system is the synchronization and timing acquisition errors. This problem can become even more pronounced under multi-user scenario. Such errors could result in higher MAI undesirably as orthogonality among users could be compromised. The work in [55] investigated the performance of a single user IR-UWB system with different synchronization procedures under IEEE 802.15.4a channel model as well as on a measured channel. The results shows that for non-line-of-sight cases, synchronization errors could degrade the system performance by 10 *dB* at $\text{BER} = 10^{-4}$. With more users entering to the system, the performance degradation caused by such issues must be studied and quantified clearly to better identify the system limitations.

Chapter 3

Interference Suppression for Transmitted-Reference Impulse-Radio UWB Systems

This work is focused on IR-UWB and its low-complexity systems. TR-UWB is one of the suitable candidates. After studying the system characteristics, it has been found that there are two major types of interference in this system, namely, NBI and IPI. We propose a receiver with the capability of mitigating both of them.

It is observed that the major drawback for the TR scheme is the low transmission rate caused by the reference pulse transmitted for each symbol. To avoid interference between reference and data pulses, they must be separated by at least the channel delay length. If higher data rate is desired, IPI is inevitable. Some researchers have proposed a receiver with the capability of IPI mitigation [45]. The main idea is to average the signals over multiple symbol intervals. Statistically, this process can purify the reference pulses from both information-bearing pulses

3.1 Digital Receiver Design

and background noise.

For practical TR-UWB systems, the presence of NBI may cause severe performance degradation. The averaging technique cannot handle the NBI suppression because the average of NBI over many frames is not necessarily zero. Note that NBI is normally much stronger than the UWB signals, which makes the latter signals noise-like at the receiver side. Thus, NBI suppression techniques must be applied before the averaging process is performed. We can adopt the well-studied NBI suppression techniques for spread-spectrum systems to UWB systems. Particularly, spectral filtering techniques are of interest because they can achieve large improvement in performance. The tapped-delay-line (TDL) structure for filter design draws much attention since it is more practical for implementation. Three design criteria for spectral filtering have been presented in [27] and the author concluded that for a strong interferer, the performance would converge to the same results under any design criterion. For UWB systems, the potential NBI will be much stronger than the UWB signals. Hence, any of these three criteria is applicable for the filter design. The third design criterion in [27] suggests placing a notch filter (NF) at the interferer's frequency.

3.1 Digital Receiver Design

Figure 3.1 shows our proposed digital receiver structure for TR-UWB systems to mitigate both NBI and IPI by using notch filtering and mean-based averaging techniques, respectively. Note that the completely digital implementation requires the sampling rate to be at least the Nyquist rate for the UWB band which could reach up to 20 GHz. The transmitted PAM signal is represented as:

3.1 Digital Receiver Design

$$s(t) = \sqrt{E_p} \sum_i [p(t - iT_f) + d_i p(t - iT_f - T_d)] \quad (3.1)$$

where $p(t)$ is a signal pulse with duration T_p and unit energy, E_p is the desired pulse energy, T_f is the frame duration, T_d is the delay of the data pulse with respect to the reference pulse, and $d_i \in \{+1, -1\}$ is the binary PAM symbol.

Figure 3.1: Proposed receiver structure with a notch filter and an improved AcR where the reference pulses are purified by averaging.

$v(t)$ denotes the NBI encountered in the channel and is modeled as a partial-band Gaussian noise with zero mean and variance σ_v^2 . The PSD is given by

$$S_v(f) = \frac{I_0}{2} \left[\text{rect} \left(\frac{f - f_v}{\Delta f_v} \right) + \text{rect} \left(\frac{f + f_v}{\Delta f_v} \right) \right] \quad (3.2)$$

where I_0 is a constant, f_v is the NBI center frequency, Δf_v is the NBI bandwidth and $\text{rect}((f - a)/b)$ represents the rectangular function centered at $f = a$ with width b and unit amplitude. $n(t)$ represents the additive white Gaussian noise (AWGN) in the channel and it has variance of $\sigma_n^2 = N_0 B_T / 2$, where N_0 is the power spectral density of AWGN and B_T is the system bandwidth.

After passing through the multipath channel $h(t)$, the received signals are sam-

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pled and processed by a notch filter (NF) which is used to suppress the NBI. The simplest digital NF structure is a second-order finite impulse response (FIR) filter with only one zero in the frequency response that provides a notch at the center frequency of the NBI. The filter structure is shown in Fig. 3.2. Even with such a simple NF, the receiver performance can be drastically improved. The impulse response of the NF can be written as

$$h_{NF}(t) = \sum_{i=0}^2 a_i \times \delta(t - iT_{ss}) \quad (3.3)$$

where $T_{ss} = 1/f_{ss}$ is the sampling interval to convert the analog signals to digital signals. The coefficient set $[a_0, a_1, a_2]$ can be determined as follows [56]:

$$\begin{aligned} a_0 &= (2 + 2 \cos(4\pi f_J/f_{ss}))^{-1} \\ a_1 &= a_0 (\pm 2 \cos(4\pi f_J/f_{ss})) \\ a_2 &= a_0 \end{aligned} \quad (3.4)$$

where the $\pm$ sign for a_1 is chosen accordingly so that the filter frequency response at two normalized frequency points $\omega = 0$ and $\omega = \pi$ will not exceed the unity gain.

Figure 3.3 shows the structure of a second-order IIR NF. The pole placed near the zero gives us the control on the notching width. By assuming the notching width as B_{NF} at $-A$ dB the coefficient sets $[a_0, a_1, a_2]$ and $[b_1, b_2]$ can be calculated as follows [56]

Figure 3.2: Design of a 2^{nd} order FIR NF with only one zero located at the NBI centre frequency.

Figure 3.3: Design of a 2^{nd} order IIR NF with one zero and one pole located near the NBI centre frequency.

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$$\begin{aligned}
 a_0 &= \left(1 + \sqrt{\frac{1 - 10^{\frac{-A}{10}}}{10^{\frac{-A}{10}}}} \tan \left(\frac{\pi \times B_{NF}}{f_{ss}} \right) \right)^{-1} \\
 a_1 &= -2a_0 \cos(4\pi f_J / f_{ss}) \\
 a_2 &= a_0 \\
 b_1 &= a_1 \\
 b_2 &= 2a_0 - 1.
 \end{aligned} \tag{3.5}$$

Figure 3.4: Magnitude response of (a) 2nd order FIR-NF (b) 2nd order IIR-NF. The notching-width ratio between (a) and (b) is 2.5 at -25 dB.

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Figure 3.4 shows the magnitude responses for both FIR-NF and IIR-NF implementations for NBI mitigation. The NBI is assumed to be WLAN signal located at 5.2 GHz and with bandwidth equal to 16.6 MHz. Hence both NF implementations have the notch placed at 5.2 GHz and the IIR NF has $B_{NF} = 16.6$ MHz defined at -40 dB. It is observed that there is about 2.5 times difference in notching width between the two NF implementations at the same gain magnitude level of -25 dB. The FIR NF shown in Fig. 3.2 is not able to control the notching width and it needs more taps to do so. More complex structures can be developed for more accurate control on the notching width for both FIR NF and IIR NF.

After the NF, the signal is processed by a matched filter (MF) which is matched to the transmitted pulse shape. Then the signal will be processed by the template purification (TP) filter. The TP filter purifies the received reference pulses by continuously averaging the received frames based on the assumption of a quasi-static channel which remains stable during the averaging process. After averaging over N frames, the resulting pulse will be used as the correlation template. As N increases, the noise component in the template will be further reduced but with a longer system delay. The TP filter can be expressed as

$$H_{TP}(z) = \frac{1}{N} (1 + z^{-T_f f_{ss}} + z^{-2T_f f_{ss}} + \dots + z^{-NT_f f_{ss}}) \quad (3.6)$$

where T_f is the symbol duration as defined in (3.1) and f_{ss} is the sampling frequency of the filter. Then the data pulses are obtained by subtracting the purified reference pulses from the received signal train. After that, the AcR will perform correlation between the data pulse train and the reference pulse train. Decisions are made according to the polarity of the correlation outputs.

3.1 Digital Receiver Design

Note that the system will have a long delay when N is of a large value. To reduce the delay, the running average structure can be used. Each time only the current frame is averaged with the last output of the TP filter. This implementation reduces the delay to 1 frame duration only. However, the receiver performance is poor at the beginning because the reference pulses are still quite noisy. As more signals are received, the error probability of the receiver decreases. Figure 3.5 illustrates the TP filter with running average structure.

Figure 3.5: Template purification filter with running average.

The performance of the proposed digital filter is shown in Fig. 3.6. The pulse chosen for the simulation is the 6th order derivative Gaussian pulse with a duration of 0.7 ns such that the spectrum is within the UWB band. The channel model adopted is the CM1 (Residential LOS) environment of the IEEE 802.15.4a model [26], whereby the type of fading is based on the modified Saleh-Valenzuela model. The delay spread is assumed to be 40 ns. System parameters are chosen as

Figure 3.6: Performance comparison for $SIR = -15$ dB. (a) comparison between the receiver with and without notch filtering (FIR and IIR NF) (b) receiver sensitivity to center frequency change (5% higher or lower) and notching width change (IIR to FIR width) of the NF.

3.2 Analog Receiver Design

$N = 200$, $T_f = 52$ ns and $T_d = 10$ ns. From Fig. 3.6, we can see that the system performance improves drastically for both FIR and IIR NFs. We also observe that the BER is not very sensitive to the centre frequency of the NF. With 5% variance, there is only less than 1 dB difference. In addition, the BER is not very sensitive to the notching width. There is about 2.5 times difference between FIR and IIR NF at the same gain level of -25 dB as shown in Fig. 3.4, but the BER difference is only about 1 dB. This is because most of the NBI has been removed and most of the signal energy is still preserved. With the averaging technique, reference pulses can be greatly purified. Note that the conventional TR scheme does not take care of the noise reduction in the signal template. The simulation curve for the case with IPI and NBI but without NF moves upward a little at higher E_b/N_0 . This could be due to the fact that the residue IPI increases when E_b increases. The residue IPI could become more dominant than the white noise N_0 .

3.2 Analog Receiver Design

For the analog implementation of the proposed receiver in Fig. 3.1, we just change the NF to an analog band-stop filter (BSF) designed based on Butterworth filter and change the summation in the AcR to integration. The received signal after MF can be written as

$$r(t) = \sum_i r_i(t) \quad (3.7)$$

where $r_i(t)$ is the signal at the output of the MF for $iT_f \leq t \leq (i+1)T_f$. It is defined as

3.2 Analog Receiver Design

$$r_i(t) = \sqrt{E_q} [q(t - iT_f) + d_i q(t - iT_f - T_d)] + \tilde{n}_i(t) + \tilde{v}_i(t) \quad (3.8)$$

where E_q is the energy of the received waveform $q(t)$. Note that $q(t) = \sqrt{E_p/E_q} * h_{CH}(t) * h_{BSF}(t) * p(-t)$ contains all the multipath channel information of $h_{CH}(t)$, $h_{BSF}(t)$ represents the impulse response of the BSF and $p(-t)$ represents the impulse response of the MF. The residual noise term $\tilde{n}_i(t)$ has zero mean. The term $\tilde{v}_i(t)$ represents the residual NBI after passing through the MF and has zero mean. Both $\tilde{n}_i(t)$ and $\tilde{v}_i(t)$ are uncorrelated Gaussian processes.

With averaging consecutively over N frames, the TP filter outputs a purified reference waveform as

$$\begin{aligned} \hat{r}(t, N) &= \frac{1}{N} \sum_{i=0}^{N-1} r_i[t - (N - 1 - i) T_f] \\ &= \sqrt{E_q} q(t) + \frac{1}{N} \sum_{i=0}^{N-1} \{ \sqrt{E_q} d_i q(t - T_d) + \tilde{n}_i[t - (N - 1 - i) T_f] \\ &\quad + \tilde{v}_i[t - (N - 1 - i) T_f] \}. \end{aligned} \quad (3.9)$$

This output waveform will be used as the correlation template for the AcR. As N increases, the noise component in the template waveform will be further reduced but with a longer system delay.

3.2.1 Performance Bound of the Analog Receiver

The lower bound of the BER performance is derived based on the assumption that N is approaching infinity. According to the strong law of large numbers [57], the output of the TP filter will approach the mean value of the input. Hence,

3.2 Analog Receiver Design

the output waveform will be the clean reference waveform without any noise or reference components which can be written as

$$\begin{aligned}\hat{q}(t) &= \lim_{N \rightarrow \infty} \hat{r}(t, N) \\ &= \sqrt{E_q} [q(t) + \mu_d q(t - T_d)] + \overline{\tilde{n}_i(t)} + \overline{\tilde{v}_i(t)} = \sqrt{E_q} q(t)\end{aligned}\quad (3.10)$$

where μ_d , $\overline{\tilde{n}_i(t)}$ and $\overline{\tilde{v}_i(t)}$ are the mean values of the transmitted data, the noise component and the NBI component, respectively. All these mean values are equal to zero.

Since the reference waveform is perfectly cleaned, the IPI component in the data waveform can be completely removed. The residual noise and NBI components are only present in the data waveform. Based on the central limit theorem [57], the probability density function of the residual noise and NBI terms after the AcR can be approximated as Gaussian distributed. The output of the AcR for the i -th symbol is

$$\begin{aligned}Z_i &= \int_{T_d + (i-1)T_f}^{T_d + iT_f} \hat{q}[t - (i-1)T_f - T_d] \{r_i(t) - \hat{q}[t - (i-1)T_f]\} dt \\ &= d_i E_q + N_Z + V_Z\end{aligned}\quad (3.11)$$

where N_Z and V_Z are the residual noise and NBI components after the AcR, respectively, and they can be approximated as Gaussian distributed.

The AcR output can be treated as an equivalent matched filter output for each frame duration where the equivalent matched filter of AcR has an impulse response

3.2 Analog Receiver Design

of

$$h_{AcR}(t) = \hat{q}(-t). \quad (3.12)$$

We then compute the variance of the noise and NBI terms in the frequency domain as

$$\sigma_{N_Z}^2 = \int_{-\infty}^{+\infty} S_{\hat{n}}(f) df \quad (3.13)$$

and

$$\sigma_{V_Z}^2 = \int_{-\infty}^{+\infty} S_{\hat{v}}(f) df \quad (3.14)$$

where $S_{\hat{n}}(f)$ and $S_{\hat{v}}(f)$ are the PSD expression of the corresponding noise and NBI terms, respectively. They can be expressed as

$$S_{\hat{n}}(f) = \frac{N_0}{2} |H_{BSF}(f)|^2 |H_{MF}(f)|^2 |H_{AcR}(f)|^2 \quad (3.15)$$

and

$$S_{\hat{v}}(f) = |H_{BSF}(f)|^2 |H_{MF}(f)|^2 |H_{AcR}(f)|^2 S_v(f) \quad (3.16)$$

where $H_{BSF}(f)$, $H_{MF}(f)$, and $H_{AcR}(f)$ are the frequency responses of the BSF, the MF matched to the transmitted pulse shape and the equivalent matched filter of AcR, respectively. Note that we can treat the AcR as an equivalent matched filter and the magnitude of its frequency response is $|H_{BSF}(f)H_{MF}(f)H_{CH}(f)|$. Hence, we derive that $|H_{AcR}(f)|^2 = |H_{BSF}(f)H_{MF}(f)H_{CH}(f)|^2$, where $H_{CH}(f)$ is the transfer function of the propagation channel. Following that, the lower bound of BER performance can be written as [58]

3.2 Analog Receiver Design

$$P_{e,best} = Q \left(\sqrt{\frac{E_q^2}{\sigma_{N_Z}^2 + \sigma_{V_Z}^2}} \right) \quad (3.17)$$

where $Q(x) \equiv \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$ [57].

3.2.2 Selection of N Value

Next, we study the effect of finite value of N on the BER performance of the proposed receiver. Since the residual NBI after BSF is relatively small as compared to the IPI component and the residual noise, we will only study the effect of N value on the noise and IPI components. By applying the central limit theorem [57], we can calculate N from the perspective of noise reduction at each sampling instant, which is given by

$$P \left\{ -X \frac{\sqrt{N}}{\sigma_{\tilde{n}}} \leq \frac{\sum_{j=1}^N \tilde{n}_j}{\sigma_{\tilde{n}} \sqrt{N}} \leq X \frac{\sqrt{N}}{\sigma_{\tilde{n}}} \right\} = 1 - 2Q \left(X \frac{\sqrt{N}}{\sigma_{\tilde{n}}} \right) = Y \quad (3.18)$$

where $\tilde{n}_j$ is the Gaussian noise at certain sampling instant from the j -th symbol period before the TP filter, $\sigma_{\tilde{n}}$ is the corresponding standard deviation, and Y is the confidence level required to have a noise level lower than X volts on average. Since the IPI in the reference waveform arises from the data waveform, the mean value of IPI is zero and the standard deviation at each sampling point is equal to the strength of the data waveform at that sampling instant. Hence, by applying the Chebyshev's inequality [57], we have

$$P \left\{ \left| \frac{\sum_{j=1}^N \gamma_j}{N} \right| \geq X \right\} \leq \frac{\sigma_\gamma^2}{NX^2} = 1 - Y \quad (3.19)$$

3.2 Analog Receiver Design

where X and Y are defined in a similar manner as in (3.18), γ_j is the IPI component at the sampling instant from the j -th symbol period and σ_γ^2 is the corresponding variance.

Note that the Chebyshev's inequality does not require any particular distribution for the random variables it operates on, and hence, it tends to over-estimate the N value. When the allowed N value is relatively small, Chebyshev's inequality gives a better estimation. When a large value of N is permitted, the averaged IPI component will approach Gaussian distribution. Under this assumption, we can provide better estimation by adopting (3.18) for estimation of N based on IPI reduction, which can be expressed as

$$P \left\{ -X \frac{\sqrt{N}}{\sigma_\gamma} \leq \frac{\sum_{j=1}^N \gamma_j}{\sigma_\gamma \sqrt{N}} \leq X \frac{\sqrt{N}}{\sigma_\gamma} \right\} = 1 - 2Q \left\{ X \frac{\sqrt{N}}{\sigma_\gamma} \right\} = Y \quad (3.20)$$

3.2.3 Numerical Results of the Analog Receiver

Figure 3.7 shows the effects of BSF on the BER performance with the signal-to-interference power ratio (SIR) equal to -15 dB. The performance of the generalized-likelihood-ratio-test (GLRT) receiver for TR-UWB systems [21] is also presented for comparison. Note that the GLRT receiver is optimal for the conventional TR-UWB systems without NBI and IPI. From Fig. 3.7, we observe that the BER performance of the proposed receiver is even better than that of the GLRT receiver under the interference-free environment at the expense of a longer system delay. The corresponding simulation curve labeled with "Clean Ref Wav ($N \rightarrow \infty$)" represents the lower bound of the BER performance of the proposed receiver under the given environment. The corresponding analytical curve is also presented in Fig.

Figure 3.7: Performance comparison among various receivers with $SIR = -15$ dB.

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3.7. The discrepancies between the simulation and analytical curves are mainly due to the Gaussian approximation adopted for the NBI components at the output of the AcR. The simulation and analytical curves corresponding to the ideal case are also presented for comparisons. The ideal case refers to (a) the reference and data pulses are well separated so that there is no IPI; (b) NBI is not present and hence BSF is not required; (c) N approaches infinity and hence the reference waveforms are noise-free. These two curves represent the best performance under the IEEE 802.15.4a fading channel with white noise. The gap between the lower-bound performance and ideal case is caused by the signal energy loss due to the BSF. It is observed that the performance of the proposed receiver is fairly close to the lower-bound performance and it is only around 2 dB away from that of the ideal case at $\text{BER} = 10^{-3}$. We can conclude that the proposed receiver is able to suppress both NBI and IPI effectively.

In Fig. 3.8, the BER performance of the proposed receiver versus the time delay T_d between reference and data pulses is presented. We observe that the BER results are insensitive to the change in T_d or the change in IPI level. That is because reference waveforms can be greatly purified from the IPI and the noise with the averaging technique. Note that the conventional TR scheme does not take care of the noise reduction in the signal template. Even when there is no delay between the data pulse and the reference pulse, i.e., $T_d = 0$ ns the receiver performance is only slightly worse than the case when $T_d = 40$ ns. In other words, the data rate of the TR scheme can be increased appreciably as the delay is decreased.

Figure 3.8: BER performance versus T_d with $N = 200$.

3.3 Chapter Summary

In this chapter, the main discussion is on interference suppression for impulse-radio UWB (IR-UWB) communication systems. Narrow band interference (NBI) is universal to all UWB systems. Suppression or mitigation of NBI can be achieved through proper filtering. Studies have shown that notch filtering is as good as other complex filtering schemes when the interference is strong and localized. As our main focus is on low-complexity realization of IR-UWB systems, notch filtering is adopted for NBI suppression. Various realizations of the notch filters have been proposed and their design criteria have been discussed.

Due to the same concern on low system complexity, transmitted-reference (TR) scheme has been studied as the signal transmission scheme. The interference arising from TR scheme itself is also investigated and suppression by statistical averaging is adopted. Combined with NBI suppression scheme, the whole single-user system performance is evaluated via numerical results.

Chapter 4

Baseband Signal Processing

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

Two popular transmission schemes for IR-UWB systems are on-off keying (OOK) and transmitted-reference (TR) signaling. The corresponding receiver structures are energy detection based non-coherent receiver and autocorrelation-based differential receiver. Coherent RAKE receivers are usually not considered for practical use since they are highly complex and the performance relies greatly on precise estimation of channel information and timing offset [59].

OOK transmission with an energy detection receiver provides the simplest system structure and it is the most practical one to implement UWB systems [60]. However, the bit-error-rate (BER) performance is very sensitive to the threshold setting at the output of the square-law device. Besides, improper threshold can also result in unbalanced error probability for data 1 and 0 since the detection statistics for data 1 and 0 are non-central chi-square and central chi-square distributed,

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

respectively. In general, pilot symbols are needed for better threshold setting. To avoid these issues, we can apply binary orthogonal coding to OOK signals [10] or, equivalently, we can adopt binary orthogonal pulse positioning modulation (BO-PPM) [60] for encoding. Due to the receiver characteristic of energy detection, the BER performance is much worse compared with that of coherent receivers. Improvement is desirable for reliable data transmission with lower transmission power.

It is observed that OOK signals can be treated as a special case of pulse-amplitude-modulated (PAM) TR signals when the delay between reference and data pulses becomes zero. We investigate the possibility of applying AcR's for the TR signals to the OOK signals in order to improve the overall system performance. As mentioned earlier, AcR's are conventionally used to detect TR signals. Since reference pulses are completely overlapped with data pulses for the zero-delay cases, we need to recover the correlation template from the received signal. Inspired by the work in [45] to suppress the interference between reference and data pulses, we adopt the statistical averaging technique to obtain a clean correlation template. However, it is not possible to know the exact number of transmitted pulses beforehand, resulting in additional uncertainty on the estimated correlation template in the presence of additive white Gaussian noise (AWGN). To avoid this difficulty, we propose a UWB system with binary-orthogonal-coded OOK (BOC-OOK) transmission. A pseudo-coherent detector which is improved from the conventional AcR is proposed to detect the received signal as a zero-delay PAM-TR (ZD-PAM-TR) signal.

4.1.1 System Model

Signaling Scheme

For the data transmission scheme, OOK is preferred due to its low complexity. To solve the problem of uneven error probability of detecting data 1 and 0, we propose BOC-OOK as the signaling scheme. BOC-OOK is basically the same as the BO-PPM. There are two time slots in every symbol duration. For each transmission, only one of them is occupied by the desired UWB pulse. When data 1 is transmitted, only the first time slot contains a UWB pulse and the second slot remains silent. In other words, the code word for data 1 is $[1 \ 0]$. For data 0, the code word is $[0 \ 1]$. The transmitted signal can be expressed as

$$s(t) = \sqrt{E_p} \sum_i [d_i p(t - iT_s) + (1 - d_i) p(t - iT_s - 0.5T_s)] \quad (4.1)$$

where E_p is the desired pulse energy, $p(t)$ is the signal pulse with duration T_p and normalized energy, T_s is the symbol duration, and $d_i \in \{1, 0\}$ is the i th binary OOK symbol. Note that $T_p \ll 0.5T_s$ so that there will be no inter-chip interference (ICI) after propagating through the channel.

By observing the relationship between OOK and TR signaling, we can conclude that OOK can be seen as a special case of the PAM-TR signaling with zero delay between reference and data pulses. Hence, the BOC-OOK signals can be treated as zero-delay orthogonal-coded PAM-TR (ZD-OCPAM-TR) signals with the data symbols coded as $[0.5 \ -0.5]$ and $[-0.5 \ 0.5]$ for data 1 and 0, respectively. Figure 4.1 illustrates the basic idea by using the code words.

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Figure 4.1: BOC-OOK and the equivalent ZD-OCPAM-TR schemes.

Pseudo-Coherent Detection Scheme

For OOK signals, an energy detection based non-coherent receiver is normally adopted because of its low complexity. However, due to the nature of energy detection, the BER performance is far worse than that of coherent detection. Inspired by the statistical averaging concept presented in [45], we propose a pseudo-coherent detector which is improved from the conventional AcR to detect OOK signals as ZD-TR signals so that the BER performance can be improved significantly. Assuming perfect synchronization, the performance of the proposed detection scheme is able to approach that under the AWGN environment at the expense of longer system delay. The proposed receiver structure is presented in Fig. 4.2.

Figure 4.2: Pseudo-coherent detector for BOC-OOK signals.

The received signal can be expressed as

$$r(t) = h(t) * s(t) + n(t) = \sum_i r_i(t) \quad (4.2)$$

where $h(t)$ is the channel impulse response, $n(t)$ is the AWGN component, and the symbol $*$ denotes the convolution operation. The received signal $r_i(t)$ for interval

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

$iT_s \leq t \leq (i+1)T_s$ is given by

$$r_i(t) = \sqrt{E_q} [d_i q(t - iT_s) + (1 - d_i) q(t - iT_s - 0.5T_s)] + n_i(t) \quad (4.3)$$

where E_q is defined as the energy of the received waveform $q(t)$, taking into account the channel dispersion. Note that $q(t) = p(t) * h(t)$ contains all the multipath channel information of $h(t)$. The AWGN noise term $n_i(t)$ has zero mean and variance $\sigma_n^2 = N_0 B_T$, where N_0 is the single-sided power spectral density of AWGN and B_T is the transmission bandwidth.

The template cleaning filter (TCF), or equivalently a TPF with half of the output magnitude, is to reconstruct the correlation template which is equivalently $0.5\sqrt{E_q}q(t)$ from the received signals by a two-step averaging process. The channel is assumed to be quasi-static which remains constant over the averaging process. This is a reasonable assumption since the IEEE 802.15.4a channel model has a coherence time in the order of 30 ms [61]. The first step is to average the received signal over the first and second time slots (each of $T_s/2$ seconds) within one symbol period to get a noisy correlation template. For the second step, these noisy templates over N symbols are averaged to get a much cleaner template. As N increases, the noise component in the template will be further reduced but with a longer system delay. The filter output can be expressed as

$$\hat{r}(t, N) = 0.5\sqrt{E_q}q(t) + \hat{n}(t, N) \quad (4.4)$$

where $\hat{n}(t, N)$ is the noise component in the estimated template with variance of $\sigma_n^2/(2N)$.

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

It is worthwhile noting that the maximum allowable N is limited by the channel coherence time. If the channel fading pattern is fast changing, the number of averaging N in the TCF will need to be reduced within the channel coherence time. This will result in higher noise residue in the output of TCF, and hence, a higher BER can be expected.

The equivalent OCPAM-TR data symbols are then obtained by subtracting the purified template in (4.4) from the received signals in *both* time slots of every symbol. Following that, the AcR will perform correlation between the equivalent OCPAM-TR data symbols and the template. Decisions are then made based on the polarity of the difference between the correlation outputs for the first and second time slots.

4.1.2 Performance Analysis of the Proposed Scheme

Lower-Bound BER Performance

We derive the lower-bound BER performance for the proposed UWB system. The lower bound is obtained when N approaches ∞ or the correlation template is perfectly recovered from noise. The clean template is given by

$$\tilde{r}(t) = \lim_{N \rightarrow \infty} \hat{r}(t, N) = 0.5\sqrt{E_q}q(t). \quad (4.5)$$

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

The decision variable for the i th symbol is

$$\begin{aligned}
 Z_i &= \int_{(i-1)T_s}^{(i-0.5)T_s} \tilde{r}[t - (i-1)T_s] [r_i(t) - \tilde{r}(t - (i-1)T_s)] dt \\
 &\quad - \int_{(i-0.5)T_s}^{iT_s} \tilde{r}[t - (i-0.5)T_s] [r_i(t) - \tilde{r}(t - (i-0.5)T_s)] dt \\
 &= \frac{1}{2}(2d_i - 1)E_q + n_Z
 \end{aligned} \tag{4.6}$$

where n_Z is the noise component before decision making. It is Gaussian distributed with zero mean and variance of

$$\sigma_{n_Z}^2 = \frac{1}{4}N_0E_q \int_0^{0.5T_s} q^2(t)dt = \frac{1}{4}N_0E_q. \tag{4.7}$$

Since the detection statistics for data 1 and 0 are symmetric, the lower bound of BER performance can be expressed as [58]

$$\begin{aligned}
 P_{e,best} &= P\{Z_i > 0 | d_i = 0, N \rightarrow \infty\} \\
 &= Q\left(\sqrt{\frac{E_q}{N_0}}\right)
 \end{aligned} \tag{4.8}$$

where $Q(x)$ is defined as $\frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$ [57].

Effect of N on the BER Performance

When N is finite, the estimated template contains a non-zero noise component. Therefore, the BER performance degrades as N decreases. The estimated template expressed in (4.4) will be used for correlation. The noise component $\hat{n}(t, N)$ can

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

be expressed as

$$\hat{n}(t, N) = \frac{1}{N} \sum_{j=1}^N \left(\frac{n_{1,j}(t) + n_{2,j}(t)}{2} \right) \quad (4.9)$$

where $n_{1,j}(t)$ and $n_{2,j}(t)$ represent the noise components in the first and second time slots of symbol j , respectively.

By considering the effect of N , we can express the decision variable for data 1 as

$$\begin{aligned} Z_i(N|d_i = 1) &= \int_0^{0.5T_s} \hat{r}(t, N) [\sqrt{E_q}q(t) + n_{1,i}(t) - n_{2,i}(t)] dt \\ &= \frac{1}{2}E_q + n_{Z,1}(N) + n_{Z,2}(N) + n_{Z,3}(N) \\ &= \frac{1}{2}E_q + n_Z(N) \end{aligned} \quad (4.10)$$

where $n_{Z,1}(N)$ and $n_{Z,2}(N)$ correspond to the product term of the clean data waveform with the noise in the estimated correlation template and the product term of the clean template waveform and the noise in the data waveform respectively. They are Gaussian random variables with zero mean and variance of $N_0E_q/(4N)$ and $N_0E_q/4$, respectively. $n_{Z,3}(N)$ corresponds to the product of both noise components in the template and data waveforms and it can be approximated as Gaussian distributed according to the central limit theorem [57]. To calculate its variance, we adopt the sampling theorem [58]. $n_{Z,3}(N)$ can be represented by $0.5T_s \times 2W = T_sW$ virtual samples, where W is the signal bandwidth. Hence,

$$n_{Z,3}(N) = \frac{1}{2W} \sum_{l=1}^{T_sW} [\hat{n}(N)_l (n_{1,i,l} - n_{2,i,l})] \quad (4.11)$$

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

with variance

$$\begin{aligned}\sigma_{n(Z,3,N)}^2 &= \frac{1}{(2W)^2} T_s W \left[\frac{\sigma_n^2}{2N} \times 2\sigma_n^2 \right] \\ &= \frac{T_s W}{4N} N_0^2.\end{aligned}\tag{4.12}$$

Since the mean values of the cross-term products are zero, the total variance of $n_Z(N)$ is the sum of the variance of its three components. The exact BER is then represented as

$$\begin{aligned}P_{e,N} &= P\{Z_i(N|d_i = 1) < 0\} \\ &= Q\left(\sqrt{\frac{E_q^2}{(\frac{1}{N} + 1)N_0 E_q + \frac{T_s W N_0^2}{N}}}\right).\end{aligned}\tag{4.13}$$

As $N \rightarrow \infty$, the BER values in (4.13) will approach the lower-bound BER values evaluated in (4.8).

4.1.3 Numerical Results and Discussions

For simulation, we adopt the 6th order derivative Gaussian pulse with duration of 0.7 ns which satisfies the spectrum mask set by Federal Communications Commission (FCC). The channel model adopted is the CM1 (Residential LOS) environment of the IEEE 802.15.4a standard [26], whereby the type of fading is based on the modified Saleh-Valenzuela model. We design the transmission rate such that $T_s/2$ is longer than the maximum channel delay, and hence, there is no inter-chip-interference in the received signal.

In Fig. 4.3, we compare the performance of the conventional energy detector (CED) with the proposed pseudo-coherent detector on detecting the BOC-OOK

4.1 Proposed Pseudo-Coherent Detection Scheme for OOK/PPM IR-UWB Signals

signals. It shows that by treating the BOC-OOK signals as ZD-OCPAM-TR signals and detecting them using the proposed receiver, the system performance improves greatly even when the number of signal symbols used for template cleaning is only $N = 5$. When N increases, the performance is approaching the lower-bound performance derived in Section 4.1.2. It is observed that when $N = 500$, the performance curve is quite close to the theoretical lower bound which converges to the AWGN baseline environment. The analytical BER curves for different values of N are obtained based on (4.13). They are very close to their respective simulation curves, which validates the Gaussian approximation used to the output noise component.

In Fig. 4.4, we compare the BER performance of the proposed receiver for detecting non-coded conventional OOK and the BOC-OOK signals. We have made some modifications to the proposed receiver to make it able to detect the conventional OOK signals. Firstly, the TCF reconstructs the correlation template by using one-step averaging process to average over N symbols, since in each symbol duration, there is only one time slot and during which either a positive pulse or a null is transmitted. Secondly, the correlation process corresponding to interval $0 \leq t \leq 0.5T_s$ in Fig. 4.2 is kept and its output is served as the decision variable, whereas the one corresponding to $0.5T_s < t \leq T_s$ is removed. The comparison is based on equal transmission rate and equal effective received E_b/N_0 . It is observed that the system with BOC-OOK constantly performs better than that with non-coded OOK. When the value of N is relatively small, significant improvement is achieved and the slope of the BER curve under BOC-OOK is much larger than that under non-coded OOK. That is mainly attributed to the balanced number of pulses sent for data 1 and 0, respectively, in the BOC-OOK scheme. The uneven number

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Figure 4.3: BER performance comparison between the proposed receivers and the conventional energy detector (CED).

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of pulses sent for non-coded OOK will introduce more uncertainty and more noisy component into the template estimation. When N increases, this uncertainty will reduce, and hence, the BER curves under different OOK schemes will converge. At $E_b/N_0 = 14 \text{ dB}$, the BER under the BOC-OOK scheme with $N = 50$ is almost as good as the BER under non-coded OOK scheme with $N = 100$. That means that BOC-OOK can provide similar system performance with a smaller system delay, which is desirable for implementation.

Figure 4.4: BER performance comparison between non-coded OOK and BOC-OOK.

4.2 Proposed Bandpass Downsampling Scheme

To implement the above proposed pseudo-coherent receiver in analog components, we need long analog delay lines for the TP filter. However, those long delay lines are very difficult to build in analog components. For digital implementation of the proposed receiver, delay lines are easy to realize but the sampling rate requirement is quite stringent at the receiver front end. The Nyquist rate of sampling for UWB signals can go up to 20 GHz which requires ADC of extremely high rate and high complexity. Since our objective is to reduce the system complexity and cost, we need to reduce the sampling rate. We propose the bandpass downsampling (BPDS) scheme which is able to reduce the sampling rate significantly with small degradation in BER performance of the receiver.

4.2.1 Theory of Bandpass Downsampling

In general, signal processing focuses on signal reconstruction or signal recovery after signal reception. The main purpose is to reveal the signal properties intended to be transmitted. These properties include but not limited to signal shape, amplitude, position, and spectrum. For that purpose, signal aliasing must be avoided. In previous work on sampling rate reduction using bandpass sampling theory for UWB systems [62], the emphasis was mainly on reconstruction of the actual UWB pulse shape in the lowest possible sampling rate to avoid aliasing. In the context of our work on UWB signal detection, the main concern is to determine any presence of UWB pulses in terms of energy rather than the actual UWB pulse shape. This discovery has enabled us to take a different perception on signal processing. Aliasing or distortion in signal shape is acceptable as long as it does not affect the

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detection on UWB pulse presence. Besides, aliasing is inevitable to further reduce the sampling rate below the bandpass sampling rate [63].

Our proposed BPDS scheme is different from the existing bandpass sampling scheme in a way that our scheme allows and utilizes certain pattern of aliasing. The BPDS consists of a bandpass filter (BPF) and an ADC with sampling rate equal to the bandwidth of the BPF. There is a requirement on the location and bandwidth of the BPF such that $2f_c/B_{BPF}$ is of an integer value, where f_c is the center frequency and B_{BPF} is the bandwidth of the BPF. BPDS is different from the bandpass sampling theory proposed in 1991 [63]. Bandpass sampling theory is applicable to bandpass signals and it tries to avoid any form of aliasing in the signal spectrum. The minimum sampling rate is then derived based on the location of the signal spectrum. Some research work has been done to apply the bandpass sampling theory to UWB systems [62] since UWB is basically a bandpass systems located from 3.1 GHz to 10.6 GHz. The BPDS allows a certain pattern of aliasing and utilizes it to jointly achieve low sampling rate and satisfactory performance. The BPF is located to extract part of the signal spectrum which contains the highest power spectral density (PSD). Then the bandwidth of the BPF is adjusted to meet the requirement described earlier. The sampling rate of the ADC is required such that the bandpass filtered signal spectrum will have full aliasing that the negative and positive spectra are folded and overlapped exactly with each other. Then the received signals are converted to relatively narrowband signals of bandwidth equal to the sampling frequency. B_{BPF} is the minimum required sampling rate for BPDS. By using a sampling rate lower than B_{BPF} , self aliasing will occur which means copies of the signal spectrum will interfere with the original spectrum. Self aliasing enhances noise floor and reduces signal

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power which causes the reduction in effective signal-to-noise power ratio (SNR). Sampling rates higher than B_{BPF} do not necessarily result in better performance. The sampling rate under BPDS should be chosen to avoid any partial aliasing when copies of the negative signal spectrum only partially overlap with the copies of the positive signal spectrum. Partial aliasing is also more destructive to the effective SNR compared to full aliasing.

BER Performance Bound of BPDS Scheme

In this part, we will derive the BER performance bound of an IR-UWB receiver adopting BPDS scheme and compare it with the optimal performance under Nyquist sampling. For easy reference, we name these two receivers as BPDS receiver and Nyquist receiver, respectively. Both receivers are assumed to have perfect knowledge on IR-UWB pulse shape and signal propagation channel. Matched filters are used in both receivers to maximize the SNR. By comparing the resultant SNR of both receivers is equivalent to comparing their BER performance, as the BER performance is calculated using $Q(x)$ function with x equivalent to the square root of SNR.

We will start from some signal processing theory first to establish the generalized equation to calculate the signal samples energy. Assuming $x[n]$ is the sampled sequence of a continuous signal $x(t)$, the Discrete-Time Fourier Transform (DTFT) can be written as:

$$X(\omega) = \sum_{n=-\infty}^{\infty} x[n]e^{-i\omega n}. \quad (4.14)$$

Another frequency function $X_T(f)$ is defined based on the Poisson summation

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formula [64] which is written as:

$$X_T(f) \stackrel{\text{def}}{=} \sum_{k=-\infty}^{\infty} X(f - kf_s) = T \sum_{n=-\infty}^{\infty} x(nT) e^{-i2\pi f T n}, \quad (4.15)$$

where

$$X(f) = \int_{-\infty}^{\infty} x(t) e^{-i2\pi f t} dt \quad (4.16)$$

is the continuous-time Fourier transform of $x(t)$. In the formulae above, T is the sampling interval, $f_s = 1/T$ is the sampling frequency, $\omega = 2\pi f T = 2\pi(f/f_s)$ is the normalized radian sampling frequency, f represents frequency, and t represents time.

As $x[n] \equiv x(nT)$, we can use the inverse transform of $X_T(f)/T$ to recover the discrete-time sequence:

$$x[n] \equiv x(nT) = \int_{-\frac{1}{2T}}^{\frac{1}{2T}} X_T(f) \cdot e^{i2\pi f n T} df. \quad (4.17)$$

From the above equations, we can derive the signal energy after being sampled at f_s as:

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$$\begin{aligned}
E_x(f_s) &= \sum_{n=-\infty}^{\infty} |x[n]|^2 \\
&= \sum_{n=-\infty}^{\infty} |x(nT)|^2 \\
&= \sum_{n=-\infty}^{\infty} x(nT) \overline{x(nT)} \\
&= \sum_{n=-\infty}^{\infty} \left[\int_{-\frac{1}{2T}}^{\frac{1}{2T}} X_T(f) \cdot e^{i2\pi f n T} df \int_{-\frac{1}{2T}}^{\frac{1}{2T}} \overline{X_T(f')} \cdot e^{-i2\pi f' n T} df' \right] \\
&= \sum_{n=-\infty}^{\infty} \left[\int_{-\frac{1}{2T}}^{\frac{1}{2T}} \int_{-\frac{1}{2T}}^{\frac{1}{2T}} X_T(f) \overline{X_T(f')} e^{-i2\pi n T(f'-f)} df' df \right] \\
&= \int_{-\frac{1}{2T}}^{\frac{1}{2T}} \int_{-\frac{1}{2T}}^{\frac{1}{2T}} X_T(f) \overline{X_T(f')} \left[\sum_{n=-\infty}^{\infty} e^{-i2\pi n T(f'-f)} \right] df' df \\
&= \frac{1}{T} \int_{-\frac{1}{2T}}^{\frac{1}{2T}} \int_{-\frac{1}{2T}}^{\frac{1}{2T}} X_T(f) \overline{X_T(f')} \left[\sum_{n=-\infty}^{\infty} \delta(f' - f - k f_s) \right] df' df \\
&= \frac{1}{T} \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} X_T(f) \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} \sum_{k=-\infty}^{\infty} \overline{X_T(f')} \delta(f' - f - k f_s) df' df \\
&= f_s \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} X_T(f) \overline{X_T(f)} df. \tag{4.18}
\end{aligned}$$

Note that based on (4.15), the item $\sum_{n=-\infty}^{\infty} e^{-i2\pi n T(f'-f)}$ in (4.18) can be calculated as

$$\begin{aligned}
\sum_{n=-\infty}^{\infty} e^{-i2\pi n T(f'-f)} &= \frac{1}{T} \cdot \left\{ T \sum_{n=-\infty}^{\infty} y[nT] e^{-i2\pi n T(f'-f)} \right\} \\
&= \frac{1}{T} \sum_{k=-\infty}^{\infty} Y(f' - f - k f_s) \\
&= \frac{1}{T} \sum_{k=-\infty}^{\infty} \delta(f' - f - k f_s) \tag{4.19}
\end{aligned}$$

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with $y[nT] = 1$ as the samples of an all one signal $y(t) = 1$ and the Fourier transform of $y(t)$, $\mathcal{F}\{y(t)\} = Y(f) = \delta(f)$.

With the above derivation on the generalized equation on samples energy, we will move on to derive the signal samples energy under BPDS and Nyquist sampling (NS) schemes, respectively.

Figure 4.5: Signal spectrum with bandpass downsampling.

Figure 4.5 illustrates an example of a signal frequency spectrum before sampling in green lines. The signal spectrum is double sided with component located at positive frequency band and negative frequency band represented by $X^+(f)$ and $X^-(f)$ respectively. The bandwidth of the signal is $f_c + B$. The dashed line represents one shifted copy of the signal spectrum after BPDS sampling.

For the scenario of BPDS, the two implementation requirements are as follows:

I. The required components are a BPF located at f_c with bandwidth equal to $2B$ and an analog-to-digital convertor (ADC).

II. To generate full aliasing, the positive frequency portion $X^+(f)$ must completely overlaps with the negative frequency portion $X^-(f)$ after sampling. The following two conditions must be satisfied:

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- i) Sampling rate $f_s = 2B$;
- ii) Centre frequency is at f_c with f_c/B equal to an integer value.

If the phase components of the signal have a relationship such that $\angle X^+(f + f_c) = \angle X^-(f - f_c)$, the signal energy at the BPDS output is maximized and optimal. The maximized signal energy can be calculated using the generalized sample energy equation 4.18 as follows:

$$\begin{aligned}
 E_x^{BPDS} &= 2B \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} X_T(f) \overline{X_T(f)} df \\
 &= 2B \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} |2X^-(f - f_c)|^2 df \\
 &= 8B \int_{-B}^B |X^-(f - f_c)|^2 df.
 \end{aligned} \tag{4.20}$$

For the scenario of NS, the signal is sampled at Nyquist rate of $f_s = 2(f_c + B)$. The signal energy after sampling is:

$$\begin{aligned}
 E_x^{NS} &= 2(f_c + B) \int_{-\frac{1}{2}f_s}^{\frac{1}{2}f_s} X_T(f) \overline{X_T(f)} df \\
 &= 2(f_c + B) \int_{-(f_c+B)}^{(f_c+B)} 2|X^-(f)|^2 df \\
 &= 4(f_c + B) \int_{-B}^{f_c} |X^-(f - f_c)|^2 df.
 \end{aligned} \tag{4.21}$$

Assuming that the channel is corrupted by AWGN with power spectral density of $N_0/2$, the SNR ratio after match filters between BPDS and NS receivers is as follows:

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$$\begin{aligned}
\frac{SNR_{BPDS}}{SNR_{NS}} &= \frac{E_x^{BPDS}/(2N_0B)}{E_x^{NS}/[N_0(f_c + B)]} \\
&= \frac{2B \int_{-B}^B |X^-(f - f_c)|^2 df}{(f_c + B) \int_{-B}^{f_c} |X^-(f - f_c)|^2 df} \cdot \frac{f_c + B}{2B} \\
&= \frac{\int_{-B}^B |X^-(f - f_c)|^2 df}{\int_{-B}^{f_c} |X^-(f - f_c)|^2 df} \\
&\leq 1, \forall f_c \geq B.
\end{aligned} \tag{4.22}$$

It shows that the SNR ratio between BPDS and NS is the energy ratio of the signal $x(t)$ with and without the BPF required by BPDS. If the signal $x(t)$ itself does not have any energy beyond the range of the BPF, the SNR ratio between BPDS and NS is at its maximum which is equal to 1.

With the above theoretical derivation, we conclude that the performance bound of the proposed BPDS technique under the assumption of $\angle X^+(f + f_c) = \angle X^-(f - f_c)$ is closely linked to the signal energy captured by the BPF. If the signal by nature is a bandpass signal with centre frequency at f_c and signal energy within $f_c \pm B$, then the resulted SNR of BPDS is equivalent to that of NS, provided that the BPF is able to capture all the energy.

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In this section, we are going to look at the application of BPDS to our proposed IR-UWB receiver with pseudo-coherent OOK detection presented in Fig. 4.2.

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Figure 4.6: Proposed pseudo-coherent OOK detector with BPDS.

Figure 4.6 shows the pseudo-coherent OOK detector with BPDS component. The proposed receiver is a hybrid structure that contains both analog and digital components. The BPDS component can be realized in analog circuits with BPFs and ADCs and the subsequent parts of the receiver can then be built in digital components.

4.3.1 Performance Analysis of the Combined Scheme

BER Performance with Single Receiver Unit

Firstly we consider the most straight forward realization of the proposed pseudo-coherent receiver with only one receiver unit as shown in Fig. 4.6. The signal for the i -th symbol after the BPDS component will be in discrete samples and it can be expressed in vector form

$$r_i = \sqrt{E_q} \left\{ d_i \begin{bmatrix} \mathbf{q} \\ \mathbf{0}_{(2L-k) \times 1} \end{bmatrix} + (1 - d_i) \begin{bmatrix} \mathbf{0}_{L \times 1} \\ \mathbf{q} \\ \mathbf{0}_{(L-k) \times 1} \end{bmatrix} \right\} + \mathbf{n}_i \quad (4.23)$$

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where $\mathbf{q}$ is the vector representation of the continuous signal after the BPF in the BPDS component $q(t) = \sqrt{E_p/E_q}p(t) * h_{CH}(t) * h_{BPF}(t)$. $L = T_f f_s$ is the number of the samples obtained within each frame duration T_f which is half of T_s , the symbol duration. f_s is the sampling rate. k is the length of $\mathbf{q}$.

The output of TP filter is the estimated correlation template and it can be expressed as

$$\hat{\mathbf{q}}(N) = 1/2\sqrt{E_q}\mathbf{q} + \hat{\mathbf{n}}(N) \quad (4.24)$$

where $\hat{\mathbf{n}}(N)$ is the noise component vector in the estimated template and each element in it has variance of $\sigma_n^2/(2N)$ and $\sigma_n^2 = N_0 B_{BPF}$ is the variance of the bandpass filtered AWGN. $\hat{\mathbf{n}}(N)$ can be expressed as

$$\hat{\mathbf{n}}(N) = \frac{1}{N} \sum_{j=1}^N \left(\frac{\mathbf{n}_{1,j} + \mathbf{n}_{2,j}}{2} \right) \quad (4.25)$$

where $\mathbf{n}_{1,j}$ and $\mathbf{n}_{2,j}$ represent the noise vectors in the first and second frames of the j -th symbol, respectively. Then we can express the decision variable for symbol 1 as

$$\begin{aligned} Z_i(N|d_i = 1) &= \hat{\mathbf{q}}(N)^T \left[\sqrt{E_q}\mathbf{q} + \mathbf{n}_{1,j} - \mathbf{n}_{2,j} \right] \\ &= \frac{1}{2}E_q \times \eta + n_{Z,1}(N) + n_{Z,2}(N) + n_{Z,3}(N) \\ &= \frac{1}{2}E_q \times \eta + n_Z(N) \end{aligned} \quad (4.26)$$

where $n_{Z,1}(N)$ and $n_{Z,2}(N)$ correspond to the product term of the clean data waveform with the reference noise and the product term of the clean reference

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waveform and data noise respectively. They are Gaussian random variables with zero mean and variance of $\sigma_{n_{Z,1}}^2 = N_0 E_q \times \eta / (4N)$ and $\sigma_{n_{Z,2}}^2 = N_0 E_q \times \eta / 4$, respectively. $\eta = E_q / E_p \times 100\%$ represents the percentage loss of the symbol energy due to downsampling. $n_{Z,3}(N)$ is the inner product term of both reference and data noise vectors with zero mean and variance equal to $\sigma_{n_{Z,3}}^2 = k \times \sigma_n^2 / 2N \times 2\sigma_n^2$.

Approximately, the total variance of $n_Z(N)$ is the sum of the variance of its three components. Since the decision variable for data 0 is symmetrically distributed to that of data 1, the BER can then be represented as

$$P_{e,N} = P \{Z_i(N) | d_i = 1 < 0\} \\ \approx Q \left(\sqrt{\frac{\lambda^2}{(\frac{1}{N} + 1) \lambda + \frac{k}{N}}} \right) \quad (4.27)$$

where $\lambda = \eta(B_T / B_{BPF}) \times (E_q / N_0)$ represents the effective E_b / N_0 after downsampling and B_T is the signal transmission bandwidth which is much larger than B_{BPF} .

We can observe that the BER performance in (4.27) depends on the parameter N . When N is a finite number, the correlation template will contain noise component. Whereas, when N approaches infinity, the correlation template can be reconstructed perfectly without noise and the BER performance approaches its lower bound. Therefore, the BER performance can be enhanced with larger N at the expense of longer system delay.

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BER Performance with Multiple Receiver Units

The above single-unit receiver recovers signals from a particular frequency band of the original transmitted IR-UWB signal. To enhance the frequency diversity and ultimately the total SNR of the decision statistics, we can implement multiple units to cover multiple frequency bands simultaneously. We assume that the receiver has M units and each unit has its own decision variable generated from different frequency bands. Our task then is to combine the M decision variables effectively. There are various diversity-combining schemes including equal-gain combining (EGC), selection combining (SC), and maximal-ratio combining (MRC) [58]. As our focus is on low implementation complexity, we adopt EGC as it does not need extra information or estimation on weightage for each receiver unit, and hence, less circuitry in the final receiver design. The BER performance of a receiver with M parallel units under EGC scheme can be represented as

$$P_{e,N,M} = P \left\{ \sum_{m=1}^M Z_{i,m}(N|d_i = 1) < 0 \right\} \\ \approx Q \left(\sqrt{\frac{(\sum_{m=1}^M \lambda_m)^2}{(\frac{1}{N} + 1) \sum_{m=1}^M \lambda_m + \frac{Mk}{N}}} \right) \quad (4.28)$$

where $Z_{i,m}$ represents the decision variable of symbol i from receiver unit m and λ_m represents the effective E_b/N_0 of receiver unit m .

4.3.2 Numerical Results and Discussions

We use the same channel model and pulse shape as in Section 4.1.3. All the performance is compared at $\text{BER} = 10^{-4}$. T_f is designed to be longer than the maximum channel delay assumed to be 40 ns to avoid inter-frame-interference

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(IFI) in the received BOC-OOK or equivalent BO-PPM signals. Hence, we set $T_f = 41$ ns and $T_s = 2 \times T_f = 82$ ns. For the proposed receiver with single unit, we locate the BPF at $f_c = 6$ GHz with $B_{BPF} = 2$ GHz to capture the signal spectrum containing the highest signal power. The sampling rate is set at $f_s = B_{BPF} = 2$ GHz, the smallest value that satisfies the requirement for BPDS. The number of symbols used for template cleaning is set at $N = 200$. Correspondingly, a system delay of 200×82 ns = 16.4μ s is introduced.

Figure 4.7: BER performance comparison between the proposed single-unit receiver and the conventional energy detector.

Figure 4.7 compares the performance of the proposed single-unit receivers with conventional energy detection (CED) receiver on detecting the BOC-OOK or

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equivalent BO-PPM signals. It shows that by treating the BOC-OOK signals as ZD-OC-PAM-TR signals and detecting them using the proposed receiver, the system performance improves significantly. It is observed that our proposed receiver is able to offer more than 9 dB improvement from the energy detection receiver at $\text{BER} = 10^{-4}$. Analytic curves are also provided which are well matched with the simulation ones. Beside the performance of the proposed receiver with BPDS, we also provide the theoretical curve for the proposed receiver operating at 20 GHz sampling rate without BPDS, which is roughly the Nyquist rate for UWB band. We observe that the proposed BPDS technique is able to reduce the sampling rate to one tenth of the Nyquist rate and yet provide performance with only 0.5 dB degradation at $\text{BER} = 10^{-4}$ compared with the performance under 20 GHz sampling rate.

Besides our proposed single-unit receiver, the BER performance comparison in Fig. 4.8 includes the proposed multiple-unit receiver, 3-unit receiver to be specific. The 3 receiver units are located at $f_{c1} = 4$ GHz, $f_{c2} = 6$ GHz, and $f_{c3} = 8$ GHz correspondingly with the same bandwidth of the BPFs at $B_{BPF} = 2$ GHz and the same sampling rate at $f_s = 2$ GHz. The 3-unit receiver is able to cover most of the UWB band from 3 GHz to 9 GHz, and hence, able to utilize most of the UWB signal energy for detection. This increase in frequency diversity brings a further benefit of 0.5 dB at $\text{BER} = 10^{-4}$ compared with the single-unit receiver when $N = 200$. Furthermore, we also observe that the BER performance of the proposed receivers is quite close to the all-RAKE receiver which is the optimal receiver structure with its BER equal to $Q(\sqrt{E_b/N_0})$ [58]. Figure 4.8 also confirms that our proposed BOC-OOK system outperforms the conventional TR system with PAM signaling [65,66] under the same data rate, SNR and propagation chan-

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Figure 4.8: BER performance comparison between all the proposed receivers and the conventional detectors.

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nel conditions in the expense of longer system delay. With the above observation in performance, we recommend the proposed single-unit receiver for practical implementation. It requires only a third of the circuitry compared to that of 3-unit receiver and it can still provide satisfactory performance with only 0.5 dB degradation from that of the 3-unit receiver.

In Fig. 4.8, we also include the performance curve of Averaged TR (ATR) receivers proposed in [41, 44] for PAM and PPM signaling respectively. Both ATR receiver designs assume ideal implementation is possible. The ATR outperforms conventional TR because of the clean correlation template obtained by statistical averaging over $N = 200$ consecutive reference waveforms. The ATR with PAM signaling has comparable performance to the BOC-OOK receiver operating at 20 GHz sampling rate without BPDS. Taking into account of practical implementation issue, our proposed system with only 2 GHz sampling rate is still able to provide satisfactory performance with only 0.5 dB difference in SNR. Considering that our proposed system has signaling scheme equivalent to PPM, we are able to improve the BER performance of PPM signaling to that of PAM signaling. It shows that our proposed 3-unit receiver outperforms the suggested ATR receivers with PPM signaling for about 1.5 dB at $\text{BER} = 10^{-4}$ as our system does not require an extra signal frame to transmit the reference pulse for each symbol.

In our proposed BPDS receivers, the sampling rate f_s is a key factor that influences the overall BER performance. Figure 4.9 illustrates such impact of different sampling rate to the BER performance of our proposed receiver. The performance curves are obtained from single-unit receiver with $N = 200$, $f_c = 6$ GHz, and $B_{BPF} = 2$ GHz. When $f_s < B_{BPF}$, self aliasing phenomena occurs and the effective SNR at the decision variable degrades significantly once f_s is less

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Figure 4.9: BER performance of our proposed receiver with single receiver unit and various f_s .

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than B_{BPF} . That also leads to consistent BER performance degradation once f_s is less than B_{BPF} . When f_s is only slightly greater than B_{BPF} , partial aliasing takes over. Although f_s has increased, it is not necessarily equivalent to better BER performance compared with the full aliasing case at $f_s = B_{BPF}$. Both destructive and constructive partial aliasing scenarios exist. Destructive partial aliasing occurs when one copy of the signal spectrum is aliased by more than one copy of the spectrum after sampling, e.g., when $f_s = B_{BPF}/0.9$. Otherwise, relatively constructive partial aliasing occurs. However, the relatively constructive partial aliasing can lead to inconsistent BER performance even with consistent increase in f_s . This is dominated by the actual signal energy collected during sampling process. It is worth mentioning that BPDS is still essentially a sub-sampling technique with sampling rate much lower than the Nyquist rate. Various levels of signal energy are collected with various f_s . From the frequency spectrum point of view, one of the reasons could be that the shifted copies of the original signal spectrum after sampling have amplitudes proportional to f_s [58]. In Fig. 4.9, we observe that some of the relatively constructive aliasing such as $f_s = B_{BPF}/0.8$ is able to capture more signal energy, and hence, performs better than that of full aliasing at $f_s = B_{BPF}$.

Besides f_s , there is another design parameter N that influences the BER performance of our proposed receivers. In Fig. 4.10, we compare the receiver BER performance with respect to various N values for both single-unit receiver and 3-unit receiver structures. We observe that even with a small N value such as $N = 5$, the proposed receivers are able to deliver a much improved performance compared with that of conventional energy detector. To the other extreme, when N approaches infinity, the BER performance approaches its theoretical lower bound.

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Figure 4.10: BER performance comparison of the proposed receiver with various N values.

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We observe that when $N = 200$, the BER performance is fairly close to the theoretical lower bound. Hence, we recommend $N = 200$ for the implementation of our proposed receivers for great performance under the condition that the system can tolerate a throughput delay of 200 symbols. If the allowed delay is not as long, $N = 50$ is recommended as it can reduce a lot of delay time and yet is able to provide a satisfactory performance. There is another observation worth mentioning that the multiple-unit receiver may not always outperform the single-unit receiver, especially when N is relatively small. Fig. 4.10 shows that when $N = 5$, the single-unit receiver actually outperforms the 3-unit receiver. This can be well explained through the theoretical performance analysis for M -unit receiver in (4.28). In the decision variable, the noise variance is jointly affected by N and the effective E_b/N_0 . When N is small, the noise product item Mk/N becomes the dominant factor of the BER. If M receiver units are adopted, the extra signal energy collected will be neutralized by the noise product item. As N increases, the dominant factor in BER will shift towards the effective E_b/N_0 . With more receiver units, more signal energy can be captured, and hence, the BER performance will be enhanced.

Although the proposed receivers are initially developed for BOC-OOK signals, they actually can be adopted to detect conventional non-coded OOK signals with two modifications. Firstly, the Template Purification Filter only averages across symbols rather than frames as there is only one frame per symbol in non-coded OOK signals. Secondly, the summation of sampled signals from $[1 \ T_f f_s]$ in Fig. 4.6 is kept and the corresponding output serves as the decision variable, whereas the summation branch corresponding to $[T_f f_s + 1 \ 2T_f f_s]$ is removed. The resultant receiver could have some estimation errors in the amplitude of the recon-

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Figure 4.11: BER performance comparison of the proposed receiver with BOC-OOK and conventional OOK transmissions.

4.3 Bandpass Downsampling Applied to the Proposed Detection Scheme

structured correlation template as the actual number of pulses received is unknown. This estimation error will contribute as an extra interference in addition to the residual white noise. If N is a small and odd number, the template estimation error will be exaggerated which leads to BER performance degradation.

The BER performance of the modified single-unit receiver with non-coded OOK signaling is presented in Fig. 4.11. It is also compared against the BER performance of unmodified single-unit receiver with BOC-OOK signaling, based on equal signal transmission rate and equal effective received E_b/N_0 . We observe that the BOC-OOK system consistently outperforms the non-coded OOK system. As discussed earlier, when N is a small and odd number such as $N = 5$, there is a large error in the estimated template amplitude for non-coded OOK due to unbalanced data 1 and data 0. There is no such issue in BOC-OOK thus we can observe a huge performance improvement by adopting BOC-OOK. As N increases, the template estimation error from non-coded OOK signals reduces, and hence, the BER performance curves under different OOK schemes converge eventually.

Before we proceed to the next section, we would like to highlight the advantages and constraints of the proposed receivers as well as the potential applications. The proposed receivers with BPDS is capable of noticeably reducing the required sampling rate to one tenth of the Nyquist rate. This significantly eases the practical implementation of the receivers in digital circuit components. Besides, the proposed receivers are able to achieve performance close to that of an all-RAKE receiver. However, the proposed receivers require a statistical averaging process which causes a relatively long system throughput delay in the order of tens of microseconds. To make the averaging process meaningful, the signal propagation channel must be quasi-static with coherent time longer than the system delay.

4.4 Improved Detection Scheme from Initial Sampling Instant

Furthermore, inter-frame interference (IFI) and inter-symbol interference (ISI) is not allowed in our propose receivers. This is different from the system proposed in [67] which focuses on the modeling and suppression of IFI and ISI for high-data-rate applications. With these constraints, our proposed receivers are suitable for low-data-rate applications that operate under quasi-static channel condition and can tolerate certain amount of system delay.

4.4 Improved Detection Scheme from Initial Sampling Instant

We have proposed and studied the pseudo-coherent receivers operating at sub-sampling BPDS rate for OOK or PPM signals in the context of IR-UWB communications. In this section of the work, we propose to further improve the receivers by investigating the impact of various initial sampling instants on the resultant SNR. With a properly allocated initial sampling instant in the BPDS component, we can enhance the SNR of the sampled signals and thereafter reduce detection error.

4.4.1 Observation in Initial Sampling Instant and Receiver Improvement

From (4.27) and (4.28), we observe that λ , the effective E_b/N_0 , is one of the dominant factors in the BER performance of the previously proposed receivers. To enhance the detection performance, λ must be as large as possible. We observe that the actual sampling instant has great impact on the effective signal energy

4.4 Improved Detection Scheme from Initial Sampling Instant

collected by the receiver. After we determine the values of f_c and B_{BPF} of the BPF and the sampling rate of the ADC inside the BPDS component according to the design requirement, the only variable related to λ is the initial sampling instant. Since our system is of uniform sampling, if an improper initial sampling instant is chosen, significant amount of signal energy may be lost during the sub-sampling process.

Figure 4.12 illustrates the noise-free samples obtained after the BPDS component with respect to different initial sampling instant. The BPF adopted has $B_{BPF} = 2$ GHz and $f_c = 4$ GHz. The pulse shape adopted is the 6th order derivative Gaussian pulse of 0.7 ns which satisfies the spectrum mask set by the Federal Communications Commission (FCC) and the channel model is CM1 (residential-line-of-sight) of IEEE 802.15.4a standard [26], whereby the type of fading is based on the modified Saleh-Valenzuela model. The maximum channel delay is assumed to be 40 ns. The top graph shows the signal waveform sampled at 20 GHz, which is the Nyquist sampling rate of the UWB signals, and the corresponding initial sampling instant is at $t_i = 0$ ns. This graph serves as the benchmark for the subsequent graphs obtained with sub-sampling rate at 2 GHz and initial sampling instants delayed by multiples of $t_{Nyquist} = 1/(20 \text{ GHz}) = 0.05$ ns. We can observe that the sub-sampled waveforms are different under different initial sampling instant and the actual signal energy captured varies largely. Hence it is necessary to pay attention to the initial sampling instant. We can calculate the proper initial sampling instance from the time locations of the pulse peak and the peak of the filter impulse response. For example, we consider a nearly ideal BPF with impulse response as a truncated sinc function with carrier frequency at 4 GHz. If the time-truncated impulse response has a length of 10 ns, the peak value will

4.4 Improved Detection Scheme from Initial Sampling Instant

Figure 4.12: Captured samples with various initial sampling instance under noise-less condition.

4.4 Improved Detection Scheme from Initial Sampling Instant

appear at $t = 10 \text{ ns}/2 = 5 \text{ ns}$. The peak value of the 6th order Gaussian pulse is at $t = 0.7 \text{ ns}/2 = 0.35 \text{ ns}$. Then the initial sampling instant should be chosen at $t_{i,best} = (5 + 0.35) \text{ ns} - \Gamma t_s$ so that peaks of the BPF output can be captured. Note that t_s is the sampling period of the ADC and Γ is the largest integer that satisfies $0 \leq t_i \leq t_s$.

Based on the above observation, we propose an improved version of the pseudo-coherent receiver with an improved initial sampling instant controlled by a time delay of $t_{i,best}$ in the ADC which is shorter than t_s . From Fig. 4.12, we can observe that the graph corresponding to $t_i = 9t_{Nyquist}$ captures the largest amount of signal energy and it corresponds to the case of $t_{i,best}$. For other parallel receiver units located at frequency bands other than 4 GHz, proper time delay is also added to the corresponding ADC in order to capture as much signal energy as possible.

4.4.2 Numerical Results and Discussions

For simulation, we still adopt the 6th order derivative Gaussian pulse with duration of 0.7 ns and the channel model is the CM1 environment of the IEEE 802.15.4a standard [26], whereby the type of fading is based on the modified Saleh-Valenzuela model. We design the transmission rate such that T_f is longer than the maximum channel delay, and hence, there is no IFI in the received signal. The signaling scheme is BOC-OOK. The number of symbol period required for template cleaning is set as $N = 200$. A 3-unit receiver is adopted with BPF's of 2 GHz bandwidth located at 4 GHz, 6 GHz and 8 GHz respectively to cover most of the UWB band which is from 3.1 GHz to 10.6 GHz. The outputs from these three receiver units are combined using equal gain combining and detection decision is made based on the polarity of the combined output. The sampling rate of the ADC's used in the

4.4 Improved Detection Scheme from Initial Sampling Instant

BPDS component is 2 GHz which is equal to the bandwidth of the bandpass filter and it is merely one tenth of the Nyquist sampling rate of UWB signals.

Figure 4.13: BER performance comparison between the improved receiver and the original pseudo-coherent receiver.

Figure 4.13 compares the performance of the improved receiver proposed in Section 4.4.1 with the original BPDS receiver. The performance of the conventional energy detector (CED) and the all-RAKE receiver is also presented as benchmarks. It is observed that with improved sampling instant, the receiver performance can be improved by 0.7 dB at $BER = 10^{-4}$. Although the improvement is not significant, we would like to highlight that there is huge performance difference between the optimized and the worst sampling instant scenarios shown in the figure. At

4.4 Improved Detection Scheme from Initial Sampling Instant

the worst sampling instant scenario, a lot of signal energy is lost. The resulted noisy decision variable greatly affects the BER performance. We also have observed that the improved receiver can deliver comparable performance to the same receiver without the BPDS component and operating at 20 GHz sampling rate. The best possible performance of the improved receiver occurs when $N \rightarrow \infty$. It is very close to the performance of the optimal all-RAKE receiver under the AWGN environment. Note that large N value implies not only better detection performance but also longer system delay. From the above observation, we can conclude that proper control on the sampling instant is crucial to the system design.

Figure 4.14: BER performance comparison with various N values.

4.5 Chapter Summary

Figure 4.14 compares the BER performance of the improved receiver with the original receiver with respect to various N values. It is observed that by adopting the best sampling instant, the improved receiver requires much less symbol periods to obtain a cleaned correlation template to achieve the same performance of the original receiver. At $E_b/N_0 = 12$ dB, the BER of the improved receiver with $N = 50$ is as good as the BER of the original receiver with $N = 200$. That means that the improved receiver can provide similar system performance with a much shorter system delay, which is desirable for implementation.

In short, we have proposed an improved version of our previously proposed pseudo-coherent BPDS receiver for OOK/PPM signals in the context of IR-UWB communications. With proper control on the initial sampling instant, the improved receiver operating at 2 GHz sampling rate can achieve the performance of receiver operating at 20 GHz. It is also shown that the initial sampling instant has great impact on the detection performance of the receiver and it is crucial to the system design.

4.5 Chapter Summary

Under single-user scenario, this chapter emphasizes on the baseband signal processing for impulse-radio UWB (IR-UWB) communication systems. Firstly, we have proposed a pseudo-coherent detection scheme for on-off keying (OOK) and PPM UWB signals through investigating the interconnection between OOK/PPM signaling and transmitted reference (TR) signaling schemes. The main advantage of the proposed pseudo-coherent detection scheme is that it eliminates the zero-lag auto-correlation component of noise arisen from conventional energy detection

4.5 Chapter Summary

scheme, and therefore, enhances the signal-to-noise ratio of the detection statistics and the bit-error-rate performance. It also embraces threshold-less detection, which does not require extra circuitry to estimate the optimal threshold as required by conventional energy detection schemes.

Furthermore, considering the low-complexity digital implementation of the proposed detection scheme, we have realized the bottleneck on sampling rate requirement. With the aim of signal detection rather than signal reconstruction, we have proposed a novel sub-sampling scheme, namely bandpass downsampling (BPDS) scheme. With BPDS, the IR-UWB receiver can provide a performance near to that under Nyquist sampling rate and yet the actual sampling rate is only one tenth of the Nyquist rate. Note that as our proposed IR-UWB receiver contains a statistical averaging component, their applications are constrained to quasi-static channel condition and to systems that can tolerate certain system delay. Quasi-static condition means the channel fading pattern will not change in the template cleaning process of less than $50 \mu s$ for $N = 500$ data bits to be detected as a group. This is also a reasonable assumption as the IEEE 802.15.4a channel model has a much longer coherence time in the order of $30 ms$, as discussed in this chapter.

We would like to highlight that the application of Q -function in our analysis framework is valid. Q -function is applicable if the final decision making statistics are Gaussian distributed or can be approximated as Gaussian distribution. Our proposed receivers consider the convolution of the fading channel $h(t)$ with the transmitted UWB pulses $s(t)$ as the equivalent transmitted information. Hence, the proposed receivers treat the communication channel as an equivalent AWGN channel. Furthermore, our proposed receivers also integrate the autocorrelation results across the length of the fading channel. The integration results can be

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approximated as Gaussian distributed as there are nearly a hundred samples of the same stochastic process being integrated. The Gaussian approximation, and hence the use of Q -function, is valid as supported by our simulation results.

To maintain the consistency of our work, we have focused our study on one specific pulse shape, namely, the sixth-order Gaussian monocycle. We choose this pulse shape as it can be easily implemented with low-complexity circuitry. Furthermore, it is within the FCC spectrum mask as discussed in the pulse shape design portion of Chapter 2. For other potential easy-implementation pulse shapes, our work on pseudo-coherent detection and BPDS can be extended further.

Although we have not explicitly discussed on the detection of PPM or multiple-pulse-position-modulation (MPPM) signals, our receivers can be generalized and applied to detect both of them for multi-user systems. Furthermore, a general k -length orthogonal code can be applied and detected by our receiver. This will be discussed in detail in the next Chapter on multiple-access IR-UWB systems.

Chapter 5

Multiple-Access Impulse-Radio UWB System

In chapter 3 and 4, we mainly focused on the single-user UWB system. In this chapter, we extend our study to multiple-access techniques and multiple access interference (MAI) mitigation schemes for broadcasting or downlink IR-UWB systems and to provide improved performance and capacity.

For IR-UWB communication systems, there are two popular multiple-access schemes, namely, time hopping (TH) and direct sequence (DS) [68]. TH scheme distinguishes users by their pulse-arrival-time sequences whereas DS scheme differentiates users by their pulse-polarity sequences.

Extensive studies have been carried out on the multiple-access performance of TH-UWB and DS-UWB systems [69], [70]. In particular, a comprehensive performance comparison between TH-UWB and DS-UWB has been presented in [70]. It has been shown in [70] that DS-UWB system is superior to the TH-UWB system in terms of the bit-error rate (BER) performance. Hence, we adopt

the DS scheme as the multiple-access scheme of our proposed downlink UWB communication system.

Walsh codes are well known balanced codes to provide perfect cross correlation performance so long as code pairs are synchronous and they are widely adopted in code-division-multiple-access (CDMA) systems to provide orthogonality for downlink transmission [71]. Some applications of Walsh codes in UWB systems can be found in [72], [73], and [74]. Unlike most existing applications which use bipolar Walsh codes (BWCs), our proposed system adopts unipolar Walsh codes (UWCs) which can reduce hardware complexity for the transmitter. Although the orthogonality across users is destroyed at transmission by using UWCs, but it can be fully restored at the receivers by using the corresponding BWCs as the despread-ing codes, thanks to the property of UWCs that each code has equal numbers of ones and zeros. Transmission using UWCs is also helpful in noise reduction during reconstruction of the correlation template at the receiver side. With UWCs, our proposed multi-user system has the potential to outperform the single-user system in terms of BER, conditioned on equal transmitted signal energy for each user and perfect synchronization. The details will be discussed later in the corresponding numerical results.

Based on the pseudo-coherent OOK detector with BPDS as shown in Fig. 4.6, we propose a receiver for each user to detect the targeted UWB signals modulated by UWCs. This proposed receiver can be implemented using digital circuit components with low complexity, as the sampling rate required is significantly reduced by BPDS technique.

When the timing acquisition is perfect, the orthogonality among all users can be perfectly restored at the receivers of our proposed system. However, when timing

5.1 Multiple-Access UWB System Model

acquisition error (TAE) appears, MAI will occur. Attributed to the property of our proposed receiver that there is always a portion of the correlation template in phase with the user's signal without knowing and correcting the exact TAE, the proposed receiver can tolerate certain amount of TAE and recover some of the transmitted data. It is observed that the BER sensitivity of the proposed system is not symmetric with respect to the TAE. Due to the asymmetric energy distribution of the propagation channel, the impact on the BER performance is larger when TAE is positive compared with the case when TAE is negative. The BER performance of the proposed system is evaluated with respect to TAE.

In the following sections of this chapter, we will describe in detail the system model of the proposed multi-user UWB communication system, the theoretical analysis on the BER performance, and the numerical results with discussions.

5.1 Multiple-Access UWB System Model

5.1.1 Base Station Signaling Scheme

In this work, we focus on the broadcasting or downlink scenario, where the base station transmits to all the users in a synchronous manner. Our system can also be used for uplink scenario. However, the synchronization difference among users must be within $\pm 0.5T_c$ for satisfactory performance, where T_c is the chip duration. This will be discussed in detail later under the timing acquisition error section. For the data transmission scheme of our proposed system, we adopt UWCs in order to benefit from the low complexity of on-off-keying (OOK) type of transmission and orthogonality of Walsh codes. The transmitted signal for the k -th user from the base station can be expressed as

5.1 Multiple-Access UWB System Model

$$s^k(t) = \sqrt{E_p} \sum_{j=-\infty}^{+\infty} \sum_{i=1}^{N_c} a_{i,j}^k p(t - jT_s - (i-1)T_c) \quad (5.1)$$

where E_p is the pulse energy to be transmitted, $p(t)$ is the signal pulse with duration T_p and normalized energy, T_c is the chip duration with $T_c \gg T_p$ so that there will be no inter-chip interference (ICI) or inter-symbol interference (ISI) after signal propagates through the channel. Note that N_c is the number of chips in one symbol, $T_s = N_c T_c$ is the symbol duration, and

$$a_{i,j}^k = \begin{cases} c_i^k, & \text{for } d_j^k = 1 \\ |c_i^k - 1|, & \text{for } d_j^k = -1 \end{cases} \quad (5.2a)$$

$$(5.2b)$$

where $a_{i,j}^k \in \{0, 1\}$ is the amplitude of the i -th chip in the j -th symbol of the k -th user. Note that $d_j^k \in \{1, -1\}$ is the j -th binary data symbol, $c_i^k \in \{0, 1\}$ is the i -th chip in the UWC for the k -th user. We illustrate the 4-by-4 unipolar Walsh matrix $\mathbf{C}_4$ in (5.3). By using its row vectors as the spreading codes, 3 users can be supported by the system, since the first row consisting of all ones has different properties from other rows. Note that for applications adopting BWCs, the first all-one row is normally occupied for the transmission of pilot symbols or auxiliary symbols to help signal detection as in [74]. Our proposed system does not require pilot symbols for template reconstruction, and hence, it is not necessary to transmit the all-one row for our proposed system.

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$$\mathbf{C}_4 = \begin{bmatrix} 1 & 1 & 1 & 1 \\ 0 & 1 & 0 & 1 \\ 0 & 0 & 1 & 1 \\ 1 & 0 & 0 & 1 \end{bmatrix} \quad (5.3)$$

Walsh codes are commonly used in the bipolar form to achieve orthogonality. By adopting OOK modulation, we transform the BWCs into the UWCs. In our proposed system, the orthogonality among users is destroyed at the transmission side because of OOK modulation, however, it can be restored at the receivers by using the corresponding BWCs as the despreading codes. The UWCs are also helpful in reconstructing the correlation template at the receivers. We will discuss them further in the proposed receiver detection scheme.

5.1.2 Receiver Detection Scheme

Figure 5.1: Proposed receiver structure for the k -th user.

The receiver structure for the k -th user is shown in Fig. 5.1. It is developed based on the pseudo-coherent receiver for single-user systems proposed in Chapter 4. The BPDS component enables analog-to-digital conversion with a subsampling

5.1 Multiple-Access UWB System Model

rate significantly below the Nyquist sampling rate for UWB signals. Constituted by an analog bandpass filter (BPF) and a relatively low-rate analog-to-digital converter (ADC), the BPDS component attempts to achieve full aliasing with the proper design on the center frequency of the BPF f_c , the bandwidth of the BPF B_{BPF} , and the sampling frequency of the ADC f_s . With the conditions that $2f_c/B_{BPF}$ is a positive integer and $f_s = B_{BPF}$, full aliasing occurs such that for the bandpass filtered signal, the spectrum of negative frequency completely overlaps with the spectrum of positive frequency after subsampling. For the single-user system in Chapter 4, it has been shown that with BPDS, the sampling rate requirement can be reduced to one tenth of the Nyquist rate and the detection performance only has small degradation compared with that of a system operating at the Nyquist rate. Besides, by adopting the BPDS technique, the subsequent part of the proposed receiver can then be fully implemented digitally without difficulty in building the delay component required by the template cleaning filter (TCF), which is hard to build by using analog component.

Conditioned on perfect timing acquisition, the subsampled signal of the i -th chip of the j -th received symbol after the BPDS component can be represented as

$$\mathbf{r}_{i,j} = \sqrt{E_q} \sum_{\kappa=1}^K a_{i,j}^{\kappa} \begin{bmatrix} \mathbf{q} \\ \mathbf{0}_{(L-y) \times 1} \end{bmatrix} + \mathbf{n}_{i,j} \quad (5.4)$$

where $L = T_c f_s$ is the number of samples obtained within each chip duration after the signal being sampled at the rate of f_s , y is the length of the received waveform vector $\mathbf{q}$ which is the sampled version of $q(t)$ and E_q is the energy of $q(t)$. Note that $q(t) = \sqrt{E_p/E_q} p(t) * h_{ch}(t) * h_{BPF}(t)$ where $h_{ch}(t)$ and $h_{BPF}(t)$ corresponds to the impulse response of the frequency-selective fading channel and the BPF

5.1 Multiple-Access UWB System Model

used in the BPDS component, respectively. $\mathbf{n}_{i,j}$ is the noise vector and its element has zero mean and variance of $\sigma_n^2 = N_0 B_{BPF}$, where N_0 is the single-sided power spectral density of the AWGN and B_{BPF} is the bandwidth of the BPF which is much smaller than the original UWB signal transmission bandwidth B_T .

After being converted into digital form, the received signal will enter the TCF which estimates the correlation template by averaging the input signals over K users and N chips, where $N = N_s N_c$ and N_s indicates the number of symbols used for averaging. By assuming a quasi-static channel condition where the channel impulse response does not change during the averaging process, the noise level in the estimated correlation template can be significantly reduced and the reduction level is proportional to the number of chips used for averaging. The output of the TCF can be expressed as:

$$\hat{\mathbf{r}}(N) = \frac{\sqrt{E_q}}{2} \begin{bmatrix} \mathbf{q} \\ \mathbf{0}_{(L-y) \times 1} \end{bmatrix} + \hat{\mathbf{n}}(N) \quad (5.5)$$

where $\hat{\mathbf{n}}(N)$ is the noise vector in the estimated template and the variance of its element is equal to $\sigma_n^2/(K^2 N)$.

The noise level in the estimated template can be controlled and further reduced when we adopt the UWCs. The UWCs are of OOK modulation and they have a nice property that the numbers of “on” chips equals that of the “off” chips in each code except for the first row of Walsh matrices which consists of all “on” chips. By adopting the UWCs, equal numbers of “on” and “off” chips are transmitted in one symbol duration regardless of the actual symbols transmitted by different users. Thus, we can have a predictable and controllable noise level in the reconstructed correlation template. The BWCs are also predictable in their chip polarities,

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however, the first row of the Walsh matrices must be transmitted for template estimation since the summation of the chips within one symbol is zero for the other rows. The UWCs do not encounter this issue and the first row needs not be transmitted. Given the same number of users and the same code length adopted in the system, the TCF is able to deliver estimated template with higher signal-to-noise ratio (SNR) by adopting UWCs rather than bipolar ones. We can represent the SNR of the TCF output as:

$$\frac{SNR_{UPW}}{SNR_{BPW}} = \frac{\left(\frac{NK\sqrt{E_q}}{2}\right)^2 \Upsilon_{\mathbf{q}}/E_{\hat{\mathbf{n}}}}{(N\sqrt{E_q})^2 \Upsilon_{\mathbf{q}}/E_{\hat{\mathbf{n}}}} = \frac{K^2}{4} \quad (5.6)$$

where SNR_{UPW} and SNR_{BPW} represent the SNR values of the TCF output by adopting UWCs and BWCs, respectively. Note that $\Upsilon_{\mathbf{q}}$ is the energy of the $\mathbf{q}$ vector, K is the number of users in the system, $E_{\hat{\mathbf{n}}}$ represents the noise energy per chip after TCF which is the same under both cases since it only depends on N , which is the number of chips used for template cleaning. It is observed that as long as $K > 2$, the UWCs are better than the bipolar ones from the perspective of template reconstruction.

Following the TCF is the correlation section of the receiver. The subsampled signal string delayed by N_s symbol durations is then correlated with the TCF output chip by chip. Note that N_s corresponds to the number of symbols required by the TCF for template cleaning. Next, the correlation outputs will undergo the despreading process. The despreading codes orthogonalize different users which are not orthogonal at transmission. The despreading codes are the BWCs mapped from the UWCs adopted by the transmitter. The mapping mechanism from unipolar to bipolar is “off” to -1 and “on” to 1 or $\mathbf{c}^k$ to $\hat{\mathbf{c}}^k = 2\mathbf{c}^k - \mathbf{1}$, where $\mathbf{1}$ is an

5.1 Multiple-Access UWB System Model

all 1 vector with the same length of $\mathbf{c}^k$. By showing the cross-correlation value between the despreading code of user k and the spreading code of user b in (5.7), we can demonstrate that the orthogonality across all users can be restored after despreading, thanks to the nice property of the bipolar Walsh matrices that the summation of any of the row vectors is zero excluding the first row, i.e.,

$$\begin{aligned}\mathbf{c}^b \cdot \hat{\mathbf{c}}^k &= \frac{\hat{\mathbf{c}}^b \cdot \hat{\mathbf{c}}^k}{2} + \frac{1}{2} \sum_{i=1}^{N_c} \hat{c}_i^k \\ &= 0\end{aligned}\tag{5.7}$$

where $\cdot$ represents the dot product operation, $k, b \in \{1, 2, \dots, K\}$ and $k \neq b$. With the notations from (5.7), the correlation value of $\mathbf{c}^k$ and $\hat{\mathbf{c}}^k$ can be expressed as:

$$\begin{aligned}\mathbf{c}^k \cdot \hat{\mathbf{c}}^k &= \frac{\hat{\mathbf{c}}^k \cdot \hat{\mathbf{c}}^k}{2} + \frac{1}{2} \sum_{i=1}^{N_c} \hat{c}_i^k \\ &= N_c/2.\end{aligned}\tag{5.8}$$

It is worth mentioning that the proposed downlink system targets low-rate communication and meanwhile it has the requirement on quasi-static channel condition and users can tolerate certain system delay introduced by the TCF.

The receiver structure presented in Fig. 5.1 only has one BPDS component that covers a fraction of the whole signal spectrum. To enhance the frequency diversity, we can treat the structure in Fig. 5.1 as a basic receiver unit and multiple such units can be implemented in parallel so that the BPFs in the BPDS components can cover distinct parts of the desired frequency band. Then the decision variables

5.2 Performance Analysis of the Proposed System

from each receiver unit can be combined to achieve better detection performance.

5.2 Performance Analysis of the Proposed System

The bit error probability of the proposed communication system is derived in this section. We assume that the system is operating under a quasi-static channel condition and there are K users in the system. Each user has a receiver consisting of M receiver units. Equal gain combining (EGC) is adopted in each user's receiver to combine the outputs from these M receiver units, since EGC will not add extra complexity into the receiver for weight estimation required by other diversity combining schemes such as selection combining (SC) and maximal-ratio combining (MRC) [58].

5.2.1 Perfect Timing Acquisition

Firstly, we discuss the detection error probability when the receivers can recover the timing information perfectly. Under this circumstance, information from different users is well separated with no multi-user interference. Without loss of generality, we assume that user k is the targeted user since all users' BER performance is the same. The chip energy E_c of the signal after BPDS is of the form $E_c = E_q \mathbf{q}^T \mathbf{q}$, where $()^T$ denotes the matrix transpose operation. We first consider the simplest case that only one receiver unit is adopted as shown in Fig. 5.1. The correlation output for the i -th chip of j -th symbol $U_{i,j} = \hat{\mathbf{r}}(N)^T \mathbf{r}_{i,j}$ and the decision variable for the j -th symbol can be expressed as

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$$\begin{aligned}
Z_j^k &= \sum_{i=1}^{N_c} \hat{c}_i^k U_{i,j} \\
&= \sum_{i=1}^{N_c} \hat{c}_i^k \left\{ \frac{E_q}{2} \sum_{\kappa=1}^K a_{i,j}^\kappa \mathbf{q}^T \mathbf{q} + n_{U_{i,j},1}(N) + n_{U_{i,j},2}(N) + n_{U_{i,j},3}(N) \right\} \\
&= \frac{E_c}{2} \sum_{i=1}^{N_c} \hat{c}_i^k \sum_{\kappa=1}^K a_{i,j}^\kappa + n_{Z,1}(N) + n_{Z,2}(N) + n_{Z,3}(N) \\
&= \pm \frac{E_c N_c}{4} + n_{Z,1}(N) + n_{Z,2}(N) + n_{Z,3}(N)
\end{aligned} \tag{5.9}$$

where $n_{U_{i,j},1}(N)$ corresponds to the correlation output of the noise-free data waveform of the i -th chip in the j -th symbol with the noise in the estimated correlation template $\hat{n}(N)$, $n_{U_{i,j},2}(N)$ corresponds to the correlation output of the noise-free template waveform and the noise in the data waveform $n_{i,j}$, and $n_{U_{i,j},3}(N)$ is the inner product of both noise components in the template and data vector. Furthermore, $n_{Z,1}(N)$, $n_{Z,2}(N)$ and $n_{Z,3}(N)$ correspond to the noise components after despreading and they have zero mean and variance of $N_c^2 E_c \sigma_n^2 / (4K^2 N)$, $E_c N_c \sigma_n^2 / 4$, and $N_c L \sigma_n^4 / (K^2 N)$, respectively. We approximate the sum of these three noise components as a Gaussian random variable of zero mean and variance equal to the sum of the above three noise variances.

Since the decision variable for symbol “1” is symmetrically distributed to that of symbol “-1”, the BER can be represented as

$$\begin{aligned}
P_{e,N}^k &= P\{Z_j^k(N|d_j^k = 1) < 0\} \\
&\approx Q \left(\sqrt{\frac{\lambda^2}{(\frac{1}{K^2 N} + \frac{1}{N_c})2\lambda + \frac{4L}{N_c K^2 N}}} \right)
\end{aligned} \tag{5.10}$$

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where $\lambda = E_c/(2N_0B_{BPF})$ and $Q(x)$ is defined as $\frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$ [57].

More receiver units can be adopted to increase the frequency diversity. When M receiver units are used and the outputs from all the receivers are combined based on the EGC scheme, the probability error can be expressed as

$$P_{e,N,M}^k = P \left\{ \sum_{m=1}^M Z_{j,m}^k (N|d_j^k = 1) < 0 \right\} \\ \approx Q \left(\sqrt{\frac{(\sum_{m=1}^M \lambda_m)^2}{(\frac{1}{K^2N} + \frac{1}{N_c})2 \sum_{m=1}^M \lambda_m + \frac{4ML}{N_c K^2N}}} \right) \quad (5.11)$$

where $Z_{j,m}^k$ represents the decision variable of symbol j from receiver unit m and $\lambda_m = E_{c,m}/(2N_0B_{BPF})$. Note that $E_{c,m}$ represents the E_c of receiver unit m . It is worth mentioning again that the channel condition considered is a multipath fading channel. The multipath component in $h_{ch}(t)$ has been captured in the received waveform $q(t) = h_{ch}(t) * p(t)$ and subsequently the digitized vector $\mathbf{q}$. $q(t)$ is not the received pulse shape but rather the complete waveform within a channel delay length.

5.2.2 Imperfect Timing Acquisition

When timing acquisition error occurs, MAI will be introduced. For different users, the interference level is different since the asynchronous cross-correlation values of Walsh codes varies. It is assumed that coarse timing acquisition can be achieved in the chip level [10], and hence, the timing acquisition error is in the range of $t_e \in [-0.5T_c, 0.5T_c]$. For each user, we can calculate the exact interference pattern which contains all possible interference levels and their probability of occurrence.

5.2 Performance Analysis of the Proposed System

Conditioned on each interference level, we can have a BER expression in the form of (5.11) with numerator being subtracted by the interference level. The interference level is determined by the users' sequences and t_e . When $t_e > 0$, the receiver has a timing lagging behind the true one, whereas when $t_e < 0$, the receiver has a timing leading the true one. Since t_e is introduced right at the beginning of signal reception, the shape of the reconstructed correlation template is shifted accordingly in a circular manner with respect to the shape under perfect timing. In other words, the reconstructed correlation template always has some portion in phase with the received signals. Attributed to this characteristic of the proposed receiver, the targeted user can still detect the data without estimation and correction on t_e . Note that the data detection will have worse BER performance, as the out-of-phase portion of the correlation template leads to MAI.

The in-phase correlation and despreading contribute positively to the decision variables whereas the out-of-phase part generates interference. The in-phase and out-of-phase chip energies are labeled as $E_{c,in}$ and $E_{c,out}$, respectively. They satisfy the relationship that $E_{c,in} + E_{c,out} = E_c$. $E_{c,in}$ can be calculated as $E_{c,in} = E_q \bar{\mathbf{q}}^T \bar{\mathbf{q}}$ where $\bar{\mathbf{q}}$ is determined by t_e as $\bar{\mathbf{q}} = [q_1, q_2, \dots, q_{(T_c+t_e)f_s}]^T$ for $t_e < 0$ and $\bar{\mathbf{q}} = [q_{t_e f_s + 1}, q_{t_e f_s + 2}, \dots, q_{T_c f_s}]^T$ for $t_e > 0$. Note that $q_i, i \in [1, T_c f_s]$ is the element in vector $\mathbf{q}$.

The overall interference level to the targeted user is a multiple of $E_{c,out}$, which can be determined by the cross-correlation values of the UWCs and the corresponding BWCs. The code-level interference is dependent on the observation time instant and the actual data symbols transmitted by all users. Assuming that user k is the targeted user, the code length is N_c for each user, and the observation time is at symbol j , the code-level interference to user k can be expressed as

5.2 Performance Analysis of the Proposed System

$$I_j^k = \sum_{b=1}^K R_{t_j}^{k,b} \quad (5.12)$$

where

$$R_j^{k,b} = \begin{cases} \sum_{i=1}^{N_c-1} \hat{c}_{i+1}^k a_{i,j}^b + \hat{c}_1^k a_{N_c,j-1}^b, & \text{for } -0.5T_c \leq t_e < 0 \\ \sum_{i=1}^{N_c-1} \hat{c}_i^k a_{i+1,j}^b + \hat{c}_{N_c}^k a_{1,j+1}^b, & \text{for } 0 < t_e < 0.5T_c. \end{cases} \quad (5.13a)$$

$$(5.13b)$$

Note that the self-interference from the previous symbol or following symbol of user k is also considered. Since each user is free to transmit either symbol “1” or “-1”, each user can introduce 8 interference levels to the targeted user under different t_e . There are K users in total and some interference levels from different users may have the same value, and hence the number of I_j^k with distinctive values is no larger than $8K$. Since the interference patterns to symbol “1” or “-1” are symmetrically distributed, we only need to consider the BER performance when $d_j^k = 1$. Conditioned on $d_j^k = 1$, we then obtain the probability of occurrence for each possible interference level $I_{j,\alpha}^k$ and we label it as p_α^I where $\alpha \in \{1, 2, \dots, A\}$ points to the different interference levels. Note that $A \leq 4K$ represents the total number of interference levels.

The BER for user k with single receiver unit can be represented as

$$P_{e,N}^k = P\{Z_j^k(N|d_j^k = 1) < 0\} \\ \approx \sum_{\alpha=1}^A p_\alpha^I Q\left(\frac{\hat{\lambda}_\alpha}{\sqrt{(\frac{1}{K^2N} + \frac{1}{N_c})2\lambda + \frac{4L}{N_c K^2 N}}}\right) \quad (5.14)$$

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where $\hat{\lambda}_\alpha = (E_{c,in} + I_{j,\alpha}^k \frac{2}{N_c} E_{c,out}) / (2N_0 B_{BPF})$, and for M receiver units, the BER can be calculated as

$$P_{e,N,M}^k = P \left\{ \sum_{m=1}^M Z_{j,m}^k (N | d_j^k = 1) < 0 \right\} \\ \approx \sum_{\alpha=1}^A p_\alpha^I Q \left(\frac{\sum_{m=1}^M \hat{\lambda}_{m,\alpha}}{\sqrt{(\frac{1}{K^2 N} + \frac{1}{N_c}) 2 \sum_{m=1}^M \lambda_m + \frac{4ML}{N_c K^2 N}}} \right) \quad (5.15)$$

where $\hat{\lambda}_{m,\alpha} = (E_{c,m,in} + I_{j,\alpha}^k \frac{2}{N_c} E_{c,m,out}) / (2N_0 B_{BPF})$ and $E_{c,m,in}$ and $E_{c,m,out}$ represent the in-phase and out-of-phase chip energy for receiver unit m , respectively.

The averaged BER performance across all users for single-unit receiver or M -unit receiver implementation can be represented as

$$P_{e,N} = \frac{1}{K} \sum_{k=1}^K P_{e,N}^k \quad (5.16)$$

and

$$P_{e,N,M} = \frac{1}{K} \sum_{k=1}^K P_{e,N,M}^k, \quad (5.17)$$

respectively.

5.3 Numerical Results and Discussions

For simulation, we adopt the 6th order derivative Gaussian pulse with duration of 0.7 ns which satisfies the spectrum mask set by the Federal Communications Commission (FCC) and the channel model is the CM1 environment of the IEEE 802.15.4a standard [26], whereby the type of fading is based on the modified Saleh-

5.3 Numerical Results and Discussions

Valenzuela model. We design the transmission rate such that there is no ICI or ISI in the received signals, and hence, we set $T_c = 41$ ns which is slightly longer than the maximum channel delay of 40 ns. The signaling scheme adopts UWCs. The number of symbols required for template cleaning is set to N_s symbols, and hence the number of chips required is $N = N_s N_c$. Three receiver units are adopted with BPFs of $B_{BPF} = 2$ GHz each located at 4 GHz, 6 GHz and 8 GHz respectively to cover most of the UWB band from 3.1 GHz to 10.6 GHz. The outputs from these three receiver units are combined using EGC and detection decision is made based on the polarity of the combined output. The sampling rate of the ADCs used in the BPDS component is $f_s = 2$ GHz which is equal to the bandwidth of the BPF and it is merely one tenth of the Nyquist sampling rate of the UWB signals.

Figure 5.2 compares the performance of our proposed system under perfect timing acquisition with the lower-bound performance for direct-sequence systems which is equivalent to that of a single-user all-RAKE receiver. The BER is equal to $Q(\sqrt{E_b/N_0})$ for the all-RAKE reception on the unipolar signals [58]. The number of users is assumed to be $K = 7$, the code length is $N_c = 8$ and the number of symbols used for template cleaning is $N_s = 50$ which corresponds to a delay of $N_s N_c T_c = 50 \times 8 \times 41$ ns = 16.4 μ s. It is a reasonable assumption that the channel remains static during the template cleaning process since the IEEE 802.15.4a channel model has a coherence time on the order of 30 ms [61]. With perfect timing acquisition, MAI is avoided at the receiver. It is observed that with certain system delay, our proposed system with 3-unit receiver can have a performance close to the all-RAKE performance with less than 1-dB away at BER of 10^{-4} . We also provide the performance of a 1-unit receiver with the BPF placed at 6 GHz where the

5.3 Numerical Results and Discussions

Figure 5.2: BER performance comparison between the proposed system and the conventional ones.

5.3 Numerical Results and Discussions

signal's highest power-spectral density is located. Because of the loss in frequency diversity, the 1-unit receiver performs worse than the 3-unit receiver. Our proposed receiver is easier to implement than the all-RAKE receiver in practice. Unlike the all-RAKE receiver, our proposed receiver does not require the exact knowledge on the channel impulse response and the received pulse shape beforehand. Instead, this information is jointly estimated during signal detection. Besides, our proposed receiver is a sub-sampling digital receiver without the needs for numerous analog correlators and precise delay lines to capture multipath information. Besides the all-RAKE receiver, we also provide the performance comparison between our system and the non-coherent energy detection receiver in which each chip is processed by energy detector before despreading process. The proposed receiver can provide a significant amount of performance improvement as compared with the energy detection receiver with acceptable increase in complexity. Theoretical performance of the proposed system is also shown in Fig. 5.2 which matches well with the simulation results.

In addition to the performance comparison among multi-user systems, we also show in Fig. 5.2 the single-user performance of our proposed system with $N = 400$. It is observed that the multi-user detection is able to outperform the single-user case conditioned on equal transmission energy for each user and perfect timing acquisition. The performance gain comes from the clean correlation template obtained with multiple users' signal. Since the template reconstruction utilizes all users' signal energy, the SNR of the reconstructed correlation template will be much higher with more users in the system.

Figure 5.3 presents the BER performance of the proposed multi-user system with respect to the timing acquisition errors. We still consider the case where

Figure 5.3: BER performance with various timing error t_e .

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$M = 3$, $K = 7$, $N_c = 8$ and $N_s = 50$. For neat presentation, we only include two simulation results for $t_e = -0.5T_c$ and $t_e = +0.1T_c$. As t_e is within a chip duration, the variation of t_e does not affect the code-level interference but only the in-phase and out-of-phase chip energy denoted by $E_{c,in}$ and $E_{c,out}$, respectively. It is shown that the BER performance is not symmetric with respect to $t_e = 0$. As the channel impulse response does not distribute the signal energy evenly, the clustering of the signal energy in certain region results in the unbalance in the error performance. As most of the signal energy concentrates in the front part of the channel response, the proposed system can tolerate more timing error when $t_e < 0$, because $E_{c,in}$ can still be much larger than $E_{c,out}$. When $t_e > 0$, $E_{c,in}$ decreases very fast as t_e increases, which results in more degradation in BER performance.

Besides UWC, another candidate of coding scheme could be the balanced unipolar orthogonal Gold code (BUOGC) which is a subset of the unipolar orthogonal Gold code [75, 76]. Each BUOGC has equal number of “on” and “off” chips. Assuming the code length is N_c , there will be $N_c/2$ number of codes available in the corresponding BUOGC set. Similar to UWC, BUOGC has perfect cross-correlation performance when they are aligned properly. With timing offset of less than one chip duration, the cross-correlation performance of BUOGC is better than that of UWC.

Figure 5.4 compares the BER performance of the proposed receiver under both coding schemes of UWC and BUOGC. The parameters are set as $M = 3$ and $N = 320$ for all performance curves in this figure. It is observed that there is a trade-off between these two coding schemes on system capacity and the BER performance. Conditioned on equal code length, UWC can accommodate nearly double the number of users that can be supported by BUOGC with some degradation in the

5.3 Numerical Results and Discussions

Figure 5.4: BER performance of UWC and BUOGC with equal N_c and t_e .

5.3 Numerical Results and Discussions

BER performance. When the out-of-phase chip energy $E_{c,out}$ is large as in the case of $t_e = 0.2T_c$, the advantage of BUOGC in the BER performance is more observable. When $E_{c,out}$ is small, the overall MAI level is not very high and the BER performance of BUOGC is comparable with that of UWC. Hence, we suggest that when the system has small number of users and low detection error is desired, BUOGC can be considered, otherwise, UWC is a better candidate.

Figure 5.5: BER performance of the proposed system with various number of users at $t_e = -0.5T_c$.

In Figure 5.5, the effects of the number of users K on the BER performance of a 3-unit receiver are presented with a fixed value of $t_e = -0.5T_c$. The number

5.3 Numerical Results and Discussions

of chips required for template cleaning is kept at $N = 320$ for all cases. As the number of users increases, the BER performance varies little. It is observed that both numerator and denominator in (5.15) decreases as K and N_c increase. Intuitively speaking, when K increases, signal energy from more users can be used to reconstruct the correlation template which results in a cleaner correlation template. However, an increase in K also requires an increase in N_c which causes a decrease in E_c , $E_{c,in}$ and $E_{c,out}$ for all receiver units since the symbol energy $E_s = E_c N_c / 2$ for each user remains unchanged. Meanwhile, the change in users' sequences due to the change in K causes the change in p_α^I , $I_{j,\alpha}^k$ and subsequently the overall MAI. The compound effect keeps the BER performance relatively stable when K changes.

Lastly, we would like to highlight the characteristics of the proposed system followed by the potential applications. The proposed multi-user UWB system adopts Walsh codes in unipolar format which is easier to implement for transmitters without affecting the implementation of receivers. UWCs can also facilitate the reconstruction of correlation template at the receivers. By doing so, we are able to enhance the SNR of the correlation template compared with that under BWCs. To maintain the orthogonality among all users, Walsh codes must be transmitted synchronously, and hence, our proposed system is more suitable for downlink transmission. For the receiver design, we adopt BPDS technique which can greatly reduce the sampling rate requirement to one tenth of the Nyquist sampling rate for UWB signals. However, the proposed receiver operates under a quasi-static channel condition and it needs a relatively long system delay in order to obtain a clean correlation template. The proposed receiver can tolerate certain amount of timing acquisition error as long as the in-phase chip energy is much

larger than the out-of-phase counterpart. Since the proposed system can tolerate certain amount of error in timing acquisition or error in synchronization, we can apply time-hopping techniques during transmission to further smooth the signal spectrum if required. With these characteristics, our proposed system is suitable for low-data-rate downlink applications that operate under quasi-static channel and can tolerate certain system delay.

5.4 Chapter Summary

In this chapter, we have proposed a low-rate UWB communication system with multiple access capability for downlink transmission. Unipolar Walsh codes (UWCs) are adopted for transmission of multi-user information. The adoption of UWCs is helpful in reconstruction of the correlation template at the receivers. Although the orthogonality among all the users is destroyed at the transmitter, it can be restored after despreading by using the corresponding bipolar Walsh codes (BWCs). To facilitate digital implementation of the receiver, bandpass downsampling (BPDS) technique is adopted to significantly reduce the sampling rate requirement to one tenth of the Nyquist sampling rate of impulse-radio UWB (IR-UWB) signals. We also consider the timing acquisition issue of the receivers. For both perfect and imperfect timing acquisition scenarios, theoretical analysis on the bit-error-rate (BER) performance has been provided, which has also been validated by simulation results adopting IEEE 802.15.4a channel model. The proposed receiver can provide performance close to that of an all-RAKE receiver when perfect timing acquisition is achieved and it is easier to implement than the all-RAKE receiver. Besides, the proposed receiver can tolerate certain amount of timing acquisition

5.4 Chapter Summary

error and yet provide satisfactory BER performance.

Chapter 6

Conclusion and Future Work

6.1 Conclusion

We have proposed a Transmitted Reference Ultra-Wideband (TR-UWB) system with both narrowband interference (NBI) and inter-pulse interference (IPI) mitigation capability where NBI is suppressed by a notch filter or a band-stop filter and IPI is reduced by the statistical averaging technique. This system can achieve twice the data rate compared with that of conventional TR systems and the bit-error-rate (BER) performance is able to approach that under the additive white Gaussian noise (AWGN) environment. Illustrations on digital notch filter design with finite impulse response (FIR) and infinite impulse response (IIR) implementations have been provided. Besides, the theoretical lower-bound performance has been derived based on Gaussian approximation. Numerical results show that our proposed system is relatively robust to some variations in system design parameters.

We have observed the relationship between the conventional on-off keying

6.1 Conclusion

(OOK) and TR systems. OOK signals can be seen as a special case of pulse-amplitude-modulation transmitted-reference (PAM-TR) signals when the delay between the reference and data pulses is zero. Inspired by the auto-correlation receiver (AcR) used in the TR systems, we proposed a pseudo-coherent detector for OOK signals, which is able to offer significant improvement on BER performance compared with the conventional energy detector. We have also looked into the implementation issues of the proposed OOK detector. Digital implementation is preferred, however, a very high sampling rate is required. To tackle this issue and reduce the receiver complexity, we have proposed a subsampling scheme named bandpass downsampling (BPDS). The BPDS scheme is able to effectively reduce the sampling rate requirement to one tenth of the Nyquist rate of the UWB band with small performance degradation. Theoretical analysis has been provided for the proposed pseudo-coherent OOK detector with BPDS.

Utilizing the study for single user case, we have proposed a multiple access UWB system with balanced unipolar coding in which the number of “on” symbols is equal to that of “off” symbols. Unipolar coding is favored in our system as it can facilitate in obtaining a clean correlation template for receivers, and hence, enhance the detection accuracy. BPDS technique is also adopted in the receiver design to significantly reduce the sampling rate requirement to one tenth of the Nyquist sampling rate for UWB signals. The analysis framework under quasi-synchronous condition is provided for the BER performance of the proposed system. Numerical result matches well with our theoretical analysis. Beyond that, it also indicates that our system can deliver near-optimal performance if perfect timing acquisition is achieved. With certain amount of timing acquisition error, our system performance is still satisfactory.

6.2 Future Work

We have emphasized in low-complexity system design throughout our work. In Chapter 2, we have reviewed the existing low-complexity IR-UWB schemes including TR and OOK schemes with various receiver designs. Our system is built upon those low-complexity schemes. Furthermore, considering the real implementation issues, we have achieved further reduction in receiver complexity by bringing in the BPDS subsampling scheme. Comparing our proposed receivers with the popular RAKE receiver, our receivers do not require exact channel information to get a near-RAKE performance. In addition, we only require one finger unit of the RAKE receiver to achieve the claimed performance. That saves $r-1$ fingers' circuit components assuming there are r multi-paths in the channel. Comparing with the lowest-complexity energy detection receivers, our proposed receivers are slightly more complex as an extra delay line is needed. However, the BER performance enhancement is significant as illustrated in Fig. 4.7. Recently, there is another low-complexity IR-UWB system proposed in [77]. The system in [77] is based on pulse shape design consisting of orthogonal sinusoidal signals with carriers. Without collecting multipath information, the receiver in [77] is already more complex as a RF-IC is needed to remove the carrier first. With that, we can conclude that overall we have achieved our goal in low-complexity system design.

6.2 Future Work

Thus far, our research work has focused on quasi-synchronous multiple-access IR-UWB system with minor timing acquisition error. For future work, one can extend our study to asynchronous multiple access techniques and the corresponding MAI mitigation schemes for IR-UWB systems and to provide improved performance

6.2 Future Work

and capacity. Besides, it is possible to explore more on the sampling theorem to consummate our proposed bandpass downsampling technique.

A multi-user TR-UWB communication system has been proposed in [65]. The proposed system incorporates pseudo-noise (PN) sequences for UWB systems with data pulses jointly modulated by PAM and PPM to enable multiple access. Two PN sequences are assigned to each user; one for reference pulses and the other one for data pulses. Since all the pulses are scrambled by the PN sequences, the interference from other users is noise like to the target user. A mean-based waveform estimation method is derived to obtain a satisfactory template for each correlation detector. It was demonstrated in [65] that the proposed detector substantially outperformed conventional TR detectors and large system capacity became achievable. The work in [65] assumed perfect synchronization across all users. This assumption may not be always valid. It is motivating to investigate more on asynchronous IR-UWB communication systems such as the work in [54]. Moreover, because of the relationship of OOK with TR systems, it is interesting to investigate on the potential of modifying the multi-user TR (MUTR) receiver in [65] for multi-user OOK systems. It could also be interesting to compare the performance of the modified MUTR receiver with that of an energy-detection based multi-user receiver in [78].

The mean-based waveform estimation method in [65] is quite effective in cleaning the reference waveforms from noise and interferences. However, the MAI in the data waveforms remains and each user's signals are detected separately and independently. Thus, multi-user detection (MUD) technique can be applied to the MUTR systems to jointly mitigate MAI. MUD technique is well studied for conventional CDMA systems. Reference [79] provides a comprehensive review on

6.2 Future Work

MUD for DS-CDMA communications. MUD considers all users as signals for each other and jointly detects all signals. It is robust to the near/far effect and the reduced interference leads to higher capacity. It is possible to explore the advantages of MUD for MUTR systems. Successive interference cancellation scheme in MUD concept could be one suitable candidate for MUTR systems to jointly detect signals and mitigate MAI.

A MAI mitigation technique has been presented in [80] for TR or differential UWB communication systems. The work has investigated the case when only a small number of users are present in the system. Conventional Gaussian distribution model is not well suited for the actual MAI statistics. Instead, the author in [80] has proposed generalized Gaussian distribution and Laplace distribution models. The numerical results have supported the author's claim of significant improvement on the receiver BER performance. The detection scheme proposed in [80] requires more computational resources and it can be considered for uplink scenarios.

In Chapter 4, we have presented a new sub-sampling technique namely BPDS. It is capable of reducing the sampling rate to one tenth of the Nyquist rate for the proposed pseudo-coherent OOK detector with little performance degradation. However, if the sampling rate is further reduced, performance degradation will become severe. It is important to explore whether there are other sampling techniques able to further reduce the sampling rate and the relationship between our proposed BPDS and other sampling techniques. One potential solution is to apply the innovation rate of sampling [81]. The innovation rate of sampling provided a theoretical minimum sampling rate necessary to fully recover a signal with finite rate of innovation. The minimum rate is equal to the finite rate of innovation.

6.2 Future Work

The key for signal reconstruction is to identify the innovative part of a signal (e.g., time instants and weights of Diracs) using an annihilating or locator filter: a device well known in spectral analysis and error-correction coding. Applied to IR-UWB signals, the finite rate of innovation can be calculated as $\rho = 2K/\tau$, where K is the number of multipath components and τ is the maximum delay of the channel. For example, if there are 40 paths in a UWB channel with maximum delay of 40 ns, the minimum sampling rate required to recover the signal is $\rho = 2 \times 40/40 \text{ ns} = 2 \text{ GHz}$ which coincides with the sampling rate used in our proposed BPDS. However, our proposed BPDS scheme does not recover the original signals but uses the distorted version to do the detection. It would be quite interesting to explore the interconnection between these two techniques. With the innovation rate of sampling, it is possible to fully recover the channel information and, hence, it is possible to further enhance the performance of multiple-access IR-UWB systems.

Another extension worthwhile investigating is applying our proposed scheme on different pulse shape designs and channel models. Particularly, for fast fading channels, our proposed scheme will face more challenges in getting a good channel estimation through statistical averaging. It is somewhat equivalent to the scenario whereby only a small N value is applicable. We have established the the N related BER performance analysis in both Chapter 4 and Chapter 5. However, if the channel condition could vary within a symbol duration, another analysis model will be beneficial. In addition, to tackle the fast changing channel condition, an improved detection scheme will be necessary.

Author's Publications

Journal Papers

S. Cui, Y. L. Guan, K. C. Teh, and K. H. Li: "Pseudocoherent detection of OOK/PPM signals as zero-delay-transmitted-reference signals with bandpass down-sampling for UWB communications," *IEEE Trans. Veh. Technol.*, vol. 58, pp. 4141-4148, Oct. 2009.

S. Cui, K. C. Teh, K. H. Li, Y. L. Guan, and C. L. Law, "Performance analysis of transmitted-reference UWB systems with narrowband interference suppression," *Wireless Communications and Mobile Computing*, vol. 9, pp. 1081-1088, Aug. 2009.

S. Cui, K. C. Teh, K. H. Li, and Y. L. Guan, "Performance Study of Low-Rate UWB Multiple-Access Communication Systems with Unipolar Coding," under preparation for submission.

Conference Papers

S. Cui, K. C. Teh, K. H. Li, Y. L. Guan, and C. L. Law, "BER performance of transmitted-reference UWB systems with notch filter in the presence of inter-pulse interference and partial-band interference," in *Proc. ICICS*, 2007, pp. 1-5.

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