Orthogonal pulse based modulation schemes for time hopping ultra wideband radio systems

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ORTHOGONAL PULSE BASED MODULATION SCHEMES FOR TIME HOPPING ULTRA WIDEBAND RADIO SYSTEMS

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List of Abbreviations

AC	Absolute Combining
AWGN	Additive White Gaussian Noise
ARAKE	All RAKE
BER	Bit Error Rate
BPM	Bi-Phase Modulation
BPPM	Biorthogonal Pulse Position Modulation
BPSM	Biorthogonal Pulse Shape Modulation
BPSK	Binary Phase Shift Keying
CF	Characteristic Function
DS-UWB	Direct-Sequence UWB
EGC	Equal Gain Combining
EIRP	Effective Isotropic Radiated Power
ESD	Energy Spectral Density
FCC	Federal Communication Commission, USA
FT	Fourier Transformation
GPR	Ground Penetrating Radar
GPS	Global Positioning System
GSSI	Geophysical Survey Systems, Inc.
HP	Hermite Polynomials
IR	Impulse Radio
IR-UWB	Impulse Radio UWB
IDA	Infocomm Development Authority, Singapore
ISI	Inter Symbol Interference
IPI	Inter Pulse Interference

LOS	Line of Sight
MA	Multiple Access
MRC	Maximal Ratio Combining
MAI	Multiple Access Interference
MPI	Multi Pulse Interference
MAI	Multiple Access Interference
MHPs	Modified Hermite Pulses
MMHP	Modulated Modified Hermite Polynomials
NB	Narrowband
NLOS	Non Line of Sight
OFDM	Orthogonal Frequency Division Multiplexing
OOK	On-Off Keying
OPM	Orthogonal Pulse Modulation
OPPM	Orthogonal Pulse Position Modulation
PAM	Pulse Amplitude Modulation
PDF	Probability Density Function
PG	Processing Gain
PPM	Pulse Position Modulation
PSD	Power Spectral Density
PSM	Pulse Shape Modulation
PSWFs	Prolate Spheroidal Wave Functions
PRAKE	Partial RAKE
QPSK	Quadrature Phase Shift Keying
SD	Selection Diversity
SGA	Standard Gaussian Approximation
SIR	Signal to Interference Ratio
SNR	Signal to Noise Ratio
SRAKE	Selected RAKE
SSA	Soft Spectrum Adaption
TH-UWB	Time-Hopping UWB
UFZ	UWB Friendly Zone

U-NII	Unlicensed National Information Infrastructure
UWB	Ultra Wideband
WLAN	Wireless Local Area Networks
WPAN	Wireless Personal Area Networks

List of Symbols

j	$\sqrt{-1}$
$\lfloor . \rfloor$	The floor operator
$\lceil . \rceil$	The ceil operator
$\%$	Remainder
$. $	Norm
$n!$	$n(n-1)(n-2)....1$
$\delta(.)$	Dirac delta function
$\sinh(x)$	$\frac{e^x-e^{-x}}{2}$
$E\{. \}$	Expectation
$\mathcal{F}\{. \}$	Fourier transformation
$\mathcal{N}(a, b)$	Normal distribution function with mean a and variance b
$Q(x)$	$\frac{1}{\sqrt{(2\pi)}} \int_x^\infty e^{-\frac{y^2}{2}} dy$
$w(.)$	Pulse waveform
$s(.)$	Transmitted signal
$h(.)$	Channel response
$r(.)$	Received signal
T_f	Pulse repetition interval
T_c	Chip duration

List of Publications

- [1] S. Majhi, A. S. Madhukumar, and A. B. Premkumar, "Performance of orthogonal based modulation schemes for TH-UWB communication systems," *To appear in IEICE Electronics Express*, 2007.
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 - [10] S. Majhi, A. S. Madhukumar, A. B. Premkumar, and W. Xiang, "Theoretical Capacity Analysis of TH-UWB Systems for Orthogonal Pulse Based Modulation Schemes," To be appeared in *IEEE Vehicular Technology Conference (VTC 2008-Fall)*.
 - [11] S. Majhi, A. S. Madhukumar, and W. Xiang, "Power Spectral Analysis of Orthogonal Pulse Based Modulation Schemes for TH-UWB Systems," Submitted to *IEEE Global Communications Conference (GLOBECOM 2008)*.

Abstract

Time hopping ultra wideband (TH-UWB) communication promises to be a viable technique to build a relatively simple, low-cost and low-power transceiver that can be used for short range high speed wireless communication over the multipath channel. The successful deployment of TH-UWB systems strongly depends on the development of pulse waveforms, modulation techniques and system complexity. The TH-UWB message symbols are transmitted by short analog pulse waveforms. Because of the short pulse waveforms, UWB is capable of providing high data rates for short range communication. Various kinds of modulation schemes have been proposed for the purpose of achieving high data rates. But a major challenge is the selection of an appropriate modulation scheme in designing a system. System model, channel model, channel capacity, interference, system performance, spectral characteristics and coexistence capability are all related issues with in the type of modulation employed. It is in this context, dissertation focusses on modulation schemes based on orthogonal pulses such as modified Hermite pulses (MHPs) and Prolate spheroidal wave functions (PSWFs).

To achieve cost effective system design, a combined modulation scheme, called on-off keying (OOK) and pulse shape modulation (PSM)(OOK-PSM), for M -ary signaling has been proposed. It provides low power and low cost system by using fewer orthogonal pulses and receiver correlators than those used in PSM and biorthogonal PSM (BPSM) schemes. Various interference issues such as MAI, inter pulse interference (IPI) and multi pulse interference (MPI) are analyzed for the combined OOK-PSM scheme. Despite its several advantages, the proposed OOK-PSM scheme cannot be used for higher level modulation schemes in the presence of multipath channel.

To overcome this problem, another combined modulation scheme, using orthogonal pulse position modulation (OPPM) and BPSM (OPPM-BPSM), has been proposed for

higher level modulation schemes to achieve high data rates. An alternative method is described to select the pulse position and pulse order for a symbol instead of using lookup tables. A probabilistic approach of this scheme is presented in the presence of multiple users. The performance of the proposed system is analyzed in the presence of MAI and ISI and the results are verified using simulation studies. Channel capacity of this scheme is also investigated in detail.

Finally, since power spectral density (PSD) plays an important role in coexisting issues, UWB interference on narrowband (NB) system are also investigated based on orthogonal and non orthogonal pulse based modulation schemes. The PSD of orthogonal pulse based modulation schemes such as M -ary PSM, M -ary BPSM and M -ary OPPM-BPSM are analyzed to investigate the spectral efficiency of orthogonal pulses. A generalized pulse waveform based on non orthogonal pulses is proposed to reduce UWB interference on IEEE802.11a system. The total amount of in-band interference power in the IEEE802.11a receiver is calculated in the presence of TH-UWB systems. The performance analysis of IEEE802.11a system shows that the proposed generalized pulse waveform coexists better than the conventional non orthogonal pulse waveforms.

Chapter 1

Introduction

Ultra Wideband (UWB) radio technology has recently received significant attention in both academia and industry for applications in wireless communications. Interest in UWB communication was sparked off by FCC rulings in February 2002 that authorized the unlicensed commercial deployment of UWB technology under strict power control. The huge unlicensed bandwidth and the large available consumer market related to wireless personal area networking (WPAN) devices are the major proponents for research in this area. UWB radio technology provided a practical solution for WPAN and home networking due to its high transmission rate, high-resolution position and tracking abilities, extremely low power spectral density and potential reuse of spectrum resources, minimal interference to other existing communication systems, robust performance under multipath conditions, abilities to have scalable data communication combined with precision location tracking, low implementation cost, low power consumption, etc.

However, understanding how this characteristic affects system performance and design is critical in the implementation of UWB radio devices. From an academic perspective, many fundamental research issues remain unresolved, including the best modulation type for a particular application, propagation characteristics of UWB signal in various environments, interference issues, performance analysis and reduction of complexity in system design. Selection of an appropriate modulation scheme and its detailed analysis in terms

Chapter 1. Introduction

of spectral characteristics, capacity and performance are important for the efficient implementation of UWB radio systems for practical use. It is in this context, the present work analyzes orthogonal pulse based modulation schemes for UWB radio system design.

1.1 History of UWB Radio Technology

Ultra wideband radio technology has been in use for over five decades [1]. Gerald Ross, a researcher with the Sperry Rand Corporation, first introduced UWB radio technology in the late 1960s, which was referred to as “carrier-free”, or “impulse” technology [2]. The technology was first used in a ground penetrating radar system in 1974 and became a success at Geophysical survey systems, Inc. (GSSI) [3]. The term UWB itself was introduced by the US department of defence in the late 1980s [4, 5]. The Department of Defence began to use the technology to image through walls and into the ground [6]. In 1998, the Federal Communications Commission (FCC) realized the potential of UWB technology and began to investigate the possibility of allowing it for commercial use and formed the necessary regulations. The late 1990s saw the move to commercialize UWB communication devices and systems. Even though the knowledge has been in existence for over thirty years, research and development of UWB systems have exploded since the FCC’s allocation of a spectrum for UWB communication on February 14, 2002 [7].

1.2 Background of UWB Radio Technology

The FCC has defined a UWB system as having a -10 dB bandwidth between 3.1 and 10.6 GHz and can operate without any licence [8]. The system should have a signal bandwidth at least 500 MHz or a fractional bandwidth greater than 0.25. The fractional bandwidth is defined as $2(f_H - f_L)/(f_H + f_L)$, where f_L is the lower and f_H is the upper frequency limits, measured at -10 dB below the peak emission point for indoor communication. For outdoor communications, the values are taken at -20 dB below

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the peak emission point. The indoor and outdoor spectrum masks adopted by FCC and infocomm development authority (IDA) of Singapore for the respective countries are shown in Figure 1.1. Effective isotropic radiated power (EIRP) from a UWB system should not exceed -41.3 dBm/MHz for FCC and -35.3 dBm/MHz for IDA for use in UWB friendly zone (UFZ) in Singapore.

There are two major types of UWB transmission schemes. The most common and traditional way of emitting UWB signal is by radiating pulses that are very short in time. This transmission technique goes under the name of impulse radio (IR) [9]. Information is directly modulated on the pulses by using various methods such as pulse position modulation (PPM) or pulse amplitude modulation (PAM) [10,11]. Multiple access (MA) capability can also be supported by using spread spectrum techniques such as time hopping UWB (TH-UWB) and direct sequence UWB (DS-UWB) approaches. In TH-UWB, the entire spectrum is used for one user. Multiple access is accomplished by assigning specific time slots for different users. In DS-UWB, the same spectrum can be shared by multiple users at the same time by assigning a different code to each user. Both methods are known as IR-UWB technology [12,13]. The other transmission scheme is multiband UWB (MB-UWB) based on orthogonal frequency division multiplexing (OFDM), where the spectrum is broken up into channels and each channel is assigned for a different user [14,15]. Frequency hopping is used in this OFDM approach as a way of making this system occupy a large bandwidth for the purpose of increasing the total radiated power.

IR-UWB has several advantages over OFDM-UWB such as its ability to penetrate through materials and resolve multipath with path length differences in the order of a foot or less. IR-UWB also allows for lower power consumption with a low duty cycle, and makes it beneficial for the detection and tracking applications [1,9]. However, there are three primary drawbacks in current implementations of IR-UWB, namely lack of high data rates, lack of long communication range and necessity of analog components

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Figure 1.1: The indoor and outdoor spectral masks as ruled by FCC, USA and IDA, Singapore

in the design of such systems. The primary focus of this work is the development of orthogonal pulse based modulation schemes for IR-UWB systems particularly for TH-UWB for commercial exploitations.

1.3 Applications of UWB Radio Technology

UWB radio technology has many reasons to make it an exciting and useful technology for future wireless communication compared to other traditional narrowband (NB) communication systems [16]. It pushes the limits of high data rate communications, enabling wireless systems to be realized which were never thought to be possible. It creates large instantaneous bandwidth which enables fine time resolution for network time distribution, precision location capability, or use as a radar. The low spectral density of UWB system allows for covert transmission that makes interception or jamming from other users difficult [17]. Since its bandwidth is much larger than the coherence bandwidth

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of the channel, any deep fade affects only a small portion of the total bandwidth. This results in robustness against channel fading without multiple antenna transceivers [18]. Another major advantage is the higher channel capacity; it is expected that the data rates of UWB systems can exceed 1 Gbps as technology scales with Moore's Law [19]. The use of limited power benefits to employ it in sensitive environments, such as hospitals and airports [20].

Because of the characteristics of impulse signaling, it must operate within the noise floor, so that it cannot be easily detected by other radio systems [17, 21]. Therefore, it is suitable for secure applications in military and government sectors. For example, in military sector it is used in battlefield sensor networks, high-resolution radar and ground penetrating radar (GPR) [22, 23]. GPR can operate only when in contact with, or within close proximity of the ground for the purpose of detecting or obtaining images of buried objects.

The major application in commercial sector include communication systems including high-speed home or office networking, desktop and notebook personal computers, PDAs and handheld computers, mobile phones, printers, scanners, portable audio, multimedia players, external hard drives, digital camcorders, digital TVs, advanced set-top boxes, personal video recorders, and DVD players. However, current rules limit communication systems to low-power use that prevents UWB systems from working beyond a relatively short distance (about 30 feet) [24]. In medical imaging systems, it may be used for a variety of health applications to scan the body of a person or animal, similar to X-ray and computed axial tomography scans [25].

1.4 Research Objective and Contributions

UWB technology for communication is not all about advantages. In fact, there are several challenges involved in using nanosecond duration pulses for communications. Perhaps

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the most obvious challenge for UWB radio systems to date has been regulatory problems. Since UWB occupies a wide bandwidth, there are many users whose spectrums are affected by UWB systems. Therefore, it is necessary to insure that UWB does not cause undue interference to existing services. However, all of these depend on the type of modulation schemes used for data transfer in a UWB system.

The main objective of this work is to develop orthogonal pulses based modulation schemes for TH-UWB systems. A variety of modulation schemes have been proposed to enhance the performance of TH-UWB systems. Due to robustness against multiple access interference (MAI) and inter symbol interference (ISI), pulse shape modulation (PSM) scheme is becoming an interesting research topic in TH-UWB systems [26–28]. However, the higher order PSM scheme increases the system complexity by increasing the number of orthogonal pulses and receiver correlators. To address these problems, several combined forms of PSM schemes have been proposed to transmit the same amount of data by using a fewer orthogonal pulses and receiver correlators than those used in conventional PSM scheme [29, 30]. However, these combined schemes increase the system complexity for practical implementation.

To deal with these challenges, a combined modulation scheme based on an on-off keying (OOK) and PSM is proposed for M -ary modulation schemes. The primary focus in this scheme is to build a low power and low cost UWB system by using fewer orthogonal pulses. The use of OOK reduces both system complexity and transmission power. Since the OOK-PSM is used for higher level modulation scheme, synchronization problem for long sequence of zeroes is reduced when compared to the conventional binary OOK scheme [1]. Since the proposed scheme uses orthogonal pulses, it reduces MAI in the receiver. The reduction of MAI is shown through mathematical analysis and simulation. Other interferences such as inter pulse interference (IPI) and multi pulse interference (MPI) are also analyzed for the proposed scheme. The bit error rate (BER) performance

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of the proposed scheme is analyzed through simulation in IEEE 802.15.3a UWB multipath channel model [31]. The performance is compared against other orthogonal pulse based modulation schemes such as existing PSM and its combined modulation schemes.

Despite many advantages, OOK-PSM scheme cannot support higher level modulation schemes in multipath scenarios. Therefore, a combined modulation is proposed for higher level modulation schemes in the presence of multiple users and multipath channel. The proposed scheme is a combination of orthogonal PPM (OPPM) and biorthogonal PSM (BPSM). It allows increased number of bits per symbol and reduces duration of the pulse repetition interval by increasing the number of orthogonal pulses. To show data rate efficiency of this scheme, the channel capacity is analyzed in the presence of multipath channel. The proposed transmission scheme is investigated through mathematical analysis and simulation results in the presence of multiple users and multipath channel. The BER performance and data rate of the system depends on the number of pulse positions and orthogonal pulses. This multivariate modulation scheme can be used adaptively for multi mode data rate TH-UWB systems.

The other research target of this work is to reduce UWB interference on NB systems. Due to wide bandwidth of the transmitted signal, UWB signal energy spreads over large frequency bands which are allocated to other NB radio systems, thus NB systems are affected by UWB interference [32]. To show coexistence capability of orthogonal and non orthogonal pulse based UWB systems with other NB receiver, PSD of these modulation schemes have been analyzed. First, PSD of orthogonal pulse based modulation schemes such as PSM, BPSM and OPPM-BPSM have been studied to show the power efficiency orthogonal pulses. Finally, a set of generalized pulse waveform is proposed to reduce UWB interference power on 802.11a receivers. The reduction of UWB interference on IEEE802.11a system is done by creating nulls at the center frequency of 802.11a receivers. The proposed pulse waveform can also reduce UWB interference power for any other NB

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systems by shifting the null point into the center frequency of the desired NB system. In addition, it is observed that the proposed waveform satisfies the FCC requirements without using the carrier signal.

1.5 Preview of the Thesis

The remainder of this dissertation is organized as follows:

Chapter 2 discusses some basic information on UWB pulse waveforms such as Gaussian pulses, modified Hermite pulses (MHPs), and Prolate spheroidal wave functions (PSWFs). Non orthogonal pulse based modulation schemes, orthogonal pulse based modulation schemes and combination of them are presented in detail. The UWB multipath channel model is investigated briefly. The analysis of system capacity and PSD of TH-UWB system are presented. Finally, UWB interference issues on NB system are presented.

Chapter 3 provides a combined scheme for OOK and PSM modulation for low power and low cost system design. It covers interference issues such as IPI, MPI and MAI in orthogonal pulse scenario. The performance of TH-UWB systems in the presence of multiuser and multipath channel model is analyzed. Finally, simulation results based on MHPs and PSWFs are presented under AWGN and multipath channel.

Chapter 4 presents a combined scheme using orthogonal PPM and BPSM for higher level modulation scheme with high data rates. The performance analysis with the assumption of correlation and RAKE receiver are presented in detail. The channel capacity of this scheme is also presented. Finally, a tradeoff between the number of pulse positions and orthogonal pulses is presented based on the system performance and data rate.

Chapter 5 describes UWB interference issues on NB systems. The PSD of orthogonal pulse based modulation schemes is analyzed to show the power efficiency of orthogonal pulses. A set of generalized pulse waveform of Gaussian family is analyzed to reduce the

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UWB interference power on NB system. This chapter also provides the special feature of this generalized pulse waveform such as null spectral creation characteristic and FCC spectral requirements. The performance of 802.11a receiver in the presence of UWB systems is presented through mathematical analysis as well as simulation studies.

Chapter 6 provides overall conclusion of this report. The possible future research directions in this area are also enumerated.

Chapter 2

Ultra Wideband Radio Systems: A Review

2.1 Introduction

The purpose of this chapter is to provide a general overview of IR-UWB systems, specifically of TH-UWB radio systems in terms of pulse waveforms, modulation techniques, channel models, transceiver design, system capacity, PSD and TH-UWB interference on NB systems.

This chapter is organized as follows: section 2.2 describes non orthogonal and orthogonal pulses for TH-UWB systems. Section 2.3 describes modulation schemes based on non orthogonal pulses. Section 2.4 describes modulation schemes based on orthogonal pulses and its combined form with other conventional modulation schemes. Section 2.5 presents a system model of TH-UWB systems which includes a transmitter, a channel and a receiver. Section 2.6 provides capacity of TH-UWB system for AWGN and multipath channel models. Section 2.7 describes PSD of the TH-UWB systems for different modulation schemes. Finally, TH-UWB interference on NB system is discussed in section 2.8 followed by summary of this chapter in section 2.9.

2.2 Pulses for TH-UWB System

One of the essential functions in TH-UWB systems is the representation of a message symbol by a short duration pulse waveform for signal transmission through air [33, 34]. The pulse waveform is an important design consideration which can affect UWB system performance considerably. The successful deployment of high data rate indoor TH-UWB systems strongly depends on the development of pulse waveforms and modulation schemes. Because of the short pulse waveforms, UWB is capable of providing high data rates for short range wireless communication. Two types of pulses are used for TH-UWB systems. They are non orthogonal pulses and orthogonal pulses [35, 36].

2.2.1 Non orthogonal Pulses

For UWB radio systems with non orthogonal pulses, information is modulated by using either different shift positions or different amplitudes of pulses [35]. The shape of these pulses can be easily maintained in the transceiver. The efficiency of these pulses depends on the shape in time domain, shape in frequency domain and shape of auto-correlation functions. The conventional TH-UWB systems use Gaussian pulse family, Rayleigh monopulse, Laplacian pulse, cubic monopulse and pulses based on B-spline approximation [37, 38].

Gaussian pulse and its derivatives are useful candidates for the conventional TH-UWB systems. To be radiated in an efficient way, Gaussian pulse should have a zero DC component and zero offset. The output of the transmitter antennae can be modeled as the first order derivative of the transmitted Gaussian pulse. The general Gaussian pulse can be expressed as [12, 37]

$$y(t) = \frac{A}{\sqrt{2\pi}\sigma} \exp\left(-\frac{t^2}{2\sigma^2}\right) \quad (2.1)$$

where A is the amplitude of the pulse waveform and σ is the bandwidth scaling factor. The 1st and 2nd order derivatives of Eq. (2.1) are known as Gaussian monocycle and

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Gaussian doublet respectively [39]. The n^{th} order derivative of Eq. (2.1) can be expressed as

$$y^n(t) = -\frac{n-1}{\sigma^2}y^{n-2}(t) - \frac{t}{\sigma^2}y^{n-1}(t). \quad (2.2)$$

The frequency domain representation of Eq. (2.2) can be expressed as

$$Y(f) = A(j2\pi f)^n \exp\left(-\frac{(2\pi f\sigma)^2}{2}\right). \quad (2.3)$$

In order to satisfy the FCC spectral mask for lower order derivatives of Gaussian pulse, the scaling factor σ needs to be chosen properly. A carrier signal is also used to satisfy the spectral mask. However, higher order derivatives of Gaussian pulse satisfy the FCC requirements without using a carrier signal [37].

The other Gaussian doublet is defined by Hamalainen which has the same length and reversed amplitudes with a fixed time gap between the pulses [40,41]. It can be expressed as

$$w(t) = y(t) - y(t - t_w) \quad (2.4)$$

where t_w is the time interval between two pulses. The time and frequency domains representation of these pulses are shown in Figs. 2.1 and 2.2.

Another useful pulse for TH-UWB systems is Rayleigh monopulse which has similar spectral shapes and time duration as Gaussian monopulse. Similarly, the Laplacian and cubic monopulse are also used for TH-UWB system. These pulses are even and odd in time, and their spectral characteristics are easily distinguishable. Pulses based on B-spline are sometime used for TH-UWB systems and converge from time-limited to band-limited signal using constrained approximation technique [35,38].

Although several pulses are proposed for TH-UWB signals, mostly Gaussian pulses and its family are used in practice due to their simplicity and easiness in implementation within short interval. Gaussian pulse and its family spread the energy of the signal from near DC to a few GHz. The higher order Gaussian pulses easily comply with FCC

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Figure 2.1: Time domain representation of Gaussian pulse, Gaussian monopulse, Gaussian doublet and doublet defined by Hamalainen

Figure 2.2: Frequency domain representation of Gaussian pulse, Gaussian monopulse, Gaussian doublet and doublet defined by Hamalainen

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regulation whereas Laplacian and cubic monopulse do not satisfy the FCC spectral mask without the carrier signal. Generating B-spline pulses are rather complex. Moreover, its base band signal is not compatible with FCC mask.

2.2.2 Orthogonal Pulses

For UWB radio systems with orthogonal pulses, information is modulated by different orthogonal pulses instead of positions and amplitudes [36]. These pulses can also be used for multiple access (MA) transmission where every user assigns different order pulses [42, 43]. The orthogonal pulses are more efficient when normalized autocorrelation gives *one* and crosscorrelation gives *zero*. Sometimes, autocorrelation and crosscorrelation are referred to as self-correlation and inter-correlation respectively. Various sets of orthogonal pulses such as modified Hermite pulses (MHPs) and Prolate spheroidal wave functions (PSWFs) have been proposed for TH-UWB systems to improve system performance [30]. There are other sets of orthogonal pulses such as Battle-Lemarie wavelet orthogonal function, Haar wavelet orthogonal functions, Laguerre orthogonal functions and Bessel-Chebyshev polynomials [44–46]. However, these orthogonal pulses are not widely used for TH-UWB modulations purpose. Therefore, their discussions are not included in this dissertation.

2.2.2.1 Modified Hermite Pulses

Mostly used orthogonal pulses for TH-UWB system are a set of MHPs [47]. The MHPs are modified from Hermite polynomials [48]. The use of MHPs for UWB system was proposed in [36, 49, 50]. The main advantage of MHPs is that they are mutually orthogonal. The time domain representation of n^{th} order MHPs can be expressed as

$$w_n(t) = k_n e^{-\frac{t^2}{4\eta^2}} n! \sum_{i=0}^{\lfloor \frac{n}{2} \rfloor} \left(-\frac{1}{2}\right)^i \frac{\left(\frac{t}{\eta}\right)^{n-2i}}{(n-2i)!i!}. \quad (2.5)$$

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If it is desired that each $w_n(t)$ has an energy of E_n , then the normalization factor k_n can be written as

$$k_n = \sqrt{\frac{E_n}{\eta n! \sqrt{2\pi}}} \quad (2.6)$$

where $\eta (= 5 \times 10^{-11} \text{ sec})$ is the scaling factor which insures that pulses have unit energy. Fig. 2.3 shows the time domain representation of MHPs of order 0, 1, 2 and 3. A constant width should be maintained for all pulses to reduce the complexity of pulse generator. However, selecting a narrower pulse width introduces timing jitter error for higher order pulses [34, 51, 52].

The frequency domain representation of MHPs of order 0, 1, 2 and 3 can be expressed as

$$\begin{aligned} H_0(f) &= 2\sqrt{\pi} \exp(-4\pi^2 f^2) \\ H_1(f) &= (-j4\pi f) 2\sqrt{\pi} \exp(-4\pi^2 f^2) \\ H_2(f) &= (1 - 16\pi^2 f^2) 2\sqrt{\pi} \exp(-4\pi^2 f^2) \\ H_3(f) &= (-j12\pi f + j64\pi^3 f^3) 2\sqrt{\pi} \exp(-4\pi^2 f^2). \end{aligned} \quad (2.7)$$

Fig. 2.4 shows the frequency domain representation of MHPs. It is observed that all the pulses have zero DC component, which is the basic idea of TH-UWB systems.

The autocorrelation functions of n^{th} order MHPs are defined as [53, 54]

$$\begin{aligned} R(\eta, n) &= (n!)^2 \exp\left(-\frac{\eta^2}{2}\right) \sum_{k=0}^{[n/2]} \sum_{l=0}^{[n/2]} \frac{(-1)^{k+l} 2^{-k-l}}{k! l! (n-2k)! (n-2l)!} \\ &\times \sum_{\substack{u=0 \\ u+v \text{ even}}}^{n-2k} \sum_{\substack{v=0 \\ v \text{ even}}}^{n-2l} \binom{n-2k}{u} \binom{n-2l}{v} \\ &\times \eta^{2n-2k-2l-u-v} (-1)^{n-v} 2^{(u+v+1)/2} \Gamma\left(\frac{u+v+1}{2}\right). \end{aligned} \quad (2.8)$$

The autocorrelation functions of the MHPs for orders $n = 0, 1, 2$ and 3 are shown in Fig. 2.5. It can be seen that the width of the main peak in the autocorrelation functions

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Figure 2.3: Time response of normalized MHPs of order $n = 0, 1, 2, 3$

Figure 2.4: The frequency response of normalized MHPs of order $n = 0, 1, 2, 3$

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becomes narrower with increase in the order of pulses. This suggests that higher order pulses are more sensitive to pulse jitter [34, 50]. Hence, it is better to use lower order pulses as much as possible. Despite its disadvantages, MHPs can be used for TH-UWB communication for almost the same duration and the same bandwidth for all values of n . The number of zero crossing of pulses is equal to order of pulses which determines the complexity of the pulse generators.

2.2.2.2 Prolate Spheroidal Wave Function

The other commonly used orthogonal pulses are PSWFs. They follow the same pattern as MHPs. However, PSWFs have applications in different fields. These are solutions of a second order differential equation [55, 56]

$$\frac{d}{dt} \left[(1 - t^2) \frac{d\psi(t)}{dt} \right] + (\lambda - c^2 t^2) \psi(t) = 0 \quad (2.9)$$

where $\psi(t)$ are PSWFs, λ is the energy concentration of $\psi(t)$ that lies in the time interval $[-T, T]$, and c denotes the number of degrees of freedom. If the bandwidth of $\psi(t)$ is Ω , then $c = T\Omega$. The closed-form solution of Eq. (2.9) is difficult to find. The approximate solution of PSWFs can be expressed as

$$\psi(t) \approx \frac{\sin \left[c \sqrt{\left(\frac{t}{T}\right)^2 - 1} \right]}{\sinh(c) \sqrt{\left(\frac{t}{T}\right)^2 - 1}}. \quad (2.10)$$

Recently, some numerical solutions have been presented for short duration pulse to meet the power spectral constraint of FCC UWB mask [57]. Fig. 2.6 shows the time domain representation of PSWFs of order 0, 1, 2 and 3. From Fig. 2.6, it is observed that the width of pulses are nearly same for all order of pulses, similar to that of MHPs.

The eigenvector of PSWFs plays an important role for power efficiency. The greater the eigenvalues, the better the power spectrum fit. Therefore, only eigenvectors corresponding to large eigenvalues should be taken for pulse design and system implementation. The two strong eigenvector pulses have many advantages compared to MHPs.

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Figure 2.5: Autocorrelation function of normalized MHPs of order $n = 0, 1, 2, 3$.

Figure 2.6: The normalized PSWFs of order 1, 2, 3 and 4

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The duration of PSWFs is less than the duration of MHPs of order 3 or higher, resulting in higher data rates. The main difference of PSWFs from MHPs is its better orthogonalization characteristics [58].

2.3 Modulation Schemes Based on non orthogonal Pulses

Modulation can be defined as a process of encoding information from a message source in a manner which is suitable for transmission. Various kinds of TH-UWB modulation schemes have been investigated by using non orthogonal pulses to meet system requirements for different applications [10,11]. Conventional modulation schemes such as PPM, PAM and combination of them have been proposed for non orthogonal pulses for improving system performance [59,60]. The efficiency of these modulation schemes is determined from different aspects such as data rates, transceiver complexity, system capacity, spectral characteristics, multiple access capability, ISI, BER performance, robustness against coexisting NB systems and so on [61–63].

2.3.1 Pulse Position Modulation

PPM is a widely used modulation scheme for TH-UWB system. The data are modulated by time shifting from nominal position [64]. If data is 0, pulse takes the nominal position and if data is 1 pulse takes position shifted by δ . When the delay is less than the pulse width, the modulation is called overlapping PPM or simply PPM and when the delay is larger than the pulse width, the modulation is called orthogonal PPM (OPPM) [63,65].

The advantages of PPM mainly arises from its simplicity and ease with which the delay may be controlled. In the case of M -ary PPM schemes, the level of modulation schemes can be extended by simply increasing the number of pulse positions. Since the M -ary PPM symbols are orthogonal, obtained by maintaining their Euclidean distance,

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symbols become more power efficient when compared to other modulation schemes for higher values of M . PPM is defined by a time offset and hence it is suited to random dithering in time. M -ary PPM uses constant amplitude for all data. The probability of detecting a false data bit is reduced compared to that in amplitude based modulation schemes [18, 66].

Despite advantages of PPM scheme, extremely fine time control is necessary to modulate pulses for subnanosecond accuracy. It needs coherent energy detector receiver and perfect knowledge in delay attenuation and fading to provide perfect template signals that resemble the received pulse [1, 67, 68].

2.3.2 Pulse Amplitude Modulation

PAM is another popular modulation scheme for TH-UWB systems. It encodes data bits by employing different levels of power without changing the pulse position. For lower level modulation schemes PAM has higher power efficiency than that in PPM scheme but for higher level modulation schemes PAM becomes worse because of decreased Euclidean distances among symbols [69].

Because of the periodicity of transmitted pulses, discrete lines are present in the PSD of M -ary PAM pulses which is discussed in section 2.7. These discrete lines increase the interference to other NB or wideband systems which share the same frequency spectrum with TH-UWB systems [1]. Details of interference issues with PSD and modulation schemes are given in section 2.8 and chapter 5.

Generally, M -ary PAM cannot be used in short-range/lowpower wireless communications for higher values of M . The major reason includes the fact that an amplitude-modulated signal which has smaller amplitude is more susceptible to noise interference than its larger amplitude counterparts [70]. Moreover, more power is required for higher level modulation schemes.

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2.3.2.1 On-Off Keying

OOK modulation is a special case of PAM where 1 is represented by the presence of pulse waveform and 0 is represented by the absence of pulse waveform. Its main advantages are simplicity and low implementation cost. The OOK transmitter is designed by turning on or off of radio frequency (RF) switch for representing data. It gives similar performance as PPM scheme when the average transmitted energy is the same [18]. The demodulation can be done by using both coherent and noncoherent detector receivers.

The major problem of OOK is that it is highly sensitive to noise and interference in the presence of multipath, so an unwanted signal can be detected as a false data bit 1 [1]. It cannot be used for M -ary modulation schemes like PPM and PAM. The synchronization problem becomes more challenging task for TH-UWB systems when a long sequence of zeroes is transmitted. However, if the receiver complexity is the main design concern, OOK scheme is the perfect choice for TH-UWB systems [61].

2.3.2.2 Bi-phase Modulation

Bi-phase modulation (BPM) is another special case of PAM. It uses polarity of a pulse to represent information bits. For instance, a pulse with positive polarity represents a digital bit 1, and a pulse with negative polarity corresponds to a bit 0. Since it has larger Euclidean distance, it gives better performance compared to other TH-UWB modulation schemes [18]. Another advantage of BPM is that the change in polarity can remove discrete lines in PSD of the signal which increases the power efficiency of UWB signals and increases the coexistence capabilities to other NB systems. It has 3 dB higher power than both OOK and orthogonal PPM system in an AWGN channel [69].

However, BPM scheme increases the system complexity for the physical implementation of the transmitter. It requires two transmitters, one is to generate positive pulse and other is to generate negative pulse [1]. Moreover, for a stream of data, accurate

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timing between two transmitters is of great importance. It can be used only for binary modulation scheme and require coherent energy detector receiver.

2.4 Modulation Schemes Based on Orthogonal Pulses

Modulation schemes based on orthogonal pulses are becoming an interesting research topic in TH-UWB systems in recent days. Since pulse waveforms are orthogonal, constellation points of these schemes take place in different Euclidean planes. The Euclidean distances among symbols remain the same with increase in the level of modulation schemes. Thereby, this modulation scheme is more power efficient than other conventional M -ary modulation schemes such as M -ary PPM, M -ary PAM and their combinations. In addition, orthogonal pulse based modulation schemes reduce ISI and MAI for TH-UWB systems [31, 43, 71]. Current research in this direction includes PSM modulation schemes and many other combined forms of PSM modulation such as biorthogonal PSM, 2PPM-PSM and BPSK-PSM schemes [28–30]. Details of these modulation schemes are described in the following subsections:

2.4.1 Pulse Shape Modulation

Pulse shape modulation (PSM) scheme is used in TH-UWB system to represent different symbols by different pulse waveforms instead of different positions or amplitudes [47]. PSM uses a set of orthogonal pulses such as MHPs and PSWFs, as analog pulse waveforms to represent information [30, 56]. It is more useful in M -ary signaling systems. Sometimes it is also referred to as orthogonal pulse modulation (OPM) [36].

The orthogonality of pulses allows transmission of multiple orthogonal signals simultaneously in both time and frequency domains without any interference [72]. For example, for 4-ary PSM scheme, 00 is transmitted by 0^{th} order pulse, 01 is transmitted by assigning 1^{st} order pulse, 10 is transmitted by assigning 2^{nd} order pulse and 11 is transmitted by

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assigning 3^d order pulse. The order of pulses is selected from the same set of orthogonal pulses. In the receiver, these four different order pulses can be identified by using four different template signals which are designed based on the pulses that are used in the transmitter.

M -ary PSM scheme requires M orthogonal pulses in the transmitter and M template signals in the receiver. Physically, M orthogonal pulses are necessary to transmit M symbols. However, only the 0^{th} order pulse is sufficient to generate all other order pulses, which reduces the hardware system complexity in the transceiver [36]. Due to orthogonality of pulses, the PSM scheme reduces ISI and MAI [71]. However, it requires higher order pulses for higher level modulation schemes, which reduces the system performance due to narrower autocorrelation function of higher order pulses. On the other hand, it is undesirable to represent the same amount of information by using larger number of orthogonal pulses. This mechanism increases the number and/or magnitude of the spectral lines in the PSD of the signal.

2.4.1.1 Biorthogonal PSM

Biorthogonal PSM (BPSM) modulation is similar to BPM scheme, the only difference is that it is used for M -ary signaling. M -ary BPSM uses $M/2$ orthogonal pulses in the transmitter and $M/2$ correlators in the receiver to transmit all M possible symbols. The number of correlators or matched filters used in this scheme drops to half of those used in M -ary PSM scheme, thus reducing complexity of TH-UWB systems. An M -ary BPSM modulation has been proposed in [28]. The output of an M -ary BPSM can be a signal with $M/2$ possible pulse shapes which are biorthogonal. Orthogonal pulse shapes are represented as follows: $w_0(t), w_1(t), \dots, w_{M/2-1}(t)$. The negative ones are defined as $w_{i+M/2}(t) = -w_i(t)$, where $i = 0, 1, \dots, M/2 - 1$. The corresponding M -ary BPSM signal based on MHPs is shown in Table 2.1 for $M = 4$ and 8.

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Table 2.1: M -ary communication for biorthogonal modulation scheme

M -ary	Symbol	Hermite Pulses
binary	0,1	$w_0(t), -w_0(t)$
4-ary	00,01,10,11	$w_0(t), -w_0(t), w_1(t), -w_1(t)$
8-ary	000,001,010,011, 100,101,110,111	$w_0(t), -w_0(t), w_1(t), -w_1(t),$ $w_2(t), -w_2(t), w_3(t), -w_3(t)$

BPSM gives high data rate and makes it easier to map symbols into pulse waveforms. It has high power efficiency due to pulse polarity. However, similar to BPM scheme, it also requires two transmitters to generate BPSM signal. Maintaining bi-phase of orthogonal pulses is a challenging task. On the other hand, due to limitation of the possible number of orthogonal pulses it cannot be used for higher level modulation schemes.

2.4.1.2 Combination of PPM and PSM

Combined PPM and PSM (2PPM-PSM) is a modulation scheme which uses both pulse position and pulse shape modulation to transmit more information bits in the same time slot by using the same number of pulses than conventional schemes [29]. The corresponding 4-ary and 8-ary modulation schemes are shown in Table 2.2, where T represents the nominal time position and δ represents time shift from the nominal position. Let A and B are orthogonal pulses and coefficients of A and B represent amplitude of pulses.

2PPM-PSM scheme uses a set of MHPs. However, any mutually orthogonal pulse can be used for such a system. From the Table 2.2, one can observe that a hardware structure designed for an 8-ary system is downward compatible for a 4-ary system. This scheme can be useful for implementation in an adaptive modulation system for the different channel condition at any given instant. Since it uses fewer orthogonal pulses for the same amount of information, it is desirable to reduce the number of pulses and/or magnitude of the spectral lines in PSD.

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Table 2.2: 4-ary and 8-ary modulation scheme for combined PPM and PSM

4-ary	T_f	
	T	$T+\delta$
00	A+B	0
01	A	B
10	B	A
11	0	A+B
8-ary		
000	2A	0
001	0	2A
010	2B	0
011	0	2B
100	A	A
101	A+B	0
110	B	B
111	0	A+B

However, the performance of the system suffers from nonuniform distribution of the constellation points. Therefore, this scheme needs orthogonalized and coded modulation to improve system performance [73]. In addition, it needs map decision and lookup table to retrieve the isolated modulation alphabet. This process requires memory in the receiver to store the map decision and lookup tables. From Table 2.2 it can be seen that for symbols 100 and 110, same pulse waveforms are transmitted in same pulse repetition interval. Since the pulse shift position is not greater than the width of the pulse waveform, the performance of the system degrades due to self interference between pulses for symbols 100 and 110. Moreover, it requires a threshold value to identify the level of amplitude at different positions for different orthogonal pulses, which increases the system complexity.

2.4.1.3 BPSK-PSM Scheme

Combined BPSK-PSM scheme has been proposed for M -ary scheme based on PSWFs for achieving high data rates in multipath fading environment. However, it can be imple-

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mented by using any set of orthogonal pulses. BPSK-PSM transmits several orthogonal pulses at the same pulse repetition interval for a symbol. The number of transmitted pulses and the number of bits in a symbol are the same. Positive amplitude is selected for bit 1 and negative amplitude is selected for bit 0. Each bit is detected independently by different correlators and combines them to form a symbol. It reduces the orthogonal pulses and correlators in the transceiver than those used in PSM and BPSM schemes. The rest of the advantages and disadvantages are similar as BPM modulation scheme.

2.5 System Model of TH-UWB Radio Systems

The block diagram of a TH-UWB system is shown in Fig. 2.7. It consists of three major parts: transmitter, channel and receiver. Although the transmitter and receiver consist of many components such as encoders, interleavers, etc. to achieve better system performance, these are not taken into consideration within the scope of present research and are omitted from our dissertation [21, 69, 74].

2.5.1 Transmitter Model of TH-UWB Radio Systems

Transmitter modulates a bit stream into a train of output pulses. It is allocated a time slot where users are allowed to transmit a burst of data symbols (scrambled and coded) into the channel. Generally in TH-UWB systems, a transmitter emits sequences of pulses that are detected by corresponding receiver whose front-end amplifiers are synchronized and time-gated to the transmitted pulse sequences. The required data is modulated using the transmitted pulse with appropriate parameters. These parameters may include pulse position, amplitude, order of pulses and multiple access codes depend on the modulation schemes and number of users.

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Figure 2.7: Block diagram of a TH-UWB communication system

2.5.1.1 Single User Environment

In this discussion, TH-UWB transmitter is considered in single user environment to illustrate a simple transmitter structure. A set of generalized M -ary signals for non orthogonal pulse based modulation schemes is represented as $\{s_0(t), s_1(t), \dots, s_{M-1}(t)\}$, where $s_i(t)$, $(0 \leq i \leq M-1)$, is the signal for i^{th} symbol and can be expressed as

$$s_i(t) = \sqrt{E_{tx}} \sum_j a_i w(t - jT_f - \delta_i) \quad (2.11)$$

where for PPM scheme $a_i=1$, δ_i is the additional time shift from nominal position for the i^{th} symbols in the M -ary PPM scheme and $\delta_0=0$ indicates the nominal position with $\delta_1 < \delta_2 < \dots < \delta_{M-1} < T_f$. For M -ary PAM scheme, $a_i=(2i-1-M)$ and $\delta_i=0$. For the BPM scheme, $a_i \in \{\pm 1\}$ and $\delta_i=0$. Similarly, for OOK scheme, $a_i \in \{0, 1\}$ and $\delta_i=0$. In general, $w(t)$ is energy-normalized, $\int_0^{T_f} w(t)^2 dt = 1$, pulse waveform. E_{tx} is the transmitted energy over each signal pulse, T_f is the pulse repetition interval. For a fixed T_f , the symbol rate can be evaluated by $R_b = 1/(N_s T_f)$ where N_s is the number of pulse repetition interval for a symbol. The symbol duration is then calculated by $T_b = N_s T_f$.

Several transmission models based on orthogonal pulses are available in the literature [36, 71, 75]. However, only PSM based transmission scheme is included in this discussion. The M -ary PSM signal based on orthogonal pulses for the i^{th} symbol can be expressed as

$$s_i(t) = \sqrt{E_{tx}} \sum_j w_i(t - jT_f) \quad (2.12)$$

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where $w_i(t)$ is the i^{th} order orthogonal pulse. The performance difference of non orthogonal and orthogonal pulse based transmission schemes can be explained in many aspects, such as complexity, data rate timing error sensitivity, receiver gain and several interference parameters such as MAI, ISI, IPI and MPI. In order to make the transmitter structure simpler, orthogonal pulse based transmission scheme assumes pulse width of each order pulse as the same. However, it decreases system performance in the presence of timing jitter [76].

2.5.1.2 Multiple Users Environment

Most communication systems have a requirement to support multiple users where different users are allowed to share the same physical medium to transmit and receive different data flows at the same time. Multiple access communication in TH-UWB system has drawn significant research interest. Various multiple access schemes for UWB system and their performance have been reported in literature [15, 64, 74]. TH has been found to be a good multiple access technique for TH-UWB systems [77]. In TH-UWB systems, each user assigned the entire spectrum in one time slot. Therefore, TH-UWB provides multiple access for different users by allowing to share the same physical medium. Since the emitted energy spreads in the frequency domain it is fair to say that TH-UWB systems can be characterized as an extension of traditional spread spectrum schemes.

Multiple access can be implemented by choosing appropriate time hopping sequences for different users to minimize the probability of collision. Since multiple systems share the same medium, each user is provided with a specific TH code sequence to avoid catastrophic collision at the receiver. The addressable TH duration must be strictly less than the frame time. The sequence is known as pseudorandom PN codes [78, 79]. The same sequence serves as user's signatures and ensures to access the medium to multiple users. Each code modifies the transmitted signal in such a way that the receiver is

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capable of separating a useful signal from other user's signal. The other user's signal is seen as interfering signal at the desired receiver. The possibility of removing these unnecessary contributions depends mainly on the characteristics of codes which are used for separating the transmitted data flow.

In TH-UWB system, each frame is subdivided into N_h chips of duration T_c . Each user (indexed by k) is assigned a unique pseudorandom time shift pattern, $c_j^{(k)}$, called a TH sequence, which provides an additional time shift for each pulse in the pulse train. Each TH code is a sequence of N_h independent and identically distributed random variables, with probability $1/N_h$ of assuming one of the integer values in the range $[0, N_h - 1]$. The k^{th} pulse undergoes an additional shift of $c_j^{(k)}T_c$, where chip duration T_c is also the duration of an addressable time delay bin [12]. An example of TH-PPM modulation scheme in a multiple users environment is shown in Fig. 2.8. The M -ary TH-UWB signal corresponding to k^{th} ($k = 1, \dots, N_u$) user can be written in a general mathematical form as

$$s_i^{(k)}(t) = \sqrt{E_{tx}^{(k)}} \sum_j a_i w(t - jT_f - c_j^{(k)}T_c - \delta_i^{(k)}) \quad (2.13)$$

and the corresponding M -ary PSM signal can be expressed as

$$s_i^{(k)}(t) = \sqrt{E_{tx}^{(k)}} \sum_j w_i(t - jT_f - c_j^{(k)}T_c). \quad (2.14)$$

2.5.2 Channel Models

A propagation environment where a signal passes through from a transmitter to a receiver is referred to as the channel. The emergence of TH-UWB technology implies that understanding characteristics of the TH-UWB propagation channel is essential. Accurate channel models are vital for performance evaluation and design of TH-UWB systems [17, 31, 80, 81].

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Figure 2.8: An example of TH-PPM scheme for multi-user environment

Many important aspects of TH-UWB systems have not yet been thoroughly investigated. These reported measurements hardly coincide with each other, especially in the amplitude distribution of multipath signals. This is partly because the propagation of TH-UWB signals is susceptible to the environment and the accuracy of UWB modeling is highly dependent of the setup of measuring equipments. If the channel is well characterized, effect of disturbances and other sources of perturbation can be reduced by proper design of the transmitter and receiver [50, 82, 83]. Detailed characterization of UWB radio propagation with focus on the small-scale fading are described below.

2.5.2.1 AWGN Channel Model

A AWGN channel model is one in which the only impairment is the linear addition of wideband or white noise with a constant spectral density. It does not account for the phenomena of fading, frequency selectivity, interference, nonlinearity or dispersion. However, it produces simple, tractable mathematical model which is useful for gaining insight into the underlying behavior of a system before the above phenomena are considered.

For UWB systems, channel $n(t)$ is a statistical channel that has a flat power spectral density over the entire frequency range and whose samples are Gaussian distributed with zero mean and variance (power) $\sigma_0^2 = 1/SNR$, where SNR is the average signal to noise ratio. It is assumed that UWB channel is wide-sense stationary, i.e., the autocorrelation

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function of $n(t)$ only depends on the time difference and not on the time epoch t . The correlation receiver is used to detect the information under the assumption of AWGN channel. However, optimum receiver for the AWGN channel may not be an appropriate scheme in a multipath environment.

2.5.2.2 Multipath Channel Model

UWB signal is wideband in nature i.e. up to tens of GHz of frequency bandwidth. So it can offer extremely fine time resolution in dense multipath environment. However, number of paths, fading distribution, phase characteristic and multipath delay spread are major problems to derive a complete statistical model of the TH-UWB channel [80, 82]. The IEEE802.15.3a working group proposed a channel model which is a modified version of Saleh-Valenzuela channel model [31, 84, 85].

IEEE802.15.3a channel model is evaluated using the UWB multipath channel model based on the indoor channel measurement in the 2 – 8 GHz frequency band. Each multipath belongs to a cluster, which is a group of multipaths that arrive close together in time. The arrival times of each cluster and each multipath within a cluster are exponentially distributed random variable conditioned upon the previous arrival time. Each cluster, as well as each multipath within a cluster undergoes independent fading. The multipath components arrive in clusters and the amplitude of every multipath gain has a Log-normal distribution rather than the Rayleigh distribution. The phase is randomly set to ± 1 with equiprobability. The multipath channel can be classified by its impulse response as

$$h(t) = \sum_{l=0}^{L-1} \sum_{m=0}^{M-1} \alpha_{m,l} \delta(t - T_l - \tau_{m,l}) \quad (2.15)$$

where $\delta(\cdot)$ is the Dirac delta function, $\alpha_{m,l}$ is the multipath gain coefficients for m^{th} ray and l^{th} cluster, T_l is the delay of the l^{th} cluster and $\tau_{m,l}$ is the delay of the m^{th} multipath component relative to the l^{th} cluster arrival time T_l . Multipath channel coefficients are

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defined as $\alpha_{m,l} = p_{m,l}\beta_{m,l}$ where $p_{m,l}$ is equal likely to take on the values of ± 1 , and $\beta_{m,l}$ is the Log-normal fading [86, 87], the statistics of $\beta_{m,l}$ are as follows:

$$20 \log(\beta_{m,l}) \propto \mathcal{N}(\mu_{m,l}, \sigma^2) \quad (2.16)$$

$$E[\beta_{m,l}^2] = \Omega_0 e^{-T_l/\Gamma} e^{-\tau_{m,l}/\gamma} \quad (2.17)$$

where Ω_0 is the mean power of the first path of the first cluster, and distributions of cluster arrival time and the ray arrival time are given by

$$p(T_l/T_{l-1}) = \Lambda \exp[-\Lambda(T_l - T_{l-1})], \quad l > 0 \quad (2.18)$$

and

$$p(\tau_{m,l}/\tau_{(m-1),l}) = \lambda \exp[-\lambda(\tau_{m,l} - \tau_{(m-1),l})], \quad m > 0 \quad (2.19)$$

respectively, where T_l is the arrival time of the first path of the l^{th} cluster; $\tau_{m,l}$ is the delay of the m^{th} path within l^{th} cluster relative to the first path arrival time, T_l , Λ is the cluster arrival rate, λ is the ray arrival rate, i.e. the arrival rate of path within each cluster, Γ is the cluster decay factor, γ is the ray decay factor and σ is the standard deviation of Log-normal fading in dB.

The channel model divides possible channel conditions according to four scenarios: channel model CM1 is for the delay of the direct path or quasi line-of-sight (LOS) within the distance of 0 – 4 m, channel CM2 is for non-line-of-sight (NLOS) within the distance of 0 – 4 m, channel CM3 is for LOS within the distance of 4 – 10 m, and channel CM4 is for NLOS within the distance of 4 – 10 m. Figs. 2.9 and 2.10 show the channel response for the channel CM1 and CM3 respectively. It is observed that channel realization of LOS channel is easier than NLOS version. Channel model parameters match well with UWB channel measurements with a typical rms delay spread of 15 – 20 ns. Channel coefficients are normalized to remove the effect of path loss for the UWB channel model. Many authors [80, 83, 84] have found different channel parameters in different office building

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Figure 2.9: Impulse response for the channel model for LOS proposed by IEEE802.15.3a working group

Figure 2.10: Impulse response for the channel model for NLOS proposed by IEEE802.15.3a working group

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environments. However, the present work has used parameters based on IEEE802.15.3a channel model [31]. For the purpose of performance analysis, the present channel model is simplified to the following expression

$$h(t) = \sum_{l=1}^L \alpha_l \delta(t - \tau_l) \quad (2.20)$$

where τ_l is the l^{th} multipath delay, α_l is the l^{th} multipath gain.

2.5.3 Receiver Model of TH-UWB Radio Systems

The common TH-UWB receiver includes threshold detectors, correlation detectors, RAKE-fingers and the advanced receiver which contains complex signal processing techniques. The receiver structure fully depends on propagation environments. Two basic receivers are considered in discussion namely, correlation receiver for the consideration of the AWGN channel and RAKE receiver for the multipath channel [8, 64, 88].

The typical received signal for number of N_u active users can be expressed in the presence of AWGN channel as

$$r(t) = \sum_{k=1}^{N_u} \alpha^{(k)} s^{(k)}(t - \tau^{(k)}) + n(t) \quad (2.21)$$

where $\alpha^{(k)}$ is the channel gain of the k^{th} user. The received signal in the presence of multipath channel is

$$r(t) = \sum_{k=1}^{N_u} \sum_{l=1}^L \alpha_l^{(k)} s^{(k)}(t - \tau^{(k)} - \tau_l^{(k)}) + n(t) \quad (2.22)$$

where $\alpha_l^{(k)}$ is the gain of the l^{th} path of k^{th} user, $\tau^{(k)}$ is the time asynchronism between 1^{st} and k^{th} users and $\tau_l^{(k)}$ is the l^{th} path delay of k^{th} user.

2.5.3.1 Correlation Receiver

Correlation receiver or matched filter is perhaps a simplest approach to the TH-UWB receiver, often referred to as an optimum receiver for a single user AWGN channel [89–91].

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The principle behind this receiver is that correlation is performed by mixing the received pulse with a template pulse waveform followed by an integrator and the output of the integrator is sampled to detect data. The integration interval is somewhat longer than the pulse length in order to pick up some of multipath components. In order to optimize the processing gain and SNR, the template waveform should be the same as that of received waveform. This is however, not easily achievable as the received signal is subjected to channel distortions. Thus, generating a template signal for the received pulse is difficult without considerable increase in complexity. The complexity of template generation can be avoided by approximating the template by using a delayed version of the transmitted pulse or by using a rectangular pulse. The simplest correlator receiver structure is shown in Fig. 2.11.

In each correlator, a template signal corresponding to a specific symbol is generated and convolved with the received signal to maximize the output of SNR. Output samples are passed to a decision circuit and maximum likelihood principle is applied to detect the transmitted data. The template signal of the m^{th} correlator of user can be expressed as

$$\phi_{1,m}(t) = \sum_{j=0}^{N_s-1} \alpha_j^{(1)} w(t - jT_f - c_j^{(1)}T_c - \delta_m^{(1)}) \quad (2.23)$$

and the corresponding output of the m^{th} correlator can be written as

$$Z = \int_{jT_f}^{(j+1)T_f} r(t) \phi_{1,m}(t) dt. \quad (2.24)$$

Main challenges in this approach are providing accurate synchronization and processing speed. The optimum receiver for AWGN channel may not be an optimal scheme for a multiple user and multipath environment. More advanced receiver are required for such applications.

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Figure 2.11: Correlator receiver structure

2.5.3.2 RAKE Receiver

In TH-UWB systems, a large number of multipath components are available for detecting the transmitted signal. These multipath components are resolvable and can be effectively combined to improve the performance of the receiver. In order to do that, the received signal passes through time delay elements, whose function is to align multipath components in time. These multipath components are passing through different correlator banks with correlator mask. Correlator banks are followed by a combiner that determines the variable to be used for the decision. The received signals at a correlator is used by a set of weighting functions that are used to combine the outputs of the correlator [92–94].

The most common way to collect multipath energy is to use a RAKE receiver. The RAKE receiver is a bank of correlators. It takes advantage of multipath propagation by combining a large number of different independent replicas of the same transmitted pulse, in order to exploit time diversity of the multipath channel. The output of all the fingers are combined using several methods such as maximal ratio combining (MRC), selection diversity (SD), absolute combining (AC) and equal gain combining (EGC). A large number of RAKE fingers is needed to capture all the multipath energy sufficiently, and this increases system complexity proportionally. Therefore, an optimum receiver

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under these conditions must be included for the number of correlators associated with different replicas of the same transmitted waveform [95, 96].

The implementation of a RAKE receiver requires the same transmitted pulse, and in order to exploit time diversity, it requires the knowledge of multipath gains and multipath delays, which are obtained by channel estimators. Receiver structure and its complexity depend on demodulation technique, channel response, number of correlators and number of RAKE fingers in the receiver. Fig. 2.12 shows the receiver structure for a multipath channel. The weighting factors $\omega_1, \omega_2, \dots, \omega_N$ are decided based on channel estimation, where N is the number of resolvable paths in the receiver.

Analytically, the template signal in the m^{th} correlator for m^{th} PPM symbol can be expressed as

$$\phi_m^{(1)}(t) = \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \alpha_p^{(1)} w'(t - jT_f - c_j^{(1)}T_c - \delta_m^{(1)} - \tau_l^{(1)}) \quad (2.25)$$

where $\delta_m^{(1)}$ is the time shift position for the m^{th} symbol of user one, $w'(t)$ is the possible retrieve pulse waveform in the receiver.

The number of fingers is selected in three ways namely: all RAKE (ARAKE), selected RAKE (SRAKE) and partial RAKE (PRAKE). The term ARAKE has been largely used in several works to indicate the receiver with unlimited resources (taps or correlators) and instant adaptability, so that it can combine all of the resolved multipath components in the receiver. The SRAKE receiver selects best paths and then combines the selected subset using MRC. It provides lower complexity than that of ARAKE receiver. However, the selection procedure requires knowledge of instantaneous values of all multipath components. The PRAKE receiver uses only first partial paths of resolving paths which are not necessarily the best. The complexity reduction with respect to SRAKE is due to the absence of a selection mechanism. It only needs to find the position of the first arriving path, which leads to a substantial complexity reduction [97, 98].

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Figure 2.12: RAKE receiver for demodulating PPM signal

2.6 Channel Capacity for TH-UWB Radio Systems

The capacity of a channel is the highest data rate that the channel can reliably support. The capacity of TH-UWB system depends on many factors such as modulation schemes, pulse waveforms, channel condition and the number of users. Capacity based on the modulation schemes such as M -ary PPM, M -ary PAM and combination of them have been provided in [99–103] under the assumptions of different channel conditions such as AWGN and multipath fading. Chapters 3 and 4 describe channel capacity of multipath channel for orthogonal pulses based modulation schemes.

2.6.1 Capacity in AWGN Channel

Capacity for M -ary OPPM signals over multipath channel can be evaluated by considering random variations of the received signal energy and signal correlation values. As per Shannon, the capacity of communication system in an AWGN channel with continuous-valued inputs and outputs as [104]

$$C = W \log_2(1 + SNR) \quad (2.26)$$

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where C is the capacity in bits/s and W is the channel bandwidth.

For discrete-valued input and continuous-valued output channel, a k -bit information source $\mathbf{U} = \{U_1, U_2, \dots, U_k\}$ is mapped to a $M = 2^k$ PPM signal set $\mathbf{X} = \{X_1, X_2, \dots, X_M\}$ where every signal $\mathbf{x}_m$ can be represented as an M -dimensional vector

$$\mathbf{x}_m = \{0, \dots, 0, \sqrt{E_s}, 0, \dots, 0\} \quad (2.27)$$

where E_s is the energy per dimensional symbol. Therefore, the channel capacity is the maximum amount of information that can be transmitted reliably and is given by

$$C = \max_{p(\mathbf{x})} I(\mathbf{Y}; \mathbf{X}) \quad (2.28)$$

where $I(\mathbf{Y}, \mathbf{U})$ is the mutual information between the channel output $\mathbf{Y}$ and the channel input $\mathbf{U}$. Since $\mathbf{X}$ is the invertible function of $\mathbf{U}$, the capacity becomes

$$C = \max_{p(\mathbf{x})} I(\mathbf{Y}; \mathbf{U}). \quad (2.29)$$

The channel capacity with input signals restricted to an equiprobable M -ary signal constellation, and no restriction on the channel output over the AWGN channel was computed as [105–107]

$$C = \log_2 M - E_{\mathbf{v}|\mathbf{x}_1} \log_2 \left[\frac{\sum_{j=1}^M p(\mathbf{v}|\mathbf{x}_j)}{p(\mathbf{v}|\mathbf{x}_1)} \right] \quad (2.30)$$

where $E_{\mathbf{v}|\mathbf{x}_1}$ is the expectation of $p(\mathbf{v}|\mathbf{x}_1)$, and $p(\mathbf{v}|\mathbf{x}_j)$ is the probability density function (PDF) conditioned on $\mathbf{x}_j$ as the transmitted signal and $\mathbf{v}_j$ as the received variable can be expressed as

$$p(\mathbf{v}|\mathbf{x}_j) = \left(\frac{1}{2\pi\sigma^2} \right)^{M/2} \exp \left[-\frac{(v_j - \sqrt{E_s})^2}{2\sigma^2} \right] \prod_{\substack{i=1 \\ i \neq j}}^M \exp \left[-\frac{v_i^2}{2\sigma^2} \right] \quad (2.31)$$

If $\mathbf{v}_j$, $j = 1, 2, \dots, M$ are conditionally independent for given $\mathbf{x}$, and distributed as

$$\begin{aligned} v_1 & \text{ is } \mathcal{N}(\sqrt{\rho}, 1) \\ v_j & \text{ is } \mathcal{N}(0, 1), \quad j \neq 1 \end{aligned} \quad (2.32)$$

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then by using Eq. (4.36) in Eq. (2.30) capacity can be expressed as

$$C = \log_2 M - E_{\mathbf{v}|\mathbf{x}_1} \log_2 \sum_{j=1}^M \exp[\sqrt{\rho}(v_j - v_1)] \quad (2.33)$$

It is a well known fact that the AWGN channel is fully characterized by its SNR and does not depend on an overall scaling of the noise. For large values of M the AWGN capacity can be approximated as

$$C = \log_2 M - E_{v_1|\mathbf{x}_1} \log_2 [1 + (M - 1)e^{\rho/2} e^{-\sqrt{\rho}v_1}] \quad (2.34)$$

where v_1 is $N(\sqrt{\rho}, 1)$ given x_1 .

2.6.2 Capacity in Multipath Fading Channel

Wireless communication systems typically operate in multipath fading channels. Without loss of generality, a multipath block fading channel is assumed for TH-UWB signal transmission [108, 109]. Specifically, fading is assumed to be constant during channel coherence time and becomes independent after coherence time. This assumption is valid for high data rate broadband wireless communication systems such as TH-UWB and DS-UWB systems. Here L_c RAKE fingers are assumed for MRC combining for the given TH-UWB system. Let $\mathbf{r}$ be the channel output vector corrupted by AWGN noise vector $\mathbf{w}$, which has zero mean and variance $\sigma^2 = \frac{1}{2} \|\mathbf{H}\|_F^2 N_0$, where $\|\mathbf{H}\|_F^2 = \sum_{i=1}^{L_c} \|\alpha_i\|^2$ is the squared Frobenius norm of $\mathbf{H}$, and α_i is the channel gain for the i^{th} diversity path. After MRC combining, $\mathbf{r}$ can be expressed as

$$\mathbf{r} = \|\mathbf{H}\|_F^2 \mathbf{x}_m + \mathbf{w}. \quad (2.35)$$

The capacity of M -ary PPM for TH-UWB systems over the multipath channel was computed in [104] as

$$C = \log_2 M - \frac{1}{M} \sum_{m=1}^M \int_{\mathbf{r}} p(\mathbf{r}|\mathbf{x}_m) \log_2 \left(\frac{\sum_{q=1}^M p(\mathbf{r}|\mathbf{x}_q)}{p(\mathbf{r}|\mathbf{x}_m)} \right) d\mathbf{r} \quad (2.36)$$

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where $\mathbf{r}$ is the M -dimensional received vector, $r_i = w_i$ for $i \neq m$ and $r_m = \|\mathbf{H}\|_F^2 \sqrt{E_s} + w_m$. The received signal $\mathbf{r}$ will have an M -dimensional joint Gaussian distribution conditioned on $\mathbf{x}_m$ with PDF

$$p(\mathbf{r}|\mathbf{x}_m) = \left(\frac{1}{2\pi\sigma^2} \right)^{M/2} \exp \left[- \frac{(r_m - \|\mathbf{H}\|_F^2 \sqrt{E_s})^2}{2\sigma^2} \right] \prod_{\substack{i=1 \\ i \neq j}}^M \exp \left[- \frac{r_i^2}{2\sigma^2} \right]. \quad (2.37)$$

Using Eq. (2.37) in Eq. (2.36), the channel capacity for an M -ary PPM UWB system on a block fading channel can be expressed as

$$C = \log_2 M - \frac{1}{M} \sum_{m=1}^M E_{\mathbf{r}|\mathbf{x}_m} \left[\log_2 \left(\sum_{q=1}^M e^{\left(- \frac{r_m^2 - r_q^2 + (r_q - \|\mathbf{H}\|_F \sqrt{E})^2 - (r_m - \|\mathbf{H}\|_F \sqrt{E})^2}{2\sigma^2} \right)} \right) \right]. \quad (2.38)$$

2.7 Power Spectral Density

Power spectral density (PSD) provides a single expression that gives the distribution of power over frequency and shows its sensitivity with respect to system parameters. The PSD analysis provides good insights into effects of periodicity and structure in signals. However, under anticipated regulations, UWB transmitted power is likely be limited by the PSD of the transmitted signal [37, 110].

The efficiency of modulation scheme is determined in two ways. First, the modulation technique needs to be power efficient. In other words, the modulation needs to provide the best error performance for a given energy per bit. Second, the choice of a modulation scheme affects the structure of the PSD and thus has the potential to impose additional constraints on the total transmitted power [24, 52, 111].

On the other hand, the PSD of the TH-UWB signals plays a major role in key aspects like good understanding of the effect of TH-UWB signals on the NB receiver. If the PSD is understood, it may be possible to reduce or eliminate interference of TH-UWB signal from the selected NB system. Details of TH-UWB interference on NB system is discussed

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in section 2.8 and chapter 5. In this section both the power efficiency and the effect of modulation schemes on the PSD are discussed for PPM scheme [112, 113].

2.7.1 Analysis of PSD

A mathematical frame work is essential to enable the evaluation of the PSD of a modulated impulse radio signal using a deterministic TH code. The TH-UWB signal presented in Eq. (2.13) is required to be modified, to analyze PSD of M -ary symbols. Assuming the analysis is for user 1, superscript term in Eq. (2.13) is omitted for simplicity, so it can be defined as

$$s(t) = \sum_{l=0}^{M-1} \sum_{h=0}^{N_s-1} a_l w(t - lN_p T_f + hT_f - c_{l,h} T_c - \delta_l) \quad (2.39)$$

where a_l and δ_l are the independent and stationary process, $c_{l,h}$ is the TH-sequence and N_p is the period of TH sequence. To simplify the analysis of PSD of TH-UWB signal, it is assumed that N_s is equal to N_p . Since the Eq. (2.39) depends on the time dithering, it can be expressed as continuous form i.e.

$$y(t) = \sum_l s_{p_l}(t - lN_p T_f). \quad (2.40)$$

The PSD is computed by evaluating the Fourier transformation (FT) of the autocorrelation function of $y(t)$ i.e.

$$P_y(f) = \mathcal{F} \left\{ E \{ y(t)y(t+\tau) \} \right\} \quad (2.41)$$

where $\mathcal{F}\{.\}$ denotes the FT and $E\{.\}$ denotes the expected value. Therefore, the PSD can be expressed as [62]

$$\begin{aligned} P_y(f) = & \frac{1}{N_p T_f} \left[E \{ |S_p(f)|^2 \} - E \{ S_p(f) S_q^*(f) \} \right] \\ & + \frac{1}{(N_p T_f)^2} \sum_k E \{ S_p(f) S_q^*(f) \} \delta \left(f - \frac{k}{N_p T_f} \right) \end{aligned} \quad (2.42)$$

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where p and q are two independent random variables with the same probability distribution function as p_l . $S_p(f)$ is the FT of $s_p(t)$ which can be evaluated as

$$S_p(f) = W(f) \sum_{l=0}^{M-1} a_l e^{-j2\pi f \delta_l} T_l(f) \quad (2.43)$$

where $W(f)$ is the FT of the transmitted pulse $w(t)$ and $T_l(f)$ is the FT of the TH code which is transmitting the l^{th} symbol

$$T_l(f) = \sum_{h=0}^{N_s-1} e^{-j2\pi f (c_l h T_c + (lN_p + h)T_f)}. \quad (2.44)$$

To find a closed form expression of $P_y(f)$, the expectation of $|S_p(f)|^2$ given in Eq. (2.42) need to be evaluated and can be written as

$$E\{|S_p(f)|^2\} = E\left\{|W(f)|^2 \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} a_l a_n e^{-j2\pi f (\delta_l - \delta_n)} T_l(f) T_n^*(f)\right\}. \quad (2.45)$$

Since a_l and a_n are independent random variables from the same process and δ_l and δ_n are independent random variables from the different process, the Eq. (2.45) can be evaluated as

$$E\{|S_p(f)|^2\} = |W(f)|^2 \sum_{l=0}^{M-1} \left\{ E\{a_l a_l\} |T_l(f)|^2 + \sum_{\substack{n=0 \\ n \neq l}}^{M-1} E\{a_l a_n\} E\{e^{-j2\pi f (\delta_l - \delta_n)}\} T_l(f) T_n^*(f) \right\}. \quad (2.46)$$

Similarly the second expectation in Eq. (2.42) can be expressed as

$$E\{S_p(f) S_q^*(f)\} = |W(f)|^2 \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} E\{a_l a_n\} E\{e^{-j2\pi f (\delta_l - \delta_n)}\} T_l(f) T_n^*(f). \quad (2.47)$$

The waveforms $s_p(t)$ and $s_q(t)$ are generated by two i.i.d process. Therefore, expectations in Eq.(2.47) are independent of l and n and equal to the case $l \neq n$ of Eq.(2.46) i.e.

$$E\{S_p(f) S_q^*(f)\} = |W(f)|^2 E\{a_l a_n\} E\{e^{-j2\pi f (\delta_l - \delta_n)}\} \sum_{l=0}^{M-1} \sum_{\substack{n=0 \\ n \neq l}}^{M-1} T_l(f) T_n^*(f). \quad (2.48)$$

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To simplify the above equation new variables are introduced for expected values which can be defined as

$$\begin{aligned}
 E\{a_l a_n\} &= R_0^a & \text{if } l = n \\
 E\{a_l a_n\} &= R_1^a & \text{if } l \neq n \\
 E\{e^{-j2\pi f(\delta_l - \delta_n)}\} &= R_0^\delta & \text{if } l = n \\
 E\{e^{-j2\pi f(\delta_l - \delta_n)}\} &= R_1^\delta & \text{if } l \neq n.
 \end{aligned} \tag{2.49}$$

Using above newly defined variables in Eq.(2.46) and Eq.(2.48) and substituting these in Eq.(2.42) the final PSD can be formulated as

$$\begin{aligned}
 P_y(f) = \frac{|W(f)|^2}{N_p T_f} \left\{ R_0^a - R_1^a R_1^\delta \right\} \sum_{l=0}^{M-1} |T_l(f)|^2 + \\
 \frac{|W(f)|^2}{(N_p T_f)^2} R_1^a R_1^\delta \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} T_l(f) T_n^*(f) \sum_k \delta\left(f - \frac{k}{N_p T_f}\right).
 \end{aligned} \tag{2.50}$$

It is observed that the PSD of a TH-UWB signal in Eq. (2.50) consists of continuous and discrete PSD components. How the total power distributed over both the components depends partly on the modulation parameters i.e. expected values of Eq. (2.49). The first part of Eq. (2.50) is the continuous component which does not create serious problems on interference with other NB systems. However, the second part is a discrete part that mainly causes interference with other NB systems. The discrete PSD components can be reduced by controlling modulating parameters. The effect of PSD for two modulation schemes such as PPM and PAM have been illustrated as follows.

For binary PPM modulation scheme the amplitude remains fixed, however, the position is changed with the information, therefore Eq.(2.49) can be rewritten for binary PPM TH-UWB signal as

$$R_0^a = R_1^a = A^2, R_1^\delta(f) = \frac{(1 + \cos(2\pi f \delta))}{2} \tag{2.51}$$

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where δ is the time shift from the nominal position. Hence the Eq.(2.50) can be reformulated for PPM modulation scheme as

$$P_y(f) = \frac{|W(f)|^2 A^2}{N_p T_f} \left\{ \frac{1 - \cos(2\pi f \delta)}{2} \right\} \sum_{l=0}^{M-1} |T_l(f)|^2 + \frac{|W(f)|^2 A^2}{(N_p T_f)^2} \times \left\{ \frac{1 + \cos(2\pi f \delta)}{2} \right\} \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} T_l(f) T_n^*(f) \sum_k \delta \left(f - \frac{k}{N_p T_f} \right) \quad (2.52)$$

where, in general, both components remain in the PSD for PPM modulation scheme. Therefore, it has the most constant power spectrum over the whole frequency band. The power spectrum does not drop for the TH-PPM, so it offers high power over a long frequency range.

However for PAM modulation, the position remains the same and only the amplitude is changed with data, therefore Eq.(2.49) can be expressed as

$$R_0^a = R_1^a = 0, R_1^\delta(f) = \frac{(1 + \cos(2\pi f \delta))}{2} \quad (2.53)$$

and the corresponding PSD for TH-PAM modulation scheme can be rewritten as

$$P_y(f) = \frac{|W(f)|^2 A^2}{N_p T_f} \sum_{l=0}^{M-1} |T_l(f)|^2. \quad (2.54)$$

It shows that PSD contains only continuous frequency components for binary PAM modulation scheme. Therefore, binary PAM has the lowest possibility of detection and the lowest possibility of interfering with existing NB systems. Since PAM has amplitude variations, the time frame is longer in duration to accommodate longer pulse train. This will cause fewer spectral combs. However, due to almost no power spikes, the detection of the signal is very difficult, so it would be tactically best for covert users.

2.8 UWB Interference on NB Receiver

In the above PSD analysis, it is observed that modulation schemes play an important role in designing PSD of TH-UWB signal. On the other hand UWB interference on NB

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receiver depends on PSD of TH-UWB signals. Consequently, it is necessary to implement TH-UWB modulation schemes and corresponding pulse waveforms that effectively and adaptively suppress UWB interference on NB systems [40, 41, 114]. Therefore it is not only sufficient to develop a modulation for high data rate and low complexity system, but also required to see the PSD of the modulation and its mitigation potentiality with other NB systems. For this reason, in this dissertation the UWB interference on NB system is studied in detail to show the coexistence capability of the modulation schemes.

TH-UWB systems spread transmitted signal power over an extremely large frequency band with very low power spectral density. This large frequency band is allocated to other radio systems. The result is that UWB system degrades the performance of other systems which overlay in the same frequency band, or operate in nearby frequency band including IEEE802.11a and IEEE802.11b wireless local area networks (WLANs), global positioning system (GPS) receiver, CDMA-based cellular systems, Bluetooth, broadcasting, etc. Among these, IEEE802.11a WLAN system is the mostly affected NB systems whose performance degradation increases significantly in the presence of TH-UWB systems [115, 116]. However, coexistence of TH-UWB with WLAN system is more important to commercial consumer sectors. Therefore, this dissertation analyzes the effect of UWB interference on coexisting NB systems. We have considered IEEE802.11a WLAN as a typical example of coexisting NB systems. The system description of IEEE802.11a and its interference with TH-UWB systems are given in the following section [32, 117]. The mitigation technique and in-band interference power are described in chapter 5.

2.8.1 IEEE802.11a Systems Model

The interference issues of TH-UWB systems over a NB radio system have been studied extensively in [115, 116]. The operating band of WLAN based on IEEE802.11a standards overlaps UWB band and its performance is effected by the interference generated by

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UWB systems. The WLAN uplink receiver is generally far from the UWB systems and WLAN downlink receiver is mainly affected by the UWB interference power. The relative geometry of coexisting 802.11a and UWB systems are shown in Fig. 2.13. It is assumed that IEEE802.11a node is placed in center and surrounded by several UWB transmitters in a two-dimensional setting.

IEEE 802.11a can support data-rates up to 54 Mbps for a distance up to 30 meters, and, therefore, use of these systems has seen a significant increase in recent years [1]. These systems operate in the 5 GHz unlicensed national information infrastructure (UNII) frequency bands using OFDM based transmission. In this spectrum three 100 MHz wide frequency bands (5.15 – 5.25, 5.25 – 5.35 and 5.725 – 5.825GHz) exist, with the maximum effective radiated power (ERP) per MHz set to 2.5 mW, 12.5 mW and 50 mW respectively. Each frequency band is divided into five 20 MHz wide channels, in which 802.11a employs 52 sub-carriers (48 for data transmission and 4 dedicated for channel estimation). The IEEE802.11a standard requires receivers to have a minimum sensitivity ranging from –82 dBm to –65 dBm, depending on the chosen and/or required data rate [117, 118].

The primary interest here is the effect of a TH-UWB signal on a NB receiver. Since TH-WB energy spreads over large frequency band and NB energy is in a small portion of frequency band, only a small portion of TH-UWB signal is collected by NB receiver. Fig. 2.14 shows the TH-UWB pulse ESD and the NB filter frequency response where $W(f)$ is the FT of UWB pulse, $H_{IF}(f)$ is the frequency response of a band-pass filter with center frequency f_0 , corresponds to the center frequency of the NB systems.

2.8.2 Estimation of UWB Interference Power

The TH-UWB interference on IEEE802.11a system can be realized by evaluating the in-band interference of TH-UWB in the IEEE802.11a receiver. The in-band UWB interference power on IEEE802.11a receiver in the presence of single TH-UWB system can be

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Figure 2.13: Relative geometry of coexisting 802.11a and UWB systems

Figure 2.14: Illustration of UWB pulse ESD and narrowband filter frequency response

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expressed as [116]

$$\begin{aligned}
 P_{INB} &= \int_{-\infty}^{\infty} P_y(f) |H(f)|^2 df \\
 &= \int_{-\infty}^{\infty} \frac{|W(f)H(f)|^2 A^2}{N_p T_f} \left(\frac{1 - \cos(2\pi f \delta)}{2} \right) \sum_{l=0}^{M-1} |T_l(f)|^2 df + \\
 &\quad \sum_{\frac{k}{N_p T_f} \pm f_0 \in [-B/2, B/2]} \left\{ \frac{|W(\frac{k}{N_p T_f})H(\frac{k}{N_p T_f})|^2 A^2}{(N_p T_f)^2} \left(\frac{1 + \cos(\frac{2\pi k \delta}{N_p T_f})}{2} \right) \right. \\
 &\quad \left. \times \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} T_l(\frac{k}{N_p T_f}) T_n^*(\frac{k}{N_p T_f}) \right\}
 \end{aligned} \tag{2.55}$$

where B is the bandwidth of the bandpass filter at the front end of the NB receiver and $P_y(f)$ is the PSD of UWB signal defined in Eq.(2.50). $H(f)$ is the FT of the filter $h(t)$. Assuming $h(t)$ is given as an ideal bandpass filter, such that:

$$H(f) = \begin{cases} 1 & f_0 - B/2 \leq f \leq f_0 + B/2; \\ 1 & -f_0 - B/2 \leq f \leq -f_0 + B/2; \\ 0 & \text{elsewhere.} \end{cases} \tag{2.56}$$

Therefore, Eq.(2.55) finally can be simplified as:

$$\begin{aligned}
 P_{INB} &= \int_{f_0 - B/2}^{f_0 + B/2} \frac{|W(f)|^2 A^2}{N_p T_f} \left(\frac{1 - \cos(2\pi f \delta)}{2} \right) \sum_{l=0}^{M-1} |T_l(f)|^2 df + \\
 &\quad \sum_{\frac{k}{N_p T_f} - f_0 \in [-B/2, B/2]} \left\{ \frac{|W(\frac{k}{N_p T_f})|^2 A^2}{(N_p T_f)^2} \left(\frac{1 + \cos(\frac{2\pi k \delta}{N_p T_f})}{2} \right) \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} T_l(\frac{k}{N_p T_f}) T_n^*(\frac{k}{N_p T_f}) \right\}.
 \end{aligned} \tag{2.57}$$

The above equation shows that the in-band interference power on NB receiver fully depends on the ESD of UWB pulse waveform, TH code, pulse repetition interval, symbol duration, number of bits in a symbol and bandwidth of the NB receiver. It is to be noted that the above expression gives the total amount of interference power without assuming path losses or shadowing. However, UWB interference suppression method in the presence of pathloss is given in below and in more details at chapter 5.

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In order to calculate the Signal-to-Interference Ratio (SIR) of the 802.11a system in presence of the UWB interference, the path loss model and the UWB interference power are to be determined. The UWB path loss model for LOS and NLOS in a residential environment can be expressed as [115]

$$PL(d) = PL_0 + 10\gamma \log_{10}(d) + S. \quad (2.58)$$

Here, constants PL_0 and γ were determined empirically to be 47 dB and 1.7 respectively for LOS paths; and 51 dB and 3.5 for NLOS paths. S is the well-known shadow fading factor which represents a zero mean Gaussian random variable. The WLAN channel model is assumed to be flat-fading. A Rician channel with a K-factor of 6 dB was assumed for LOS paths and a Rayleigh channel was assumed for NLOS [115]. In both cases, the dB-received power from each transmitter is calculated by using Eqn. (2.58) as

$$\begin{aligned} P_{R,UWB}(d_{UWB}) &= P_{T,UWB} - PL(d_{UWB}) \\ P_{R,802.11a}(d_{802.11a}) &= P_{T,802.11a} - PL(d_{802.11a}) \end{aligned} \quad (2.59)$$

where $P_{T,UWB}$ is the transmit power of UWB transmitter, $P_{T,802.11a}$ is the transmit power of 802.11a transmitter, d_{UWB} is the UWB transmitter and receiver distance and the corresponding received power of UWB receiver is $P_{R,UWB}(d_{UWB})$. Similarly $d_{802.11a}$ is the 802.11a transmitter and receiver distance and corresponding received power of IEEE802.11a is $P_{R,802.11a}(d_{802.11a})$. The UWB interference power on the 802.11a system can be written as

$$\tilde{I}_{UWB}(f_{NL}, f_{NH}) = P_{T,UWB} - PL(d_{UWB}) \quad (2.60)$$

where f_{NL} and f_{NH} are the lower and upper band-edges of the 802.11a receiver respectively. The SIR of the 802.11a receiver is then

$$SIR_{802.11a} = P_{R,802.11a} - \tilde{I}_{UWB}(f_{NL}, f_{NH}). \quad (2.61)$$

To get better performance of 802.11a system, it is necessary to reduce the $\tilde{I}_{UWB}(f_{NL}, f_{NH})$ which is shown in chapter 5.

2.9 Summary

In this chapter, non orthogonal and orthogonal pulses and their advantages and disadvantages for TH-UWB systems have been described. The corresponding modulation schemes such as PPM, PAM, OOK, BPSK, PSM and their combined forms have been described in detail. The TH-UWB transmitter for single user and multi users transmission and corresponding mathematical formulations are presented. The channel models for signal propagation have been described. For TH-UWB system, IEEE802.15.3a channel model and its fading behavior have been presented by providing mathematical distribution functions. The optimum correlation receiver and RAKE receiver have been presented. The corresponding template signals have been provided to collect the received signal for the different channel models. The channel capacity for M -ary PPM scheme has been investigated to show the efficacy of TH-UWB modulation scheme in terms of capacity. The PSD efficiency of various modulation schemes have been described to show their power efficiency and mitigation capability with NB systems. Finally, TH-UWB interference issues on the NB system have been discussed based on these pulse waveforms, modulation schemes and PSD.

Chapter 3

M-ary Signaling based on OOK and PSM Modulation Schemes

3.1 Introduction

Several modulation schemes have been proposed in literature [59–63] for TH-UWB systems for different applications. Among these, orthogonal pulse based modulation schemes are popular because of their better system performance. These modulation schemes reduce MAI and ISI considerably by using correlation properties of orthogonal pulses. However, designing such modulation schemes in the presence of multiple users and multipath channel environments is a challenging task for TH-UWB systems. Several design issues related to system performance, interference reduction, capacity enhancement, reduction of system complexity, etc. still remain unsolved.

This chapter describes a combined modulation scheme for TH-UWB systems by using OOK and PSM schemes. It uses a set of orthogonal pulses to represent bits in a symbol. These orthogonal pulses are transmitted simultaneously in the same pulse repetition interval resulting in a composite pulse. MAI, IPI and MPI are analyzed in the presence of multiple users and multipath environments. The theoretical formulation for RAKE receiver is presented to collect the multipath signal. BER performance of the proposed scheme is analyzed through mathematical analysis and simulation under different

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channel conditions. The capacity of TH-UWB systems is presented in detail to analyze the efficiency of OOK-PSM scheme. The performance of this scheme is compared with existing PSM and its combined modulation schemes.

This chapter is organized as follows: The details of TH-UWB system model are described in section 3.2. Section 3.3 describes the combined modulation scheme based on combination of OOK and PSM modulation. Section 3.4 presents the transceiver structure and system performance under the AWGN channel model. Section 3.5 provides the RAKE receiver structure, analyzing of MAI, IPI, MPI and system performance in the presence of multiple users in a multipath environment. The results of simulation studies under different channel conditions are discussed in section 3.6. Finally, a summary of this chapter is provided in section 3.7.

3.2 PSM and TH-UWB Radio Systems

The successful deployment of UWB radio systems for high-speed indoor communication strongly depends on the development of pulses, modulation techniques and low complexity receivers. Message symbols are transmitted using short analog waveforms confined to within the frequency range of UWB radio. Various kinds of modulation schemes such as M -ary PPM, M -ary PAM, OOK, BPSK have been proposed for TH-UWB systems by using short pulse waveform to achieve high data rate transmission [18, 119]. The major challenge is the selection of appropriate modulation scheme to design a reliable communication system [60, 120].

PSM scheme is presently an interesting research topic for TH-UWB systems to transmit signal in efficient way. It is considered to be a suitable choice for TH-UWB systems for its robustness against MAI and ISI [47]. PSM scheme requires a large set of orthogonal pulses for higher level modulation schemes. However, due to speculative autocorrelation

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property of higher order orthogonal pulses, PSM cannot be used for higher level modulation schemes. Moreover, it increases the system complexity for higher level modulation schemes by increasing the number of correlators in the receiver [36, 49].

To address these problems, several combined forms of PSM scheme such as M -ary BPSM, M -ary 2PPM-PSM and M -ary BPSK-PSM have been proposed for M -ary signaling. These can transmit the same amount of data by using fewer orthogonal pulses and uses fewer correlators than those used in PSM schemes [28, 30]. The other objective for combined modulation schemes is that they are able to reduce the number of spectral spikes in TH-UWB signal, which helps TH-UWB system to coexist with NB systems without significant interference [29, 114, 116]. However, their performance, system capacity and several interference issues have not been analyzed so far, to get the complete knowledge of orthogonal pulse based TH-UWB systems.

In this chapter, a new combined modulation scheme based on OOK and PSM for M -ary modulation schemes is proposed to reduce system complexity and is referred to as OOK-PSM scheme. Here information bits or symbols are mapped onto orthogonal pulses by on-off-keying [121, 122]. Since orthogonal pulses are more susceptible in multipath channel, the BER performance of the proposed scheme is analyzed through mathematical analysis and simulations for different channel conditions. The existing PSM and its combined modulation schemes are also analyzed in multipath channel model and compared with the proposed scheme.

3.3 System Model of OOK-PSM Scheme

The proposed method maps a set of message bits or symbol onto one or more orthogonal pulses by on-off keying. The number of pulses in each symbol depends on the number of non-zero bits. Table 3.1 shows the examples of 2-bit and 3-bit symbol transmission and the corresponding transmitted pulses. Fig. 3.1 shows composite pulses for the symbols of

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a 3-bit modulation scheme. In general, N -bit OOK-PSM scheme requires N orthogonal pulses to transmit an N -bit symbol, where $M = 2^N$. These N independent bits are transmitted during same time frame by assigning different orthogonal pulses resulting in a composite pulse. The presence of composite pulse is decided by on-off-keying, i.e. pulse is present for *one* and absent for *zero*. Pulses overlay in both time and frequency domains without any interference [72]. The composite pulse goes through a set of correlators in the receiver. The correlation receiver is designed using a set of template signals which are similar to the set of orthogonal pulses used at the transmitter. Each correlator recovers a pulse from composite pulses by exploiting their correlation properties. The block diagram of an N -bit symbol transmitter and correlation receiver is shown in Fig. 3.2.

The proposed method has several advantages over conventional methods. It uses fewer pulses and receiver correlators than those used in PSM and BPSM schemes, which leads to lower complexity for the system design [28, 36]. Since *zero* is represented by absence of pulse, the proposed scheme uses low average transmitted power, which is crucial for energy constrained UWB systems. Furthermore due to the absence of pulse for *zero*, OOK complexity is nearly half of that of other conventional modulation schemes. The validity of complexity reduction and power consumption is true for all combined schemes of OOK. Comparing with an equivalent BPSK-PSM scheme, an N -bit OOK-PSM scheme has at most $(3/4)^{th}$ complexity of that of other orthogonal pulse based combined schemes irrespective of the level of modulation schemes. It is known that binary OOK scheme has synchronization problem when long sequence of zeros are transmitted. However, proposed scheme can be used for higher level modulation schemes, therefore it reduces the synchronization problem by reducing the length of long sequence of zeroes.

Since the proposed scheme uses orthogonal pulses, MAI can be reduced considerably by assigning different subsets of orthogonal pulses for different users. As the proposed scheme transmits more bits by using fewer orthogonal pulses, it generates fewer spectral

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Table 3.1: Transmitted pulses for 2-bit and 3-bit symbols

Schemes		T_f			Combined form of transmitted pulses
		$w_0(t)$	$w_1(t)$	$w_2(t)$	
2-bit	00	Off	Off	Off	None
	01	Off	Off	On	$w_2(t)$
	10	Off	On	Off	$w_1(t)$
	11	Off	On	On	$w_1(t) + w_2(t)$
3-bit	000	Off	Off	Off	None
	001	Off	Off	On	$w_2(t)$
	010	Off	On	Off	$w_1(t)$
	011	Off	On	On	$w_1(t) + w_2(t)$
	100	On	Off	Off	$w_0(t)$
	101	On	Off	On	$w_0(t) + w_2(t)$
	110	On	On	Off	$w_0(t) + w_1(t)$
	111	On	On	On	$w_1(t) + w_2(t) + w_0(t)$

Figure 3.1: Composite pulse waveforms and symbols for an 3-bit OOK-PSM modulation scheme

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Figure 3.2: Transceiver structure for N -bit OOK-PSM modulation scheme in AWGN channel

spikes in the signal. Therefore, the proposed scheme can coexist with overlapping NB systems without causing significant interference [29,116]. The overall scheme is downward compatible. That is, the lower-level scheme can be used in higher-level system without changing the hardware design. The 3-bit scheme can be changed into 2-bit scheme by just keeping off $w_0(t)$ and can be changed into binary scheme by keeping off $w_0(t)$ and $w_1(t)$. This property can be exploited further for an adaptive modulation system based on channel conditions at any given instant.

The designing of transmitted signal in the transmitter depends on the modulation scheme and codes for multiple users to avoid catastrophic collision among users. The OOK-PSM modulation signal of the k^{th} user and for the i^{th} symbol can be defined as

$$s_i^{(k)}(t) = \sqrt{E_{tx}^{(k)}} \sum_{j=0}^{N_s-1} \mathbf{a}_i \mathbf{w}^{(k)}(t - jT_f - c_j^{(k)}T_c) \quad (3.1)$$

where $i = 0, 1, \dots, M-1$, T_f is the pulse repetition interval, index j represents the number of pulse repetition interval for a symbol, $c_j^{(k)}$ is the TH sequence with chip duration T_c and

$$\mathbf{w}^{(k)}(t) = [w_0^{(k)}(t) \ w_1^{(k)}(t) \ \dots \ w_{N-1}^{(k)}(t)]^T \quad (3.2)$$

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is the N -dimensional column vector of k^{th} user, $w_n^{(k)}(t)$ is the n^{th} order orthogonal pulse of k^{th} user, $\mathbf{a}_i$ is the N -bit binary row data vector for the i^{th} symbol.

3.4 Performance of OOK-PSM in AWGN Channel

The receiver structure and its performance depend on modulation schemes and channel models. In this section, the correlation receiver structure and system performance are analyzed. The correlation receiver contains N correlators for N -bit OOK-PSM scheme. Since the system supports N_u users, the received signal in AWGN channel can be written as

$$r(t) = \sum_{k=1}^{N_u} \sqrt{E_{rx}^{(k)}} s^{(k)}(t - \tau^{(k)}) + n(t) \quad (3.3)$$

where $\tau^{(k)}$ is the time delay for k^{th} user, and $n(t)$ is the AWGN. The received signal passes through N correlators. In each correlator, the received signal is multiplied by a template signal and the corresponding transmitted bit is decided by exploiting the correlation properties of the pulse waveform. Hard decision decoding is assumed at the correlator, followed by a parallel to serial converter as shown in Fig. 3.2. The receiver performance can be improved by changing hard decision decoding into high performance soft decision decoding methods.

The number of correlators in the receiver is the same as the number of bits in a symbol. For simplicity, the index 1 is often omitted in the single user case. If N_s is the number of repetition of pulse waveforms for each symbol, the reference bit b is defined in the time interval $[0, T_b]$, where $T_b = N_s T_f$. The decision statistics of the user 1 is

$$\begin{aligned} y &= \int_0^{T_b} r(t) \mathbf{w}^{(1)}(t - jT_f - c_j^{(1)}T_c) dt \\ &= \sum_{k=1}^{N_u} \int_0^{T_b} \left(\sqrt{E_{rx}^{(k)}} s^{(k)}(t - \tau^{(k)}) + n(t) \right) \mathbf{w}^{(1)}(t - jT_f - c_j^{(1)}T_c) dt \\ &= [Z_0 \quad Z_1 \quad \cdots \quad Z_{N-1}]^T \end{aligned} \quad (3.4)$$

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where $\mathbf{w}^{(1)}(t)$ is assumed as template signals as defined in Eq. (3.2) by neglecting transceiver derivative characteristics, Z_l is the test statistic of l^{th} correlator which undergoes a hard decision decoding where $l = 0, 1, \dots, N - 1$. The value of Z_l can be expressed as

$$Z_l = Z_{l,s} + Z_{l,MAI} + Z_{l,n} \quad (3.5)$$

where $Z_{l,s}$ is the desired signal, $Z_{l,MAI}$ is the MAI term and $Z_{l,n}$ is the AWGN term at the l^{th} correlator. It is assumed that perfect synchronization exists between the desired transmitter and reference receiver and that the time delay is known at the reference receiver, that is $\tau^{(1)} = 0$. The desired signal $Z_{l,s}$ can be expressed as

$$Z_{l,s} = \sum_{j=0}^{N_s-1} \int_{jT_f+c_j^{(1)}T_c}^{jT_f+c_j^{(1)}T_c+T_c} \sqrt{E_{rx}^{(1)}} s^{(1)}(t) w_l(t - jT_f - c_j^{(1)}T_c) dt \quad (3.6)$$

where $w_l(t)$ is the template signal of l^{th} correlator. The transmitted pulse of the desired user occurs within the chip duration T_c , so, the time frame $[jT_f, (j+1)T_f]$ changes into $[jT_f + c_j^{(1)}, jT_f + c_j^{(1)}T_c + T_c]$. Assuming l^{th} order pulse is present in the composite pulse, the signal energy of user 1 at the l^{th} correlator for N_s time frame can be expressed as

$$E_b^{(1)} = (Z_{l,s})^2 = E_{rx}^{(1)} N_s^2 \int_0^{T_c} h_l^2(t) dt = E_{rx}^{(1)} N_s^2 \quad (3.7)$$

where $E_{rx}^{(k)}$ is the received amplitude of the k^{th} user.

Under the standard Gaussian approximation, $Z_{l,n}$ and $Z_{l,MAI}$ are assumed to be as zero-mean Gaussian random processes as characterized by variances σ_n^2 and σ_{mai}^2 respectively. Due to timing jitter error from $N_u - 1$ interfering users, the timing jitter error ϵ is uniformly distributed over $[\Delta, -\Delta]$, where $\Delta = 0.1$ for MHPs up to 4^{th} order [36, 52, 123]. $Z_{l,MAI}$ is the total MAI at l^{th} correlator and can be expressed as [120]

$$Z_{l,MAI} = \sum_{k=2}^{N_u} \sum_{j=0}^{N_s-1} \int_{jT_f}^{(j+1)T_f} s^{(k)}(t - \tau^{(k)} - \epsilon) w_l(t - jT_f - c_j^{(1)}T_c) dt. \quad (3.8)$$

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As the timing jitter error from interfering user is very small when compared to $\tau^{(k)} \in [0, N_s T_f]$, one can assume that $\tau^{(k)} + \epsilon \approx \tau$ is uniformly distributed over the interval $[0, N_s T_f]$. Therefore, the total interference energy from other users can be evaluated as

$$\sigma_{l,mai}^2 = \frac{N_s}{T_f} \sum_{k=2}^{N_u} \int_0^{T_f} \left(\sqrt{E_{rx}^{(k)}} \int_0^{T_p} w_n^{(k)}(t - \tau) w_l(t) dt \right)^2 d\tau \quad (3.9)$$

where T_p is the width of the pulses, $w_n^{(k)}$ is the n^{th} order pulse from the k^{th} user. If all users use the same set of orthogonal pulses, n takes any one value from the set $\{0, 1, \dots, N-1\}$. On the other hand, if all users use different exclusive subsets of the same set of orthogonal pulses, n is not equal to l . The order of pulse waveform of user 1 is l which is used at the l^{th} correlator. It can be assumed that $E_{rx}^{(1)} = E_{rx}^{(2)} = \dots = E_{rx}^{(N_u)} = E_{rx}$ for perfect power control for all users. Since correlation value depends on the width of the pulses, Eq. (3.9) can be expressed as

$$\begin{aligned} \sigma_{l,mai}^2 &= \frac{N_s}{T_f} E_{rx} \sum_{k=2}^{N_u} \int_0^{T_p} \left(\int_0^{T_p} h_n^{(k)}(t - \tau) h_l(t) dt \right)^2 d\tau \\ &= \frac{N_s}{T_f} E_{rx} \sum_{k=2}^{N_u} \int_0^{T_p} (R_{n,l}^{(k)}(\tau))^2 d\tau \end{aligned} \quad (3.10)$$

where $R_{n,l}^{(k)}(\tau)$ is the correlation between n^{th} and l^{th} order pulses. The term $R_{n,l}^{(k)}(\tau)$ becomes $R_{l,l}^{(k)}(\tau)$ or $R_l^{(k)}(\tau)$ if k^{th} user uses l^{th} order pulses in the given time $[0, N_s T_f]$. Due to correlation properties of orthogonal pulses, the term $R_{n,l}^{(k)}$ is always less than $R_{l,l}^{(k)}(\tau)$ for synchronized systems. In conventional systems, the above correlation term is always between same pulses and is referred to as autocorrelation value. The sum of these correlation values gives significant amount of MAI. MAI is lower for orthogonal pulse based modulation schemes, since it is the sum of autocorrelation and crosscorrelation values. The MAI can be further reduced by sharing mutually exclusive subsets of the same set of orthogonal pulses among the different users. In such cases MAI contains only crosscorrelation values.

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$Z_{l,n}$ is the AWGN at the l^{th} correlator output

$$Z_{l,n} = \sum_{j=0}^{N_s-1} \int_{jT_f}^{(j+1)T_f} n(t) w_l(t - jT_f - c_j^{(1)} T_c) dt \quad (3.11)$$

and the corresponding variance i.e, noise power of AWGN can be expressed as follows [28]

$$\sigma_n^2 = \frac{N_s N_0}{2} \int_0^{T_f} w_l^2(x) dx = \frac{N_s N_0}{2}. \quad (3.12)$$

Due to different autocorrelation values for different pulses, each correlator gives different probability of error. The corresponding probability of error of the l^{th} correlator in the presence of MAI is [123]

$$P_l = Q \left(\sqrt{\frac{E_b^{(1)}}{2(\sigma_n^2 + \sigma_{l,mai}^2)}} \right) \quad (3.13)$$

$$= Q \left(\sqrt{\frac{N_s^2 E_{rx}}{2 \left(\frac{N_0}{2} N_s + \frac{N_s}{T_f} E_{rx} \sum_{k=2}^{N_u} \int_0^{T_p} (R_{n,l}^{(k)}(\tau))^2 d\tau \right)}} \right) \quad (3.14)$$

$$= Q \left(\sqrt{\frac{E_b}{\left(N_0 + 2R_b E_b \sum_{k=2}^{N_u} \int_0^{T_p} (R_{n,l}^{(k)}(\tau))^2 d\tau \right)}} \right)$$

where $R_b = 1/N_s T_f$ is the data rate and $E_b = N_s E_{rx}$ is the received energy at the receiver.

Since each decision is independent, the average probability of bit error can be defined as

$$Pr_b = \frac{1}{N} \sum_{l=0}^{N-1} P_l \quad (3.15)$$

where N is the total number of correlators for N -bit symbols transmission. The correct decision of the l^{th} correlator is $1 - P_l$. The received symbol is perfect if all correlators make correct decisions. Since decisions are independent, the correct decision for a symbol can be defined as

$$P_c = \prod_{l=0}^{N-1} (1 - P_l) \quad (3.16)$$

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The probability of symbol error rate can be calculated by using Eq. (3.13). The N -bit symbols error rate can be expressed as

$$P_r = \left(1 - \prod_0^{N-1} \left(1 - Q \left(\sqrt{\frac{E_b}{(N_0 + 2R_b E_b \sum_{k=2}^{N_u} \int_0^{T_p} (R_{n,l}^{(k)}(\tau))^2 d\tau}} \right) \right) \right) \quad (3.17)$$

3.5 Performance in Multipath Channel

The RAKE receiver structure is more complicated when it is designed for multipath channel. Fig. 3.3 shows the RAKE receiver structure for multipath channel. The performance and robustness of a system in multipath environment are often determined by the amount of multipath energy that can be collected at the receiver. If there are N_u users and each experiences a different channel, then the received signal can be expressed as

$$r(t) = \sum_{k=1}^{N_u} \sum_{l=1}^{L_p} \alpha_l^{(k)} s^{(k)}(t - \tau_l^{(k)}) + n(t) \quad (3.18)$$

where $\alpha_l^{(k)}$ is the path gain and $\tau_l^{(k)}$ is the time delay of l^{th} path for k^{th} user, and $n(t)$ is the AWGN. The reference signal of user 1 at q^{th} ($= 0, 1, \dots, N-1$) correlator can be expressed as

$$\phi_q^{(1)}(t) = \sum_{j=0}^{N_s-1} v_q^{(1)}(t - jT_f - c_j^{(1)}T_c) \quad (3.19)$$

where N_s is the total number of time frames for a symbol and

$$v_q^{(1)}(t) = \sum_{p=1}^{L_p} \alpha_p^{(1)} w_q^{(1)}(t - \tau_p^{(1)}). \quad (3.20)$$

Since multiple pulses are transmitted in one time frame, the transmitted and received signal contain several pulses. However, template signal at each RAKE finger contains only one pulse. Therefore, when sum of several pulses is correlated with a single pulse waveform, there are several interferences in the presence of timing jitter and asynchronous systems. Pulses with short duration are not orthogonal and they may overlap one another.

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When a pulse overlaps with the same pulse, it is called inter pulse interference (IPI) or self interference and when pulse interferes with other pulses, it is called multi pulse interference (MPI). The decision statistics of the user 1 in the q^{th} correlator can be written as

$$\begin{aligned} Z_q^{(1)} &= \int_{jT_f}^{(j+1)T_f} r(t)\phi_q^{(1)}(t)dt \\ &= S_q^{(1)} + IPI_q^{(1)} + MPI_q^{(1)} + MAI_q^{(1)} + N_q^{(1)} \end{aligned} \quad (3.21)$$

where $S_q^{(1)}$ is the desired signal, $IPI_q^{(1)}$ is the IPI, $MPI_q^{(1)}$ is the MPI, $MAI_q^{(1)}$ is the MAI due to the presence of multiple users and $N_q^{(1)}$ is the AWGN term. The IPI, MPI, MAI and AWGN terms behave like interference noise mixed with the original signal. The correct decision of $Z_q^{(1)}$ can be taken only if the desired signal, IPI, MPI, MAI and AWGN are known precisely. Therefore, these terms need to be analyzed.

3.5.1 Desired Signal

For analysis, it is assumed that perfect synchronization exists between the transmitter and reference receiver. Assuming that $\tau_l^{(1)} = 0$ and that the transmitted symbol uses q^{th} order pulse $w_q^{(1)}(t)$, the desired signal $S_q^{(1)}$ can be expressed as [109, 124]

$$\begin{aligned} S_q^{(1)} &= \sqrt{E_{tr}^{(1)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \alpha_p^{(1)} \alpha_p^{(1)} \int_0^{T_f} w_q^{(1)}(t - c_j^{(1)}T_c - \tau_p^{(1)}) w_q^{(1)}(t - c_j^{(1)}T_c - \tau_p^{(1)}) dt \\ &= \sqrt{E_{tr}^{(1)}} N_s \sum_{p=1}^{L_p} (\alpha_p^{(1)})^2. \end{aligned} \quad (3.22)$$

It is observed that the received energy in the multipath channel increases with increase in the number of RAKE fingers in the receiver. It improves system performance at the cost of system complexity. Therefore, in practice a moderate number of RAKE fingers is used to combat both the complexity and system performance.

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Figure 3.3: (a) A simple Transmitter structure for N -bit OOK-PSM scheme (b) Receiver structure for combined N -bit OOK-PSM scheme (c) RAKE receiver structure for q^{th} ($q = 0, 1, \dots, N - 1$) correlator.

Chapter 3. M -ary Signaling based on OOK and PSM Modulation Schemes**3.5.2 Inter Pulse Interference (IPI)**

IPI is related to interference with the same order pulses and depends on the number of multipath in the signal but is not concerned with the number of multiple users in the system. The IPI, $IPI_q^{(1)}$, of user 1 in the q^{th} correlator can be expressed from Eq. (3.21) as

$$\begin{aligned} IPI_q^{(1)} &= \sqrt{E_{tr}^{(1)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} \int_0^{T_f} w_q^{(1)}(t - c_j^{(1)}T_c - \tau_l^{(1)}) w_q^{(1)}(t - c_j^{(1)}T_c - \tau_p^{(1)}) dt \\ &= \sqrt{E_{tr}^{(1)}} N_s \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} R_{qq}^{(1,1)}(\Delta) \end{aligned} \quad (3.23)$$

where $R_{qq'}^{(k,k)}(\Delta) = \int_0^{T_f} w_q^{(k)}(t) w_{q'}^{(k)}(t - \Delta) dt$, $q' \in \{0, 1, \dots, N-1\}$, and $\Delta = (\tau_l^{(1)} - \tau_p^{(1)})$.

The corresponding variance of IPI is σ_{IPI}^2 and can be expressed as

$$\sigma_{IPI}^2 = E_{tr}^{(1)} N_s T_f^{-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \sum_{\substack{p'=1 \\ p' \neq p}}^{L_p} \sum_{\substack{l'=1 \\ l' \neq l}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} \alpha_{p'}^{(1)} \alpha_{l'}^{(1)} X(\Delta'') \quad (3.24)$$

where $\Delta'' = (\tau_l^{(1)} - \tau_p^{(1)} - \tau_{l'}^{(1)} + \tau_{p'}^{(1)})$ and $X(\cdot)$ is the correlation function of $R_{qq}(\cdot)$. The IPI degrades system performance when systems are asynchronized and it is reduced when systems are synchronized with orthogonal pulses. Therefore, designing orthogonal pulses with short pulse width is a challenging task for OOK-PSM modulation scheme. The IPI increases with increase in the number of transmitted pulses in the same time frame, T_f . Higher level modulation scheme increases the number of transmitted pulses in the same time frame. Therefore, higher level OOK-PSM schemes are not suitable for data transmission in the presence of several multipaths.

3.5.3 Multi Pulse Interference (MPI)

MPI is related to interference with different order pulses and depends on the number of multipath. It does not depend on the number of users in the system. The MPI, $MPI_q^{(1)}$

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of user 1 in the q^{th} correlator can be written from Eq. (3.21) as

$$\begin{aligned}
 MPI_q^{(1)} &= \sqrt{E_{tr}^{(1)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \sum_{m=1}^N \alpha_p^{(1)} \alpha_l^{(1)} \int_0^{T_f} w_q^{(1)}(t - c_j^{(1)}T_c - \tau_l^{(1)}) w_q^{(1)}(t - c_j^{(1)}T_c - \tau_p^{(1)}) dt \\
 &= \sqrt{E_{tr}^{(1)}} N_s \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \sum_{m=1}^N \alpha_p^{(1)} \alpha_l^{(1)} R_{qm}^{(1,1)}(\Delta). \tag{3.25}
 \end{aligned}$$

The corresponding variance of MPI is σ_{MPI}^2 and can be expressed as

$$\sigma_{MPI}^2 = E_{tr}^{(1)} N_s T_f^{-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \sum_{\substack{p'=1 \\ p' \neq p}}^{L_p} \sum_{l'=1}^{L_p} \sum_{m=1}^N \sum_{\substack{m'=1 \\ m' \neq m}}^N \alpha_p^{(1)} \alpha_l^{(1)} \alpha_{p'}^{(1)} \alpha_{l'}^{(1)} X'(\Delta'') \tag{3.26}$$

where $X'(\cdot)$ is the correlation between $R_{qm}^{(1,1)}(\cdot)$ and $R_{q'm'}^{(1,1)}(\cdot)$. Since $R_{qm}^{(1,1)}(\cdot)$ is crosscorrelation values of two different order pulses q and m , MPI tends to be zero for perfectly orthogonal pulses for synchronized systems irrespective of number of multipaths. However, MPI degrades the system performance for higher crosscorrelation values of orthogonal pulses in both partially synchronized and asynchronous systems.

3.5.4 Multiple Access Interference (MAI)

Under ideal conditions, the receiver is not affected by the presence of multiple transmissions for perfectly orthogonal TH-codes. In reality, however, the systems do not achieve ideal synchronization and codes lose orthogonality due to different propagation delays due to different paths. The receiver might not be able to remove undesired signals completely and as a consequence system performance is affected by MAI [59, 120, 125]. The MAI term, $MAI_q^{(1)}$, of OOK-PSM schemes for the N_u users can be expressed as

$$\begin{aligned}
 MAI_q^{(1)} &= \sum_{k=2}^{N_u} \sqrt{E_{tr}^{(k)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \alpha_l^{(k)} \alpha_p^{(1)} \int_0^{T_f} w_q^{(k)}(t - c_j^{(k)}T_c - \tau_l^{(k)}) w_q^{(1)}(t - c_j^{(1)}T_c - \tau_p^{(1)}) dt \\
 &= N_s \sum_{k=2}^{N_u} \sqrt{E_{tr}^{(k)}} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \alpha_l^{(k)} \alpha_p^{(1)} R_{qq'}^{(1,k)}(\Delta') \tag{3.27}
 \end{aligned}$$

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where $\Delta' = (c_j^{(1)} - c_j^{(k)})T_c - (\tau_p^{(1)} - \tau_l^{(k)})$. The variance of MAI is σ_{MAI}^2 and can be expressed as

$$\sigma_{MAI}^2 = N_s T_f^{-1} \sum_{k=2}^{N_u} E_{tr}^{(k)} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \sum_{p'=1}^{L_p} \sum_{l'=1}^{L_p} \alpha_p^{(1)} \alpha_l^{(k)} \alpha_{p'}^{(1)} \alpha_{l'}^{(k)} X''(\Delta''') \quad (3.28)$$

where $\Delta''' = (\tau_l^{(1)} - \tau_p^{(k)} - \tau_{l'}^{(1)} + \tau_{p'}^{(k)})$, and $X''(\cdot)$ is the sum of $X(\cdot)$ and $X'(\cdot)$. In a single user system, MAI is zero and in a multiple users system MAI is zero if TH-codes are orthogonal and users are synchronized irrespective of the pulses characteristics. However, designing synchronized systems and orthogonal TH-codes are difficult tasks for TH-UWB transceiver. Therefore, MAI can be reduced by using orthogonal pulse based modulation schemes and assigning different subsets of orthogonal pulses for different users.

$N_q^{(1)}$ is the AWGN generated by q^{th} correlator, and can be expressed as

$$N_q^{(1)} = \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \alpha_p^{(1)} \int_0^{T_f} n(t) w_q^{(1)}(t - c_j^{(1)} T_c - \tau_p^{(1)}) dt \quad (3.29)$$

The corresponding noise is

$$\sigma_N^2 = \frac{N_0 N_s (\sum_{p=1}^{L_p} \alpha_p^{(1)})^2}{2}. \quad (3.30)$$

3.5.5 Bit Error Rates

Due to different autocorrelation values for different pulses, each correlator gives different probability of error. It can easily be proved that the noise/interference terms are zero-mean Gaussian variables and so corresponding probability of error of the l^{th} correlator in the presence of IPI, MPI and MAI can be written as [123]

$$P_l = Q \left(\sqrt{\frac{(S_q^{(1)})^2}{2(\sigma_{IPI}^2 + \sigma_{MPI}^2 + \sigma_{MAI}^2 + \sigma_N^2)}} \right). \quad (3.31)$$

Since each decision is independent, the average probability of bit error is defined as

$$Pr_b = \frac{1}{N} \sum_{l=0}^{N-1} P_l \quad (3.32)$$

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where N is the total number of correlators for N -bit symbols transmission. The correct decision of the l^{th} correlator is $1 - P_l$. The received symbol is perfect if all correlators take correct decisions. Since decisions are independent, the correct decision for a symbol can be defined as

$$P_c = \prod_{l=0}^{N-1} (1 - P_l) \quad (3.33)$$

and $(1 - P_c)$ gives the error probability N bit-symbols.

3.6 Simulation Studies

In this section, simulation results for 2-bit symbols i.e for 4-ary PSM schemes and its combined schemes are analyzed. The simulation studies are conducted in AWGN and IEEE802.15.3a UWB multipath channel model under the assumption of perfect synchronization between transmitter and reference receiver. The present simulation studies assume use of a fixed threshold level. Since threshold value is insensitive to number of users, a fixed threshold value $\theta_{th} = \gamma\sqrt{E_{tx}}$ has been chosen rather than selecting optimum threshold values adaptively, where γ is normalized threshold value. For multipath channel, a standard method based on [126] is used to obtain γ . The present simulation studies uses $\gamma = 0.5$ for AWGN channel and $\gamma = 0.75$ for CM1 channel. All simulation studies use MHPs and PSWFs orthogonal pulses without using any coding or guard interval [29, 30, 36]. The simulation parameters are given in Table 3.2.

3.6.1 AWGN Channel

The performance of 2-bit OOK-PSM scheme in AWGN channel is shown in Figs. 3.4 and 3.5 using MHPs and PSWFs respectively. It can be seen that combined modulation schemes outperform PSM scheme. Due to fewer pulses and receiver correlators than those used in PSM, the proposed scheme provides low complexity for the system design. It does not require large number of orthogonal pulses and receiver correlators for higher-level

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Table 3.2: Simulation parameters

Simulation	parameters
Channel model	AWGN and CM1 of IEEE802.15.3a channel model [31]
Chip duration (T_c)	15 ns
Duration of time frame(T_f)	60 ns
Number of Time frame (N_s)	4
Sampling frequency	60GHz
Pulses	MHPs, PSWFs
Pulse width (T_p)	0.5 ns, 0.7 ns
Pulse position for 2PPM-PSM	0.5 ns
Number of users	4,10,30,60

modulation schemes. Since it uses fewer orthogonal pulses for transmission, it creates fewer spectral spikes resulting in better coexistence with overlapping NB systems.

Comparing with BPSM scheme, the proposed scheme requires fewer pulses and receiver correlators for same data rates. For a 2-bit modulation scheme, OOK-PSM shows nearly the same performance as that of BPSM. Due to limited correlation properties of higher order orthogonal pulses, the proposed scheme performs better than BPSM when number of bits per symbol is increased.

From Figs. 3.4 and 3.5, it is shown that BPSK-PSM results in slightly better performance than OOK-PSM scheme. Since performance difference between conventional BPSK and OOK is 3 dB in AWGN channel, it is also expected that performance of BPSK-PSM should give 3 dB over OOK-PSM scheme. But as number of bits per symbol increases, the performance difference between BPSK-PSM and OOK-PSM decreases. It is because of the increase in average number of pulses in the BPSK-PSM modulation when compared with OOK-PSM. For example, in 2-bit BPSK-PSM scheme, each symbol requires two orthogonal pulses, whereas OOK-PSM requires one pulse except for symbol 11 which requires two pulses. This difference in number of average pulses is more visible when the number of bits per symbol is increased. Though pulses are said to be orthogonal, they are nonorthogonal in the finite time interval for the timing jitter [36]. This

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Figure 3.4: Performance of various modulation schemes for 2-bit symbols transmission in AWGN for orthogonal pulses based on MHPs .

Figure 3.5: Performance of various modulation schemes for 2-bit symbols transmission in AWGN for orthogonal pulses based on PSWFs.

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leads to a degradation in performance of BPSK-PSM when the number of average pulses is more within the same time interval.

It can be seen that the proposed scheme gives nearly the same performance with 2PPM-PSM scheme. However, due to the presence of nonorthogonal pulse position in 2PPM-PSM scheme, the ISI and MAI issues resurface in 2PPM-PSM modulation scheme, which can severely affect system performance in multipath environment. Maintaining orthogonality of constellation vector is more important for better system performance. This requires coded modulation and memory in the receiver to achieve orthogonality of constellation vectors [29, 127]. Since 2PPM-PSM scheme uses PPM, PAM, and PSM, recovering signals at the receiver is complicated in the presence of multipath. In addition, the complexity of system design for 2PPM-PSM is increased for the presence of constellation matrix, map decision vector and distance comparator vector in the receiver [128].

In Fig. 3.6, the simulation performance in multiple user environment is presented in an AWGN channel. It has been seen that the performance decreases with increase in the number of bits per symbol. It is largely because of the increase in the number of orthogonal pulses used for signal transmission. Since pulses are not strictly orthogonal within the finite interval, the interference among these pulses lead to performance degradation. However, across multiple users, the performance degradation is very minimal. On the other hand, due to the presence of a single pulse in 1-bit transmission, performance difference with respect to users is more. But in 2-bit and 3-bit schemes performance difference with respect to number of users is less. It is because, these schemes use multiple orthogonal pulses which reduce crosscorrelation terms in MAI in synchronized systems. The simulation results justify lower MAI than that in the single pulse systems as shown in Eq. (3.10).

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Figure 3.6: Performance of 1-bit, 2-bit and 3-bit symbols transmission of the OOK-PSM scheme for different numbers of users in AWGN.

3.6.2 Multipath Channel Model

Since orthogonal pulses are more sensitive in multipath channel, it is required to analyze the performance of PSM modulation and its combined scheme in the presence of multipath environment. The simulation studies are conducted in IEEE802.15.3a multipath channel under the assumption of perfect synchronization between transmitter and reference receiver [31]. The present simulation studies assume channel models CM1 (0 – 4 m LOS) from IEEE 802.15.3a channel model [31]. Channel estimation is done using selective RAKE receiver and MRC. The number of significant paths is decided by taking all paths within 10 dB of the strongest paths and 85% energy of the multipath components. To collect all these multipaths a RAKE receiver is employed in the proposed system. It is assumed that the transmitted pulse average interval is much longer than the pulse duration. In channel estimation, only distinguishable paths are selected.

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Figs. 3.7 and 3.8 show the performance of combined PSM schemes using MHPs and PSWFs respectively, where number of RAKE fingers is 17. The PSWFs give better performance than MHPs in the presence of multipath. Fig.3.9 shows the performance for 23 RAKE fingers by using MHPs. The proposed OOK-PSM shows better performance than PSM scheme, but BPSK-PSM gives better performance compared to other modulation schemes. Since *zero* is represented by pulse off, OOK complexity is nearly half of that in any other modulation scheme. The combined form OOK-PSM has $(3/4)^{th}$ complexity of that of BPSK-PSM. Therefore, the suboptimal performance of M -ary OOK-PSM can be justified in terms of implementation complexity.

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Figure 3.7: Performance of various modulation schemes for 2-bit symbols transmission in a multipath environment. The receiver assumes a RAKE combination by considering all paths within -10dB of the strongest path. The scheme uses orthogonal pulses based on MHPs.

Figure 3.8: Performance of various modulation schemes for 2-bit symbols transmission in a multipath environment. The receiver assumes a RAKE combination by considering all paths within -10dB of the strongest path. The scheme uses orthogonal pulses based on PSWFs.

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Figure 3.9: Performance of various modulation schemes for 2-bit symbols transmission in a multipath environment. Path selection is based on 85% signal energy and pulses are based on MHPs.

3.7 Summary

In this chapter, a combined modulation scheme has been proposed for low cost system design. The scheme proposed is a combination of OOK and PSM schemes. The TH-UWB system model for N -bit OOK-PSM scheme and corresponding mathematical analysis have been presented in details. Two different receiver structures based on correlation receiver and RAKE receiver have been presented for AWGN and multipath channels respectively. The MAI, IPI and MPI have been analyzed in detail to detect and estimate signal in efficient way. The system performance for N -bit OOK-PSM scheme has been provided for the correlation and RAKE receiver in the presence of multiple users. The simulation results have been provided for different sets of orthogonal pulses under different channel conditions. The performance of M -ary OOK-PSM has been compared with other pulse based modulation schemes.

Chapter 4

M-ary OPPM-BPSM Modulation for Improving Data Rates in TH-UWB Radio Systems

4.1 Introduction

One of the major drawbacks of TH-UWB systems is the low data rate. The data rate of systems can be increased by increasing the level of modulation schemes. Several higher level modulation schemes such as *M*-ary PPM, *M*-ary PAM and *M*-ary PSM have been proposed to enhance data rates and to improve system performance. Increasing the level of modulation schemes of *M*-ary signaling reduces the constellation distance and noise tolerance of the system. However, constellation distance for combined *M*-ary signaling is more than the equivalent single *M*-ary scheme. Several combined modulation schemes such as *M*-ary BPPM, *M*-ary PPM-PAM, *M*-ary BPSM and *M*-ary OOK-PSM have been proposed to enhance data rates further. However, some of the combined schemes such as *M*-ary OOK-PSM do not support higher level modulations in the multipath propagation when $M \geq 3$. To solve this problem, the present work suggests an alternative *M*-ary combined modulation scheme for TH-UWB systems. The proposed scheme is a combination of OPPM and BPSM and is referred to as *M*-ary OPPM-BPSM scheme. The system model, performance and capacity of *M*-ary OPPM-BPSM modulation scheme are

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presented in detail. The corresponding mathematical analysis and simulation results are provided to show the modulation efficiency in terms of performance and system capacity.

This chapter is organized as follows: A brief discussion of M -ary signaling such as M -ary PPM, M -ary PAM, M -ary PSM, M -ary BPPM, M -ary PPM-PAM, M -ary BPSM, and M -ary BPSK-PSM for higher level modulation schemes are described in section 4.2. Section 4.3 describes the proposed modulation scheme and its mathematical model of transmitted signal. Section 4.4 presents the probabilistic approach of OPPM-BPSM in the presence of multiple users. Section 4.5 provides the RAKE receiver structure and theoretical analysis of ISI and MAI in the presence of multiple users in a multipath environment. Section 4.6 discusses channel capacity of M -ary OPPM-BPSM scheme. Section 4.7 provides the simulation results under different scenarios. Finally, a summary of this chapter is provided in section 4.8.

4.2 M -ary Modulation Schemes for TH-UWB Radio Systems

As stated earlier, the system performance of the higher level modulation schemes decreases with increase in the level of modulation schemes [36, 129, 130]. The complexity of the system also increases with increase in the level of modulation schemes [131, 132]. For example, M -ary PPM increases the ISI and needs more accurate synchronization for higher value of M . Similarly, M -ary PAM cannot be used for short range communication when M is large and M -ary PSM increases the system complexity with increase in M [68, 71].

To address these problems, several combined modulation schemes such as M -ary BPPM and M -ary PPM-PAM have been proposed for higher level modulation schemes. These combined schemes increase system data rates with improved system performance [133–135]. However, some of these modulation schemes still degrade performance due

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to presence of ISI and MAI under multiple users in multipath environments. Owing to robustness against ISI and MAI, orthogonal pulse based and its combined forms such as M -ary BPSM, M -ary BPSK-PSM, M -ary 2PPM-PSM and M -ary OOK-PSM have been proposed to reduce system complexity and to improve system performance [29, 30, 136–139]. However, M -ary BPSK transmitter and receiver still require large set of orthogonal pulses and correlators respectively. Although M -ary OOK-PSM and M -ary BPSK-PSM schemes require fewer orthogonal pulses and correlators, these cannot be used for higher level modulation scheme due to poor system performance [121].

In order to address these problems, a combined modulation scheme for TH-UWB systems is proposed in this work [140]. The proposed scheme is a combination of OPPM and BPSM. In order to construct M -ary OPPM-BPSM signal, $L (= 2^l)$ pulse positions and $N (= 2^{k-l-1})$ biorthogonal pulses are used. The selection of number of pulse positions and pulses depends on the requirement of system performance and the availability of orthogonal pulses with estimable autocorrelation properties. It allows one to increase the number of bits per symbol and, consequently, reduces the duration of pulse repetition interval by increasing the number of orthogonal pulses. Data rates can be improved further by incorporating several orthogonal pulses in the same pulse repetition interval. It also reduces the system complexity by half of M -ary PPM scheme by introducing antipodal version of orthogonal pulses. The proposed transmission scheme is investigated through mathematical analysis and simulation studies are conducted in the presence of multiple users in multipath environment. The BER performance and data rates of the same level scheme depend on the number of pulse positions. The performance of M -ary OPPM-BPSM scheme is compared with that of the existing PSM and its combined modulation schemes. It is observed that M -ary OPPM-BPSM performs better compared to M -ary BPPM, M -ary BPSM, M -ary BPSK-PSM and M -ary OOK-PSM.

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4.3 System Model of OPPM-BPSM Scheme

A modulation scheme combining OPPM and BPSM is proposed for higher level modulation to improve system data rates. The proposed system uses an alternative method for selecting the pulse amplitude, pulse positions and order of orthogonal pulses of a symbol of M -ary OPPM-BPSM scheme. A mathematical analysis for the proposed approach together with system details are provided in the following section.

4.3.1 The Modulation Scheme

In order to transmit M symbols, one has to use L orthogonal time shift positions and N biorthogonal pulses where $M = 2^k$, $L = 2^l$, $N = 2^{k-l-1}$, $k > 1$ and $0 \leq l \leq k - 1$. Antipodal pulses are chosen to smooth the PSD of TH-UWB signal and improve its co-existence ability with NB systems without any degradation in system performance [116]. These bi-phase pulses reduce the number of correlators in the receiver. Further, the system complexity is reduced by half when compared with a scheme that uses a combination of L pulse positions and N orthogonal pulses. Theoretically, $N \times L$ correlators are required in the receiver, however, only N correlators are sufficient for N orthogonal pulses, and, L delays for each pulse can be implemented in software. Therefore, hardware implementation cost of the proposed scheme is lower than the M -ary BPSM scheme for $N < M$.

By changing the number of pulse position and orthogonal pulses, one can construct a wide variety of symbols. For example, M -ary BPPM scheme can be designed by using $M/2$ pulse positions and one bi-phase pulse whereas an M -ary BPSM scheme requires one pulse position and $M/2$ biorthogonal pulses [133]. An 8-ary OPPM-BPSM signaling can be designed by using 2 pulse positions and 2 biorthogonal pulses. The proposed scheme insures relatively constant power envelope for transmitted symbol irrespective of the number of positions and pulses. The proposed multidimensional scheme increases

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Euclidean distance of the transmitted signal and so the power efficiency increases without affecting the signal bandwidth. Since each position is able to transmit multiple orthogonal pulses, it does not require wider chip duration and wider time frame than M -ary OPPM scheme. By introducing N biorthogonal pulses in N pulse positions, data rate can be increased by $(2 + 1/k)$ times compared to that of 2^k -ary PPM scheme.

As mentioned above, the number of required orthogonal pulses decreases with increase in the number of pulse positions and vice versa. If ISI is not an issue in the presence of multipath, larger number of pulse positions can be used to reduce number of pulses and/or magnitude of the spectral lines [29, 116]. The proposed multidimensional scheme keeps fixed Euclidean distance between transmitted symbols so that the power efficiency increases without affecting the signal bandwidth [141]. On the other hand, if duration of pulse positions is fixed, the pulse repetition interval can be increased by increasing the number of pulse positions. This reduces data rates of the system but allows reduction of duty cycle in the transmitted signal. The duration of pulse repetition interval is dependent on the selection of number of pulse positions and pulses. The variation in pulse repetition duration varies system data rates for the same value of M . Therefore, this modulation scheme can be used for multivariate data rate systems. It can be used in adaptive modulation systems by changing number of pulse positions and pulses instead of changing the level of modulation schemes. However, a tradeoff between number of pulse position and pulses is required for high data rate and good system performance.

4.3.2 Signal Construction for Adaptive Systems

An adaptive UWB system based on M -ary modulations can be proposed to satisfy various constraints such as BER, data rates, energy dissipation, etc. [98, 142]. Adaptive modulation systems use several order modulation schemes with the help of lookup tables for different levels modulations. Generally, a k -bit transmission requires k lookup tables

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to map the symbol into pulse position and order of pulse. If a system uses 3 level schemes adaptively (e.g. $k - 1$, k , and $k + 1$ bit), a total of $3k$ lookup tables is necessary in the transmitter to transmit the signal and $3k$ tables in the receiver to detect the data. This needs memory in the transceiver architecture and increases the system complexity. It also requires time to retrieve the respective entries from the lookup table.

In the proposed system, an alternative method is used to select the position and pulse for a symbol instead of lookup tables. This process requires $k + 1$ bit register and uses 3 shift sequence registers of different lengths to map a symbol into the pulse position, order of pulse and amplitude. If number of pulse positions is $L(> 1)$ and number of orthogonal pulses is $N(> 1)$ for k -bit symbol transmission, the corresponding k -bit register is divided into three parts. The first part (l bits long, where $l = \log_2 L$) decides the position of the pulse. The next $(k - l - 1)$ bits (corresponding to $\log_2 N$) decide the order of the pulse. The final bit decides the amplitude of the pulse. For $L = 1$, modulation uses only one position and for $N = 1$ modulation scheme uses only one pulse. These are referred to as M -ary BPSM and M -ary BPPM schemes respectively.

Consider a typical example for the proposed system. If a system uses 3 modulation schemes such as 4-ary, 8-ary and 16-ary adaptively, it requires a 4-bit register. For 16-ary scheme with 2 positions (0^{th} and 1^{st} position) and 4 orthogonal pulses (0^{th} , 1^{st} , 2^{nd} and 3^{rd} order), the symbol 1101 is transmitted at 1^{st} position by assigning 2^{nd} order pulse with a positive amplitude whereas symbol 0110 is transmitted at 0^{th} position by 3^{rd} order pulse with a negative amplitude. If the system changes level from 16 to 8, the most significant bit will be *zero* and rest of bits are information bits. Fig. 4.1 shows the structure of the transmitter of the proposed adaptive M -ary OPPM-PSM UWB system. In the receiver side, correlator decides the amplitude (sign of the pulse), pulse position and order of the orthogonal pulse. A reverse procedure corresponding to transmitter is applied to get back the original symbol.

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Figure 4.1: Generalized block diagram of adaptive transmitter for contracting the transmitted signal of OPPM-BPSM modulation scheme.

4.3.3 Transmitted Signal for OPPM-BPSM

In TH-UWB transmission, the combined orthogonal PPM and biorthogonal PSM signal of the k^{th} user for i^{th} symbol is defined as

$$s_i^{(k)}(t) = \sum_j \sqrt{E_{tx}^{(k)}} d_{(i\%2)+1}^{(k)} w_{[i/2]\%N}^{(k)}(t - jT_f - c_j^{(k)}T_c - \delta_{[i/L]}^{(k)}) \quad (4.1)$$

where $i = 0, 1, \dots, M-1$, $d_{(i\%2)+1}^{(k)} \in \{\pm 1\}$ is the sign of amplitude of the pulse $w_{[i/2]\%N}^{(k)}(t)$ at position $\delta_{[i/L]}^{(k)}$ of the k^{th} user, T_p is the pulse duration. Pulses have finite energy and are normalized so as to insure equal energy per transmission, that is $\int_{-\infty}^{+\infty} |w_l(t)|^2 dt = 1$, where $0 \leq l \leq N-1$. $E_{tx}^{(k)}$ is the energy of k^{th} user. Pulse position and order of pulse for the i^{th} symbol are $[i/L]$ and $[i/2]\%N$ respectively, where $[\cdot]$ is the floor operator and $\%$ is the remainder. The pulse repetition interval T_f of a user is divided into N_h time slots of length T_c where $N_h T_c \leq T_f$, $\{c_j^{(k)}\}$ is a pseudorandom TH code sequence and $0 \leq c_j^{(k)} < N_h - 1$. Index j is used for representing the number of pulse repetition intervals for each symbol. For simplicity it is assumed that $\delta_q^{(k)} = \delta_q$ and $\delta_q = (q-1)\delta$ for all users, where $1 \leq q \leq L$. A guard time (g_1) is inserted to reduce ISI for longer channel delay spread. The resulting time shift is written as $\delta = T_p + g_1$. The bit rate of the system is $R_b = \log_2(N \times L \times 2)/(N_s T_f)$.

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4.4 Probabilistic Approaches of OPPM-BPSM Scheme Under AWGN channel

Without loss of generality, it is assumed that orthogonality of pulses is maintained despite the differentiating effect of the transmitter and receiver antennae. Since the symbol is unknown at the receiver, ignoring i at the left hand side of Eq.(4.1) and the received signal of the user 1 can be written as

$$r(t) = s^{(1)}(t - \tau_1) + \sum_{k=2}^{N_u} s^{(k)}(t - \tau_k) + n(t) \quad (4.2)$$

where τ_k is delay of user k , N_u is the total number of users, $n(t)$ is AWGN, and $s^{(k)}(t)$ can be expressed as

$$s^{(k)}(t) = \sum_j \sqrt{E_{rx}} d_{(i\%2)+1}^{(k)} w_{[i/2]\%N}^{(k)}(t - jT_f - c_j^{(k)}T_c - \delta_{[i/L]}^{(k)}) \quad (4.3)$$

where E_{rx} is received power. It is assumed that $\tau_k - \tau_1 = 0, \forall k$ in synchronized systems [79]. N orthogonal pulses are required at each pulse position for detecting a desired symbol. Therefore, $N \times L$ number of correlators or matched filters are required in the receiver to detect $N \times L \times 2$ symbols. The largest magnitude of correlators and the sign of this magnitude are used to detect a possible transmitted symbol. The corresponding receiver structure is shown in Fig. 4.2. The decision statistics for user 1 in m^{th} correlator is

$$\begin{aligned} \arg \max_m \alpha_m &= \sum_{j=0}^{N_s-1} \int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} r(t) \phi_m^{(1)}(t) dt \\ &= \begin{cases} s_m^{(1)} + N_{MAI,m} + N_{G,m} & , signal \\ N_{MAI,m} + N_{G,m} & , no signal \end{cases} \end{aligned} \quad (4.4)$$

where $m = 1, 2, \dots, M/2$, $s_m^{(1)}$ is desired signal of user 1, and the template signal at $m^{th}(= i' \times l)$ correlator is defined as

$$\phi_m^{(1)}(t) = w_{i'}^{(1)}(t - jT_f - c_j^{(1)}T_c - \delta_l^{(1)} - \tau_1) \quad (4.5)$$

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where $i' = \{1, 2, \dots, N\}$ and $l = \{1, 2, \dots, L\}$, and MAI and noise received by k^{th} user at m^{th} correlator is written as

$$N_{MAI,m} = \sum_{j=0}^{N_s-1} \sum_{k=2}^{N_u} N_j^{(k)} \quad (4.6)$$

where the interference at m^{th} correlator from the k^{th} user and j^{th} frame is

$$N_{mj}^{(k)} = d_{(i' \% 2)+1}^{(k)} \sqrt{E_{rx}} \sum_{j=0}^{\infty} \int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} w_{i'}^{(1)}(t - jT_f - c_j^{(1)}T_c - \delta_l^{(1)} - \tau_1) \times w_i^{(k)}(t - jT_f - c_j^{(k)}T_c - \delta_l^{(k)} - \tau_k) dt \quad (4.7)$$

and the noise component for the m^{th} correlator is

$$N_{G,m} = \sum_{j=0}^{N_s-1} \int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} n(t) \phi_m^{(1)}(t) dt. \quad (4.8)$$

In order to obtain a closed-form expression for the probability of error of this system, the characteristic function of $N_{MAI,m}$ and $N_{G,m}$ needs to be evaluated. $N_{mj}^{(k)}$ can be calculated as [143]

$$N_{mj}^{(k)} = \begin{cases} 0 & \text{if } c_j^{(1)} \neq c_j^{(k)}, \\ 0 & \text{if } c_j^{(1)} = c_j^{(k)} \ \& \ \delta_l^{(1)} \neq \delta_l^{(k)}, \\ 0 & \text{if } c_j^{(1)} = c_j^{(k)} \ \& \ \delta_l^{(1)} = \delta_l^{(k)} \ \& \ w_{i'}^{(1)}(t) \neq w_{i'}^{(k)}(t), \\ d_{(i' \% 2)+1} \sqrt{E_{rx}} & \text{if } c_j^{(1)} = c_j^{(k)} \ \& \ \delta_l^{(1)} = \delta_l^{(k)} \ \& \ w_{i'}^{(1)}(t) = w_{i'}^{(k)}(t). \end{cases} \quad (4.9)$$

Therefore probability density function of $N_{mj}^{(k)}$ can be written as

$$\begin{aligned} f_{mj}^{(k)}(x) &= (\alpha + \beta + \gamma) \delta_D(x) + \eta \delta_D(x - x_M) \\ &= \zeta \delta_D(x) + \eta \delta_D(x - x_M) \end{aligned} \quad (4.10)$$

where $\alpha = (N_u^2 - 1)/N_u^2$, $\beta = (L^2 - 1)/N_u^2 L^2$, $\gamma = (N^2 - 1)/N_u^2 L^2 N^2$, $\eta = 1/N_u^2 L^2 N^2$, $x_M = d_{i' \% 2} \sqrt{E_{rx}}$, $\zeta = \alpha + \beta + \gamma$ and $\delta_D(\cdot)$ is the Dirac delta function. The characteristic function (CF) of an interference signal from the k^{th} user in the output of m^{th} correlator on the j^{th} frame is

$$\phi_{mj}^{(k)}(w) = \zeta + \eta e^{w x_M}. \quad (4.11)$$

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Figure 4.2: Correlation receiver structure for the OPPM-BPSM modulation scheme (a) Block diagram for bank of correlators for the different pulse position. (b) Block diagram of correlators for q^{th} pulse position

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The CF of all interfering users on j^{th} frame in the output of m^{th} correlator is given by

$$\phi_{MAI,j}(w) = \prod_{k=2}^{N_u-1} \phi_{mj}^{(k)}(w) = (\phi_{mj}(w))^{N_u-1}. \quad (4.12)$$

Since MAI and AWGN noise are independent, the interfering CF for MAI and AWGN in a symbol at the output of m^{th} correlator is given by

$$\begin{aligned} \phi(w) &= \prod_{j=0}^{N_s-1} (\phi_{MAI,j}(w) \phi_{G,j}(w)) \\ &= \phi_{MAI}(w) \phi_G(w) \\ &= (\zeta + \eta e^{wx_M})^{N_s(N_u-1)} e^{(-N_0 N_s w^2/4)} \end{aligned} \quad (4.13)$$

where $\phi_{MAI}(w)$ and $\phi_G(w)$ are the CF of $N_{MAI,m}$ and $N_{G,m}$ respectively. If it is assumed that the distribution of interference at the output of the correlator is Gaussian, the BER is written as

$$P_b = Q \left(\sqrt{\frac{E_{rx} N_s^2}{\sigma^2}} \right) \quad (4.14)$$

where variance of total noise can be calculated from (4.13) and is given by

$$\sigma^2 = N_s(N_u - 1)E_{rx} \frac{N_u^2 L^2 N^2 - 1}{N_u^2 L^2 N^2} - \frac{N_0 N_s}{2}. \quad (4.15)$$

4.5 Proposed Scheme in Multipath Channel Model

This section investigates system models of M -ary OPPM-BPSM modulation scheme by providing receiver structure and several interference issues. The system performance with the channel capacity are also studied in detail.

4.5.1 Received Signal

In this discussion, the proposed scheme is analyzed in a multipath channel. Without loss of generality, it is assumed that orthogonality of pulses is maintained despite the differentiating effect of the transmitter and receiver antennal. For simplicity, it is also

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assumed that signal is transmitted by using i^{th} ($0 \leq i \leq N - 1$) order pulse in the q^{th} ($0 \leq q \leq L - 1$) pulse position. The corresponding receiver structure is shown in Fig. 4.3. Fig. 4.3 (a) shows banks of correlators for L possible pulse positions. There will be N correlators corresponding to N orthogonal pulses at each position. Fig. 4.3 (b) shows the correlator designed for 0^{th} position and i^{th} order orthogonal pulse. Therefore the signal in (4.1) is rewritten as

$$s_{iq}^{(k)}(t) = \sum_j \sqrt{E_{tx}^{(k)}} d_m^{(k)} w_i^{(k)}(t - jT_f - c_j^{(k)}T_c - \delta_q^{(k)}) \quad (4.16)$$

where $d_m \in \{-1, 1\}$. If there are N_u users and each experiences a different channel (2.15), the received signal is expressed as

$$r(t) = \sum_{k=1}^{N_u} \sum_{l=1}^{L_p} \alpha_l^{(k)} s_{iq}^{(k)}(t - \tau_l^{(k)}) + n(t) \quad (4.17)$$

where $n(t)$ is AWGN that is assumed to have a two sided power spectral density of $N_0/2$. It is assumed that reference RAKE receiver is synchronized i.e. $\tau_l^{(1)} = 0$ for l^{th} RAKE finger of user 1. N orthogonal pulses are required at each pulse position for detecting a desired symbol. Therefore, $N \times L$ correlators or matched filters are required in the receiver to detect $N \times L \times 2$ symbols. The largest magnitude of correlators and the sign of this magnitude are used to detect possible transmitted symbol. The reference signal of correlator for i^{th} order pulse in q^{th} pulse position of user 1 is expressed as

$$\phi_{iq}^{(1)}(t) = \sum_{j=0}^{N_s-1} v_i^{(1)}(t - jT_f - c_j^{(1)}T_c - \delta_q^{(1)}) \quad (4.18)$$

where N_s is the number of pulse repetition intervals for a symbol and

$$v_i^{(1)}(t) = \sum_{p=1}^{L_p} \alpha_p^{(1)} w_i^{(1)}(t - \tau_p^{(1)}) \quad (4.19)$$

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(a)

(b)

Figure 4.3: Receiver structure for the proposed system (a) Block diagram for bank of correlators for the different pulse position and different order of orthogonal pulses. (b) Block diagram of correlators for 0^{th} pulse position and i^{th} ($i = 0, 1, \dots, N - 1$) order pulse which contains L_p RAKE fingers for different delays and different weights

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4.5.2 Decision statistics

At the receiver, the reference signal, indicated as a useful signal is corrupted by mainly three additive noise components. They are ISI noise generated in the RAKE finger due to the multipath component, MAI due to presence of multiple users in the system, and thermal noise generated in the receiver antenna and the receiver circuitry. The problem of receiver design can thus be stated as follows: Finding a good when possible optimal, way for extracting the useful signal from the received signal. Solving the general problem is a complicated task leading to complex receiver structures and requiring good modeling for noise components. The combined output of first L_p paths of the correlator for i^{th} order pulse for q^{th} pulse position is written as [109, 124]

$$\begin{aligned} z_{iq}^{(1)} &= \int_{jT_f}^{(j+1)T_f} r(t)\phi_{iq}^{(1)}(t)dt \\ &= S_{iq}^{(1)} + ISI_{iq}^{(1)} + MAI_{iq}^{(1)} + N_{iq}^{(1)} \end{aligned} \quad (4.20)$$

where $S_{iq}^{(1)}$ is the useful signal, $ISI_{iq}^{(1)}$ is the ISI term, $MAI_{iq}^{(1)}$ is the MAI term and $N_{iq}^{(1)}$ is the AWGN component of user 1 in the correlator for i^{th} order pulse and for the q^{th} pulse position. In order to improve the detection performance of the system, effects of ISI and MAI have to be canceled before a symbol decision is made. In communication systems, decision feedback equalizers are commonly employed for this purpose. The main idea behind decision feedback equalization is that once a data symbol has been detected, interference induces on the following symbols is estimated and subtracted before the detection of subsequent symbols. Therefore, knowledge of the desired signal and interference noise is required to make the correct decision.

4.5.2.1 Desired Signal Energy

The transmitted signal propagates in multipath channel. To collect these multipath components, each correlator contains several RAKE fingers with different delays and

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weights. The desired multipath components are correlated with corresponding RAKE fingers, and some of these received energy are known as desired signal or useful signal. In reality, it is assumed that pulses are orthogonal in the synchronized system. The desired signal can be expressed from (4.20) as

$$\begin{aligned} S_{iq}^{(1)} &= \sqrt{E_{tr}^{(1)}} d_m^{(1)} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \alpha_p^{(1)} \alpha_p^{(1)} \int_0^{T_f} w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(k)} - \tau_p^{(1)}) \\ &\quad \times w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(1)} - \tau_p^{(1)}) dt \\ &= \sqrt{E_{tr}^{(1)}} d_m^{(1)} N_s \sum_{p=1}^{L_p} (\alpha_p^{(1)})^2. \end{aligned} \quad (4.21)$$

It is observed that the received energy in the multipath channel increases with increase in the number of RAKE fingers in the receiver, which improve system performance at the cost of system complexity. Since $d_m^{(1)} \in \{\pm 1\}$, the constellation distance in OPPM-BPSM is far from those in OOK-PSM scheme, which give better performance than OOK-PSM scheme. However, larger number of RAKE fingers increases the system complexity. Therefore, a tradeoff between performance and system complexity is required to design a reliable system in the multipath channel.

4.5.2.2 Inter Symbol Interference (ISI)

In reality, channels are not perfectly estimated and each path is not synchronized. So the decision variable is affected by other unexpected signals such as ISI. It occurs when multipath components are not received by their corresponding RAKE fingers, these are received by other RAKE fingers resulting in ISI. $ISI_{iq}^{(1)}$ is the ISI of user 1 and is expressed

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from (4.20) as

$$\begin{aligned}
 ISI_{iq}^{(1)} &= \sqrt{E_{tr}^{(1)}} d_m^{(1)} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} \int_0^{T_f} w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(1)} - \tau_l^{(1)}) \\
 &\quad \times w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(1)} - \tau_p^{(1)}) dt \quad (4.22) \\
 &= \sqrt{E_{tr}^{(1)}} d_m^{(1)} N_s \sum_{p=1}^{L_p} \sum_{\substack{l=1 \\ l \neq p}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} R_{ii}(\Delta)
 \end{aligned}$$

where in general $R_{ii'}(\Delta) = \int_0^{T_f} w_i(t) w_{i'}(t - \Delta) dt$, $i, i' \in \{0, 1, \dots, N-1\}$, and $\Delta = (\tau_l^{(1)} - \tau_p^{(1)})$.

The variance of ISI, $ISI_{iq}^{(1)}$, is σ_{ISI}^2 and can be expressed as

$$\sigma_{ISI}^2 = E_{tr}^{(1)} N_s T_f^{-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \sum_{\substack{p'=1 \\ p' \neq p}}^{L_p} \sum_{\substack{l'=1 \\ l' \neq l}}^{L_p} \alpha_p^{(1)} \alpha_l^{(1)} \alpha_{p'}^{(1)} \alpha_{l'}^{(1)} X'(\Delta'') \quad (4.23)$$

where $\Delta'' = (\tau_l^{(1)} - \tau_p^{(1)} - \tau_{l'}^{(1)} + \tau_{p'}^{(1)})$ and $X'(\cdot)$ is the correlation function of $R_{ii'}(\cdot)$ [124].

It is observed that ISI is not reduced by orthogonal pulses and its modulation schemes, but depends on the channel estimation and its delay spread. The channel delay spread cannot be controlled by modulation schemes or system design. The ISI can be reduced by increasing the duration of pulse repetition interval and this affects system data rates. Moreover, pulse repetition interval is smaller when large number of orthogonal pulses are used at the same pulse position which is not an efficient way to reduce the ISI. This is one practical limitation for higher data rate systems with good system performance.

4.5.2.3 Multiple Access Interference (MAI)

In MA systems, several users transmit signals over the same channel, pulses originating in other transmission links may collide with pulses belonging to a reference transmission, giving rise to an interference noise called MAI. $MAI_{iq}^{(1)}$ is the MAI from the $N_u - 1$ users

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and can be expressed from (4.20) as

$$\begin{aligned} MAI_{iq}^{(1)} &= d_m^{(1)} \sum_{k=2}^{N_u} \sqrt{E_{tr}^{(k)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \alpha_l^{(k)} \alpha_p^{(1)} \int_0^{T_f} w_{i'}^{(k)}(t - c_j^{(k)} T_c - \delta_{q'}^{(k)} - \tau_l^{(k)}) \\ &\quad \times w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(1)} - \tau_p^{(1)}) dt \quad (4.24) \\ &= d_m^{(1)} \sum_{k=2}^{N_u} \sqrt{E_{tr}^{(k)}} \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \alpha_l^{(k)} \alpha_p^{(1)} R_{ii'}(\Delta') \end{aligned}$$

where $R_{ii'}(\Delta')$ is the correlation between the same/different orthogonal pulses which depend on the selection of pulse waveform of user 1 and user k , and

$$\Delta' = (c_j^{(1)} - c_j^{(k)}) T_c - (\delta_q^{(1)} - \delta_{q'}^{(k)}) - (\tau_p^{(1)} - \tau_l^{(k)}) \quad (4.25)$$

where i' and q' denote order of the pulse waveform and pulse position of user k respectively. The variance of MAI, $MAI_{iq}^{(1)}$, is σ_{MAI}^2 and can be expressed as

$$\sigma_{MAI}^2 = N_s T_f^{-1} \sum_{k=2}^{N_u} E_{tr}^{(k)} \sum_{p=1}^{L_p} \sum_{l=1}^{L_p} \sum_{p'=1}^{L_p} \sum_{l'=1}^{L_p} \alpha_p^{(1)} \alpha_l^{(k)} \alpha_{p'}^{(1)} \alpha_{l'}^{(k)} X''(\Delta''') \quad (4.26)$$

where $\Delta''' = (\tau_l^{(1)} - \tau_p^{(k)} - \tau_{l'}^{(1)} + \tau_{p'}^{(k)})$. From Eq. (4.26), it is observed that MAI depends on correlation properties of orthogonal pulses, number of users and system synchronization. The term $X''(\Delta''')$ is sum of the crosscorrelation and autocorrelation of orthogonal pulses. It can be reduced by assigning different exclusive subsets of the same set orthogonal pulses. However, the receiver might not be capable of removing undesired signals completely, and as a consequence system performance is affected by MAI.

4.5.2.4 AWGN Component

$N_{iq}^{(1)}$ is the AWGN generated by correlator for i^{th} order pulse and for q^{th} pulse position, and can be expressed as

$$N_{iq}^{(1)} = \sum_{j=0}^{N_s-1} \sum_{p=1}^{L_p} \alpha_p^{(1)} \int_0^{T_f} n(t) w_i^{(1)}(t - c_j^{(1)} T_c - \delta_q^{(1)} - \tau_p^{(1)}) dt \quad (4.27)$$

The noise at each receiver for AWGN can be written as

$$\sigma_N^2 = \frac{N_0 N_s \left(\sum_{p=1}^{L_p} \alpha_p^{(1)} \right)^2}{2}. \quad (4.28)$$

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4.5.3 Performance Analysis

It is assumed that the output of correlator for i^{th} order pulse and for q^{th} pulse position, z_{iq} , is larger than the other $M/2 - 1$ correlator outputs. Since each output is corrupted by noise due to the ISI, MAI and AWGN, total noise at each correlator is defined as $\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2$. The corresponding average probability of a correct decision in the presence of ISI and MAI can be expressed as [88, 133]

$$P_c = \int_0^\infty \left(\frac{1}{\sqrt{2\pi}} \int_{-z_{iq}/\sqrt{\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2}}^{z_{iq}/\sqrt{\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2}} \exp \frac{-x^2}{2} dx \right)^{\frac{M}{2}-1} p(z_{iq}) dz_{iq} \quad (4.29)$$

where pdf of z_{iq} can be written as

$$p(z_{iq}) = \frac{1}{\sqrt{2\pi(\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2)}} \exp \left(-\frac{\left(z_{iq} - N_s \sqrt{E_{tr}^{(1)}} \sum_{p=1}^{L_p} (\alpha_p^1)^2 \right)^2}{2(\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2)} \right) \quad (4.30)$$

The probability of a symbol error for combined M -ary OPPM-BPSM is given by

$$P_M = 1 - P_c. \quad (4.31)$$

The BER of the proposed scheme can be evaluated as [88, 144]

$$P_b = \frac{2^{k-1}}{2^k - 1} P_M. \quad (4.32)$$

4.6 Channel Capacity for OPPM-BPSM Scheme

To evaluate the capacity of the OPPM-BPSM scheme, the transmitted signal in Eq. (4.1) can be rewritten as

$$s_{iqm}^{(k)}(t) = \sum_j \sqrt{E_{tx}^{(k)}} d_m^{(k)} w_i^{(k)}(t - jT_f - c_j^{(k)}T_c - \delta_q^{(k)}). \quad (4.33)$$

The corresponding transmitted signal vector s_{iqm} with a nonzero value in the (i, q) element is defined as

$$s_{iqm} = [0, \dots, d_m \sqrt{E_{tx}}, 0, \dots, 0] \quad (4.34)$$

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where $d_m \in \{\pm 1\}$, $m = 1, 2$, $\sqrt{E_{tx}}$ is the amplitude for i^{th} ($1 \leq i \leq N$) order pulse and q^{th} ($1 \leq q \leq L$) pulse position. M -ary OPPM-BPSM modulation scheme for TH-UWB has discrete-valued inputs and continuous-valued outputs, which imposes additional constraints on the capacity calculation. Therefore, the channel capacity for restricted input signal with equiprobable M -ary OPPM-BPSM constellation with no restriction on the channel output, is given from Eq. (2.30) as

$$C = \log_2 M - \frac{1}{2NL} \sum_{m=1}^2 \sum_{i=1}^N \sum_{q=1}^L \int_{\mathbf{z}} p(\mathbf{z} | \mathbf{s}_{iqm}) \log_2 \left(\frac{\sum_{n=1}^2 \sum_{j=1}^N \sum_{p=1}^L p(\mathbf{z} | \mathbf{s}_{jpm})}{p(\mathbf{z} | \mathbf{s}_{iqm})} \right) d\mathbf{z} \quad (4.35)$$

where $\mathbf{z} = [z_{11}, \dots, z_{1L}, z_{21}, \dots, z_{2L}, \dots, z_{NL}]$ and the NL -dimensional joint Gaussian distribution conditioned on $\mathbf{s}_{iqm}$ with probability density function (PDF)

$$p(\mathbf{z} | \mathbf{s}_{iqm}) = \left(\frac{1}{2\pi\sigma^2} \right)^{NL} \left(\prod_{\substack{j=1 \\ (j,p) \neq (i,q)}}^N \prod_{p=1}^L e^{-\frac{z_{jp}^2}{2\sigma^2}} \right) e^{-\frac{(z_{iq} - S_{iq})^2}{2\sigma^2}} \quad (4.36)$$

where

$$S_{iq} = \sqrt{E_{tx}^{(1)}} d_m^{(1)} N_s \sum_{p=1}^{L_p} (\alpha_p^{(1)})^2 \quad (4.37)$$

and $\sigma^2 = \sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2$. From Eq. (4.35), Eq. (4.36) and Eq. (4.37), the channel capacity can be expressed as

$$C = k - \frac{1}{2NL} \sum_{m=1}^2 \sum_{i=1}^N \sum_{q=1}^L E_{\mathbf{z} | \mathbf{s}_{iqm}} \times \log_2 \left(\sum_{n=1}^2 \sum_{j=1}^N \sum_{p=1}^L e^{-\frac{z_{iq}^2 - z_{jp}^2 + \left(z_{jp} - \sqrt{E_{tx}^{(1)}} d_n^{(1)} N_s \sum_{p=1}^{L_p} (\alpha_p^{(1)})^2 \right)^2 - \left(z_{iq} - \sqrt{E_{tx}^{(1)}} d_m^{(1)} N_s \sum_{p=1}^{L_p} (\alpha_p^{(1)})^2 \right)^2}{2(\sigma_{ISI}^2 + \sigma_{MAI}^2 + \sigma_N^2)}}} \right) \quad \text{bits / channel use} \quad (4.38)$$

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4.7 Simulation Results

In this section, simulation results for performance and capacity of OPPM-BPSM are analyzed. The simulation studies are conducted in AWGN and IEEE802.15.3a UWB multipath channel under the assumption of perfect synchronization between transmitter and reference receiver [31]. The present simulation studies assume fixed and multi mod data rate systems. The simulation parameters are given in Table 4.1.

Table 4.1: Simulation parameters

Simulation	parameters
Channel model	AWGN and CM1 of IEEE802.15.3a channel model
Order of modulation	8-ary
Chip duration (T_c)	10, 20 & 40 ns for multi mod data rate systems
Chip duration (T_c)	60 ns for fixed data rate systems
Duration of time frame(T_f)	40, 80, 160 ns
Number of Time frame (N_s)	1
Sampling frequency	60 GHz
Pulses	MHPs, PSWFs
Pulse width (T_p)	0.7 ns
Pulse position	10 ns for multi mod data
Pulse position	30 ns for 2 positions, 15 ns for 4 positions
Number of users	4

4.7.1 AWGN Channel

The performance of 8-ary OPPM-BPSM scheme in AWGN channel is shown in Fig. 4.4. The simulation system uses orthogonal pulses based on MHPs and PSWFs. The results are analyzed for different pulse positions and different order of pulses. Assuming shift position δ remains constant, the chip duration T_c and the duration of time frame T_f are varied according to number of pulse positions. Correspondingly, system data rates are also varied for the same level of modulation schemes. For example, 1 pulse position and 4 pulses scheme is referred to as simple BPSM modulation and the data rate of the system is 75 mb/s for $N_s = 1$ and $T_f = 40$ ns. Similarly system with 2 positions

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and 2 pulses scheme is referred to as OPPM-BPSM modulation and data rate of the system is 37.5 mb/s for $N_s = 1$ and $T_f = 80$ ns, and, 4 positions and 1 pulse scheme is referred to as BPPM modulation and data rate of the system is 18.75 mb/s for $N_s = 1$ and $T_f = 160$ ns. It is observed that the system with moderate data rate 37.5 mb/s gives better performance compared to system with other data rates for both MHPs and PSWFs based orthogonal pulses. However, due to the presence of ISI, the performance of these schemes is degraded in a multipath environment.

The theoretical system capacity of M -ary OPPM-BPSM scheme in AWGN channel is shown in Fig. 4.5. Since there is no ISI in AWGN channel, the system capacity for the same order modulation remains same irrespective of the number of pulse positions.

4.7.2 Multipath Channel

The proposed system is extensively simulated under different channel conditions. Simulation results of an 8-ary for different number of pulse positions and orthogonal pulses in the presence of modified IEEE 802.15.3a S-V UWB multipath channel model is discussed in this section. Channel model corresponding to line of sight (0-4m) environment (CM1) is used for this studies. The performance is analyzed by using two different sets of orthogonal pulses such as MHPs and PSWFs. Fig. 4.6-Fig. 4.8 shows the results of these simulation studies.

The performance of an 8-ary scheme obtained by using 1 position and 4 pulses, 2 positions and 2 pulses, and, 4 positions and 1 pulse. In Fig. 4.6, the number of significant paths is decided by selecting paths within 10 dB of the strongest paths. Fig. 4.7 shows the performance for RAKE receivers where paths are decided by considering 85% energy of the total multipath. The performance is evaluated for multi mode data rate system, that is, duration of pulse position is fixed and duration of pulse repetition intervals is selected according to number of positions. The data rate of 1 position, 2 positions and 4 positions

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Figure 4.4: Performance of 8-ary modulation scheme in AWGN environment for different data rate systems by using orthogonal pulses based on MHPs and PSWFs.

Figure 4.5: The theoretical system capacity for M -ary OPPM-BPSM modulation scheme in AWGN channel, where $M = 4, 8, 16$ and 32 .

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Figure 4.6: Performance of 8-ary modulation scheme in different data rate by using orthogonal pulses based on MHPs and PSWFs in a multipath channel. The receiver assumes a RAKE combination by considering all paths within -10 dB of the strongest path.

Figure 4.7: Performance of 8-ary modulation scheme in different data rate by using orthogonal pulses based on MHPs and PSWFs in a multipath channel. Path selection is based on 85% signal energy.

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are 75 mb/s, 37.5 mb/s, and 18.75 mb/s respectively, for $\delta = 10$ ns and $N_s = 1$. Since duration of pulse repetition interval increases with increase in the number of positions, the inter frame interference is reduced. Therefore, the system with more positions (4 positions 1 pulse) results in better performance than the system with 1 position and 4 pulses. However, using multiple pulse positions reduces the data rate correspondingly. On the other hand, system with 1 position and 4 pulses results in worse performance because of the speculative autocorrelation properties for higher order pulses. Therefore, number of positions and pulses can be selected adaptively based on requirements of data rate and system performance.

Fig. 4.8 shows the performance of an 8-ary scheme for the same data rate (50 mb/s). Since pulse repetition interval is fixed for all possibilities of positions and pulses, length of each pulse position is decreased when 4 pulse positions and 1 pulse are considered. The multipath signals of previous pulse positions affect at the correlators of the next pulse position resulting in performance degradation. Consequently, the simulation results in Fig. 4.8 show that a large number of pulse positions (4 positions and 1 pulse) results in performance degradation. It has been shown that moderate number of pulse positions and pulses (2 positions 2 pulses) is a better choice for an acceptable data rate and system performance.

Since autocorrelation property of 0^{th} order MHPs and 0^{th} order PSWFs is the same, 4 positions and 1 pulse gives similar performance for both the pulses. However, difference in their autocorrelation values increases with increase in order of the pulse which creates difference in their performance for use of higher order orthogonal pulses. Figs. 4.6, 4.7 and 4.8 show that PSWFs pulses give better performance than MHPs for more number of orthogonal pulses. Therefore, choice of the orthogonal pulses also influences the system performance of pulse based modulation schemes.

The theoretical system capacity of M -ary OPPM-BPSM scheme for $M = 4, 8, 16$ and 32 in multipath channel is shown in Fig. 4.9. The system capacity using same level

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Figure 4.8: Performance of 8-ary modulation scheme in the same data rate environment by using orthogonal pulses based on MHPs and PSWFs in a multipath channel. The receiver assumes a RAKE combination by considering all paths within -10 dB of the strongest path.

Figure 4.9: The theoretical system capacity for M -ary OPPM-BPSM modulation scheme in multipath channel, where $M = 4, 8, 16$ and 32 .

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modulation scheme is different before for lower E_b/N_0 (say upto 20 dB). Within E_b/N_0 , the capacity of system with 1 pulse position is higher than system with 2, 4, 8 and 16 pulse positions for 32-ary OPPM-BPSM scheme. Moreover, the capacity of system with 1 pulse position of 16-ary OPPM-BPSM scheme is higher than that of system with 16 pulse positions of 32-ary OPPM-BPSM scheme for low E_b/N_0 . Since the duration of time frame is varied with number of pulse positions, the system data rate is simply calculated as $\log_2(M)/N_s T_f$. The data rate of system with 1 pulse position of 16-ary OPPM-BPSM is 100 mb/s for $T_f = 40$ ns. However, data rate of system with 16 pulse positions of 32-ary OPPM-BPSM is 31.25 mb/s for $T_f = 160$ ns. It has been shown that the system data rate and the capacity are depend on number of pulse positions. However, system capacity for the same order modulation scheme is saturated for high E_b/N_0 (say above 25 dB), irrespective of number of pulse positions.

4.8 Summary

In this chapter, M -ary OPPM-BPSM scheme for higher data rates TH-UWB systems has been proposed. The proposed scheme is obtained by combining pulse positions and orthogonal pulses. An alternative method has been introduced to construct a signal from symbols to analog waveforms. The proposed transmission scheme and its system model have been described in detail by providing mathematical formulation. The probabilistic approach of M -ary OPPM-BPSM scheme in the presence of multiple users has been analyzed. The receiver structure and several interference issues such as MAI and ISI are analyzed in the presence of multiple users and multipath channel. The channel capacity for this modulation scheme has been provided. Finally, the simulation results have been presented to show the efficiency of performance and system capacity by using different sets of orthogonal pules.

Chapter 5

PSD of TH-UWB Systems and Its Interference with NB Systems

5.1 Introduction

Analyzing PSD of UWB systems is very important in analyzing the performance of TH-UWB systems. Since UWB systems are designed for low power detection and interception, the spectrum performance is desired to be very smooth. The more the spectral combs present in a signal, the higher the probability of UWB signal interference with other NB systems. The spectral components in the signal depend on modulation schemes and pulse waveforms. When both continuous and discrete components are present in UWB signal, they degrade the system performance of coexisting NB receivers. It is in this context, the PSD of orthogonal pulse based modulation schemes such as PSM, BPSM, OOK-PSM and OPPM-BPSM are analyzed to see the effectiveness of these modulation schemes for coexistence of TH-UWB and NB systems. This analysis is followed by a generalized pulse waveform to reduce TH-UWB interference on NB systems.

This chapter is organized as follows: Section 5.2 describes the general overview of PSD and UWB interference issues on NB systems. The PSD of orthogonal pulse based modulation schemes such as PSM, BPSM and OPPM-BPSM schemes are analyzed in section 5.3. A generalized pulse waveform and its coexistence capability with IEEE802.11a

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systems is presented in section 5.4. The simulation results of UWB interference power at IEEE802.11a device for a UWB system based on generalized pulse waveform are presented in section 5.6. Finally, a summary of this chapter is given in section 5.7.

5.2 PSD and UWB Interference Issues

Minimizing interference power of UWB systems is one of the primary concerns in heterogeneous radio environment where UWB radio system has to coexist with other NB systems. In general, UWB technology is proposed for a large band of unlicensed spectrum. The usage of UWB systems degrade performance of other systems which lie in the same frequency band, or operate in nearby frequency band. Understanding of UWB interference on NB system requires a study of both UWB and NB systems design [32,69,117,145]. The system model of both systems and their mitigation techniques have been discussed in chapter 2. In this chapter, TH-UWB signal model for orthogonal pulse based modulation schemes and a generalized pulse waveform have been considered to investigate the spectral characteristics of TH-UWB signal.

Extensive studies have shown that PSD of conventional TH-UWB signal depends on the pulse waveform, TH-code, pulse amplitude and pulse position of a signal [62,112,146,147]. The PSD of TH-UWB signal consists of continuous and discrete PSD components. The continuous component depends on the ESD of pulse waveform and is constant for a specific pulse waveform. The positions of discrete components are designed based on TH-code and pulse position of a signal. The amplitudes of discrete components are decided by ESD of pulse waveform and the pattern of amplitude modulation. However, continuous component of orthogonal pulse based modulation schemes is not constant over entire signal. It varies with the order of orthogonal pulses. Since pulses are orthogonal, their continuous and discrete components can be reduced for orthogonal pulse based

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modulation schemes. For analysis of PSD of orthogonal pulse based modulation schemes, PSM, BPSM and OPPM-BPSM modulation schemes have been considered [36,121,140].

Nonorthogonal pulse waveforms are important design consideration for TH-UWB systems to coexist with victim NB system [40,41]. Different types of pulses are used for TH-UWB systems for different purposes [37,38]. For example, the Gaussian pulse family consisting of Gaussian pulse, Gaussian monopulse, Gaussian doublet, doublet defined by Hamalianen and dualcycle can be used for TH-UWB systems [148]. However, these pulse waveforms do not satisfy the FCC spectral mask without the carrier signals. Several pulses satisfy the FCC spectral requirement but they are incapable of reducing UWB interference on the victim systems. It is in this context, the present work proposes a generalized pulse waveform of Gaussian pulse family based on null spectrum for TH-UWB systems [114,149]. This pulse waveform is a combination of higher order derivative of Gaussian pulses. A mathematical analysis of this combined pulse waveform shows that other pulse waveforms such as simple Gaussian pulse, Gaussian monopulse, higher order derivatives of Gaussian pulse, and Gaussian doublet can be obtained from this pulse waveform by changing its parameters. This means that this pulse can be used as a generalized form of all Gaussian pulses. It reduces the UWB interference power on any NB systems by shifting its null spectral to the center frequency of these NB systems. Moreover, it satisfies the FCC spectral mask without using the carrier frequency [114].

5.3 PSD of Orthogonal Pulse Based Modulation Schemes

To improve the system performance, several orthogonal pulse based modulation schemes have been proposed for TH-UWB systems [36,121,140]. The advantages of these schemes in terms of interference reduction, system capacity and low design complexity have been presented in [150–152]. However, spectral efficiency of these pulses and corresponding

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PSD of the TH-UWB signal have not been analyzed so far. The PSD provides relevant information about the signal such as coexistence capability of UWB systems and to design NB systems for avoiding UWB interference. It is in this context, the PSD of orthogonal pulse based modulation schemes is analyzed.

To evaluate the PSD of orthogonal pulse based modulation schemes, the generalized M -ary TH-UWB signal for PSM, BPSM and OPPM-BPSM schemes can be written as

$$s(t) = \sum_l^{M-1} \sum_h^{N_s-1} a_l w_l(t - lN_p T_f + hT_f - c_{l,h} T_c - \delta_l) \quad (5.1)$$

where a_l , $w_l(t)$ and δ_l are the independent and stationary processes, N_p is the period of TH sequence, T_f is the time frame, $w_l(t)$ is the l^{th} order orthogonal pulse waveform transmitted during l^{th} symbol duration and $c_{l,h}$ is designed based on the TH-sequence. The continuous form of Eq. (5.1) can be rewritten as

$$y(t) = \sum_l s_{p_l}(t - lN_p T_f). \quad (5.2)$$

The corresponding PSD of orthogonal pulse based modulation schemes from Eq. (2.42) can be expressed as

$$P_y(f) = \frac{1}{N_p T_f} \left[E \{ |S_p(f)|^2 \} - E \{ S_p(f) S_q^*(f) \} \right] + \frac{1}{(N_p T_f)^2} \sum_k E \{ S_p(f) S_q^*(f) \} \delta \left(f - \frac{k}{N_p T_f} \right) \quad (5.3)$$

where p and q are two independent random variables with the same probability distribution function as p_l . $S_p(f)$ is the FT of $s_p(t)$ which can be evaluated as

$$S_p(f) = \sum_{l=0}^{M-1} a_l e^{-j2\pi f \delta_l} W_l(f) T_l(f) \quad (5.4)$$

where $W_l(f)$ is the FT of transmitted pulse $w_l(t)$. $T_l(f)$ is the FT of the TH code presented in Eq. (2.44). The expectation of $|S_p(f)|^2$ given in Eq. (5.4) can be evaluated as

$$E \{ |S_p(f)|^2 \} = E \left\{ \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} a_l a_n e^{-j2\pi f (\delta_l - \delta_n)} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \right\}. \quad (5.5)$$

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where a_l and a_n are the independent random variables, and δ_l and δ_n are also independent random variables but from the different process. The Eq. (5.5) can be rewritten as

$$E\{|S_p(f)|^2\} = \sum_{l=0}^{M-1} \left\{ E\{a_l a_l\} |W_l(f)|^2 |T_l(f)|^2 + \sum_{\substack{n=0 \\ n \neq l}}^{M-1} E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \right\}. \quad (5.6)$$

Similarly the second expectation in Eq. (5.3) can be expressed as

$$\begin{aligned} E\{S_p(f) S_q^*(f)\} &= \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \\ &= E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\} \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \end{aligned} \quad (5.7)$$

The term $\sum_{l=0}^{M-1} \sum_{n=0}^{M-1} W_l(f) W_n^*(f) T_l(f) T_n^*(f)$ is for two cases: one is for $l = n$ and other is for $l \neq n$. Therefore, Eq. (5.7) can be simplified as

$$\begin{aligned} E\{S_p(f) S_q^*(f)\} &= E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\} \\ &\quad \times \sum_{l=0}^{M-1} \left\{ |W_l(f)|^2 |T_l(f)|^2 + \sum_{\substack{n=0 \\ n \neq l}}^{M-1} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \right\} \end{aligned} \quad (5.8)$$

Using Eqs. (5.6) and (5.8) in Eq. (5.3), the PSD of orthogonal pulse based modulation schemes can be expressed as

$$\begin{aligned} P_y(f) &= \frac{E\{a_l a_l\} - E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\}}{N_p T_f} \sum_{l=0}^{M-1} |W_l(f)|^2 |T_l(f)|^2 + \\ &\quad \frac{E\{a_l a_n\} E\{e^{-j2\pi f(\delta_l - \delta_n)}\}}{(N_p T_f)^2} \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \sum_k \delta\left(f - \frac{k}{N_p T_f}\right) \end{aligned} \quad (5.9)$$

For PSM scheme, symbols are modulated only by order of pulses, i.e. $a_l=1$ and $\delta_l = 0$. The expectations of these variables are $E\{a_l a_l\} = 1$, $E\{a_l a_n\} = 0$ and $E\{e^{-j2\pi f(\delta_l - \delta_n)}\} =$

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1 respectively. The corresponding PSD of PSM signal can be inherited from Eq. (5.9) as

$$P_y(f) = \frac{1}{N_p T_f} \sum_{l=0}^{M-1} |W_l(f)|^2 |T_l(f)|^2 + \frac{1}{(N_p T_f)^2} \sum_{l=0}^{M-1} \sum_{n=0}^{M-1} W_l(f) W_n^*(f) T_l(f) T_n^*(f) \times \sum_k \delta\left(f - \frac{k}{N_p T_f}\right) \quad (5.10)$$

It has been observed that the first part of PSD is continuous spectrum that depends on the TH-code and the ESD of the l^{th} order orthogonal pulse. Since ESDs of different order orthogonal pulses are not same, selection of order of orthogonal pulses plays an important role for continuous PSD component. The second part is discrete PSD components which contain the term $W_l(f) W_n^*(f)$. The pulses are orthogonal in time domain as well as in frequency domain. Therefore, $W_l(f) W_n^*(f)$ reduces amplitude of discrete components.

For BPSM scheme, symbols are modulated by order of pulses and amplitudes, i.e. $a_l \in \{\pm 1\}$ and $\delta_l = 0$. The expectation of these variables are $E\{a_l a_l\} = 1$, $E\{a_l a_n\} = 0$ and $E\{e^{-j2\pi f(\delta_l - \delta_n)}\} = 1$. The corresponding PSD of BPSM scheme is

$$P_y(f) = \frac{1}{N_p T_f} \sum_{l=0}^{M-1} |W_l(f)|^2 |T_l(f)|^2 \quad (5.11)$$

The continuous PSD component of BPSM signal is similar to the PSM scheme without any discrete components. The PSD of the TH-UWB signal is smoothed by this scheme which allows the signal to coexist with other NB signals.

For OPPM-BPSM modulation scheme, $a_l \in \{\pm 1\}$ and $\delta_l = (l-1)\delta$, where δ is the constant time shift length. These implies, $E\{a_l a_l\} = 1$, $E\{a_l a_n\} = 0$ and $E\{e^{-j2\pi f m T_\Delta \delta}\} = (1 + \cos(2\pi m f T_\Delta))/2$. The corresponding PSD of OPPM-BPSM signal can be expressed as

$$P_y(f) = \frac{1}{N_p T_f} \sum_{l=0}^{M-1} |W_l(f)|^2 |T_l(f)|^2 \quad (5.12)$$

The PSD of BPSM and OPPM-BPSM schemes are the same. However, OPPM-BPSM can be used for higher level modulation scheme for higher data rate systems. Therefore, OPPM-BPSM modulation is an attractive choice of TH-UWB signal from several aspects.

5.4 Generalized Pulse Waveform of Gaussian Pulse Family

In this section a generalized pulse waveform is introduced to reduce UWB interference on NB system and to satisfy the FCC spectral mask without using any carrier signal. Time and frequency domain representation, ESD, notch and reduction of in-band interference power at NB system for this pulse waveform are presented.

5.4.1 Time Domain Representation

The generalized pulse waveform is represented by a combination of two higher order derivatives of Gaussian pulses. A typical pulse waveform $p(t)$ with two components of Gaussian derivatives can be expressed as

$$p(t) = ay^{(n-1)}(t) + by^{(n-1)}(t - t_w) \quad (5.13)$$

where $a, b \in [-1, 1]$, n is the positive integer, superscript indicates the order of derivative and t_w represents the time gap between two pulses. The Gaussian pulse $y(t)$ and its n^{th} order derivatives are given in Eq. (2.1) and Eq. (2.2).

Assuming constant aperture transmitter antenna with ideal characteristics, transmitted pulse waveform of Eq. (5.13) can be represented as

$$p^{(1)}(t) = ay^{(n)}(t) + by^{(n)}(t - t_w) \quad (5.14)$$

where $y^{(n)}$ is the n^{th} order derivative of Eq. (2.1).

From Eq. (5.13), it is possible to obtain expressions for conventional Gaussian pulse family by changing the parameters of a , b and n . Table 5.1 shows the various pulse waveforms with different values of parameters. It indicates that proposed pulse waveforms behave like a generalized pulse waveform for all Gaussian pulses. In addition, this combined pulse waveform satisfies the FCC power requirements without any carrier signal

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and is also able to reduce UWB interference at the NB receivers by creating spectral null points in a specific frequency. UWB interference in NB receiver depends on the center frequency of that NB system and the PSD of UWB radio signals. Lower spectral spike can be produced at the center frequency of NB systems by creating null point of ESD of UWB signal at the center frequency of the NB system. It reduces UWB interference at NB systems [40]. Hamalianen doublet also creates the null spectral, however, it does not satisfy the FCC spectral mask without using any carrier signal. However, higher order Gaussian doublet satisfies the FCC spectral mask irrespective of modulation techniques and pulse width [116].

5.4.2 Frequency Domain Representation

The frequency domain characteristics of the proposed pulse waveforms are analyzed extensively to find the suitability of using those in the available bandwidth. The FT of Eq. (5.14) can be expressed as

$$\begin{aligned}
 P(f) &= \mathcal{F}(p^{(1)}(t)) \\
 &= aY_n(f) + bY_n(f) \exp(-j2\pi ft_w) \\
 &= A(j2\pi f)^n \exp\left(-\frac{(2\pi f\sigma)^2}{2}\right) \left(a + b \exp(-j2\pi ft_w)\right) \\
 &= A(j2\pi f)^n \exp\left(-\frac{(2\pi f\sigma)^2}{2}\right) \exp(-j\pi ft_w) \left(a \exp(j\pi ft_w) + b \exp(-j\pi ft_w)\right).
 \end{aligned} \tag{5.15}$$

The corresponding amplitude spectrum, $|P(f)|$, and the ESD, $|P(f)|^2$, can be evaluated as

$$|P(f)| = A(2\pi f)^n \exp\left(-\frac{(2\pi f\sigma)^2}{2}\right) \sqrt{(a+b)^2 \cos^2(\pi ft_w) + (a-b)^2 \sin^2(\pi ft_w)} \tag{5.16}$$

$$|P(f)|^2 = A^2(2\pi f)^{2n} \exp\left(-(2\pi f\sigma)^2\right) \left((a+b)^2 \cos^2(\pi ft_w) + (a-b)^2 \sin^2(\pi ft_w)\right). \tag{5.17}$$

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Table 5.1: Parameters for different pulse waveforms

Parameters			Types of pulse waveforms
n	a	b	
1	1	0	Gaussian pulse
2	1	0	Gaussian monopulse
3	1	0	Gaussian doublet
3	1	1	Dualcycle
1	1	-1	Doublet
> 3	1	0	Higher order Gaussian pulse
≥ 3	± 1	± 1	Higher order Doublet

The frequency at which the maximum value of Eq. (5.16) is referred to as peak emission frequency, f_M , obtained by differentiating Eq. (5.16). The maximum value of the amplitude spectrum can be expressed as

$$|P(f_M)| = A(2\pi f_M)^n \exp\left(-\frac{(2\pi f_M \sigma)^2}{2}\right) \times \sqrt{(a+b)^2 \cos^2(\pi f_M t_w) + (a-b)^2 \sin^2(\pi f_M t_w)}. \quad (5.18)$$

The normalized PSD is

$$|P_n(f)| \cong \frac{|P(f)|^2}{|P(f_M)|^2} = \left(\frac{f}{f_M}\right)^{2n} \exp\left(-\left(2\pi\sigma(f-f_M)\right)^2\right) \psi(f). \quad (5.19)$$

The transmitted power, $P_{tu}(f)$, for UWB devices is defined as

$$|P_{tu}(f)| \cong A_{max} \frac{|P(f)|^2}{|P(f_M)|^2} = A_{max} \left(\frac{f}{f_M}\right)^{2n} \exp\left(-\left(2\pi\sigma(f-f_M)\right)^2\right) \psi(f) \quad (5.20)$$

where A_{max} is the peak PSD that the FCC permit and $\psi(f)$ can be expressed as

$$\psi(f) = \frac{(a+b)^2 \cos^2(\pi f t_w) + (a-b)^2 \sin^2(\pi f t_w)}{(a+b)^2 \cos^2(\pi f_M t_w) + (a-b)^2 \sin^2(\pi f_M t_w)}. \quad (5.21)$$

In order to satisfy the spectral mask, the parameters n and σ need to be evaluated.

This is done by rewriting the Eq. (5.19) in the following form

$$20n \log_{10}(f/f_M) - \frac{10(2\pi\sigma(f-f_M))^2}{\ln 10} + 10 \log_{10}(\psi(f)) - 10 \log_{10}|P_n(f)| = 0. \quad (5.22)$$

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The backoff value $10 \log_{10} |P_n(f)|$ is decided from FCC mask, for example when $f = 10.6$ GHz, $10 \log_{10} |P_n(f)|$ is -10 dB for indoor system. Eq. (5.22) has two values of $f\sigma$ with respect to a fixed value of n and $10 \log_{10} |P_n(f)|$. Again for fixed values of f , σ can be determined by using numerical method on Eq. (5.22). The value of n is determined based on the FCC spectral mask. It has been shown in Fig. 5.1 that the peak power position of ESD is increased with the increase in n , and the proposed pulses satisfy the FCC spectral mask for $n \geq 5$.

For unit power and unit values of a and b , Eq. (5.20) gives nulls (notch) in ESD at the points $f = (k/t_w)$ for $k = 1, 2, \dots, M$, as shown in Fig. 5.1. These points can be shifted by changing the value of t_w and one null point can be placed at the center frequency of 802.11a systems for $t_w = 0.189$ ns. Values of a and b are chosen by trial and error to get optimal notch. For unit values of a and b , these null spectral points are created below -70 dB power level and in normalized form below -35 dB power level (deep nulls). However, the fractional values of a and b also create the spectral notch at power level greater than -35 dB, which is approximately the same as normalized noise power level. Due to the presence of noise, fractional values of a and b can also reduce the same amount of UWB interference power at NB systems without creating deep nulls. But comparing with unit values, it gives relatively smaller spectral spikes in the PSD which gives lower UWB interference [116]. It indicates that the proposed pulse waveforms can be used adaptively in the UWB systems from different perspective by changing the values of a and b .

5.5 SIR Calculation on IEEE802.11a Receiver

Performance of a typical NB device, based on IEEE802.11a standard, in the presence of a single UWB device with the proposed pulse waveform is analyzed in this context. The model of 802.11a and UWB system is shown in Fig. 5.2. The received power of

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Figure 5.1: Normalized ESD of 4^{th} and 5^{th} order derivatives of the combined pulse waveforms for different values of a , b , and n

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802.11a receiver is obtained by using Friis free space equation [153] and can be expressed as follows

$$P_{R,802.11a}(d_o) = 10 \log_{10} \left(\frac{p_{to} G_{to} G_{ro} c^2}{(4\pi d_{oo} f_c)^2} \left(\frac{d_{oo}}{d_o} \right)^{n_o} \right) \quad (5.23)$$

where $p_{to} = 15$ dBm/MHz is transmitted and received power within the reference distance of 802.11a system. Separation distance of 802.11a receiver and transmitter is d_o ($= 5 - 55$ m), d_{oo} ($= 3$ m) is the known as reference distance, G_{to} is the transmitter antenna gain, G_{ro} is the receiver antenna gain and f_c ($= 5.3$ GHz) is the carrier frequency of the 802.11a system. The path loss factor of 802.11a system is $n_o = 0$ for $d_{oo} \leq 3$ m and $n_o = 3.3$ for $d_{oo} > 3$ m, assuming standard operation for 802.11a device. Using combined pulse waveforms in Eq. (2.60), UWB interference power (in dB) at 802.11a receiver within the band $f_{NL} - f_{NH}$ can be defined as

$$\begin{aligned} \tilde{I}_{UWB}(f_{NL}, f_{NH}) &= 10 \log_{10} \left(\frac{G_{tu} G_{ro} c^2}{(4\pi d_{uu})^2} \left(\frac{d_{uu}}{d_{uo}} \right)^{n_u} \int_{f_{NL}}^{f_{NH}} \frac{|p_{tu}(f)|}{f^2} df \right) \\ &= 10 \log_{10} \left(A_{max} \frac{G_{tu} G_{ro} c^2}{(4\pi d_{uu})^2} \left(\frac{d_{uu}}{d_{uo}} \right)^{n_u} \int_{f_{NL}}^{f_{NH}} \frac{f^{2n-2}}{f_M^{2n}} e^{(-2\pi\sigma(f-f_M))} \psi(f) df \right) \end{aligned} \quad (5.24)$$

where A_{max} is peak PSD defined in Eq. (5.20), $|p_{tu}(f)|$ is the transmitted power of UWB device, G_{tu} is the UWB transmitter antenna gain, d_{uu} is the UWB transmitter reference distance, d_{uo} is the distance between UWB transmitter and 802.11a receiver, n_u is the path loss factor of UWB system, and f_{NL} and f_{NH} are the lower and upper band-edges of the 802.11a system. The function $\psi(f)$ is defined in Eq. (5.21). Assuming all other system-linked parameters in Eq. (5.24) are constant and defined as B , then the Eq. (5.24) analytically can be written as

$$\tilde{I}_{UWB}(f_{NL}, f_{NH}) = 10 \log_{10} (B f_1(a, b, n, t_w)) \quad (5.25)$$

where

$$B = A_{max} \frac{G_{tu} G_{ro} c^2}{(4\pi d_{uu})^2} \left(\frac{d_{uu}}{d_{uo}} \right)^{n_u}$$

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Figure 5.2: Interference model of 802.11a and UWB systems

and

$$f_1(a, b, n, t_w) = \int_{f_{NL}}^{f_{NH}} \frac{f^{2n-2}}{f_M^{2n}} e^{(-2\pi\sigma(f-f_M))} \psi(f) df.$$

The SIR of the 802.11a receiver can be calculated using Eq. (2.61) and expressed as

$$SIR_{802.11a} = P_{R,802.11a}(d_0) - \tilde{I}_{UWB}(f_{NL}, f_{NH}) \quad (5.26)$$

where $P_{R,802.11a}(d_0)$ is the received power of 802.11a receiver which is transmitted by 802.11a transmitter from a distance of d_0 . The performance of 802.11a receiver increases with the increase of $SIR_{802.11a}$. Since the interference reduction is based on UWB signal characterization, it is required to reduce the $\tilde{I}_{UWB}(f_{NL}, f_{NH})$. As shown in Eq. (5.25), $\tilde{I}_{UWB}(f_{NL}, f_{NH})$ is a function of a , b , n and t_w and these parameters can be decided for the optimal interference reduction. The details of these parameters are discussed with Eq. (5.13). By appropriately selecting these parameters for UWB systems one can improve the performance of coexisting 802.11a devices.

5.6 Simulation Results

The performance of 802.11a down-link receiver is analyzed in the presence of a TH-UWB device with the generalized pulse waveform through comprehensive computer simulation studies. The 802.11a device assumes standard system parameters and studies are conducted in a Rayleigh fading channel for 802.11a device and a multipath fading model (CM1) proposed by IEEE802.15.3a study group for UWB systems [31, 154–156]. The channel model for UWB systems is a modified version of Saleh-Valenzuela model which follows Lognormal distribution. The simulation environment for 802.11a system assumes a constant signal to noise ratio of 10 dB.

As seen in the analysis, the proposed pulse waveform has four parameters, a, b, n, t_w , for spectral planning on 802.11a receiver and to satisfy the spectral mask. Computer simulation studies show that this generalized pulse waveform can be effectively used to reduce the UWB interference on 802.11a down-link receiver. Fig. 5.3 shows the performance of the 802.11a device for the proposed pulse waveform for different values of a, b , and n . Peak power point, f_M (shown in Fig. 5.1), is shifted from the center frequency of 802.11a systems to higher frequency with the increase in n [37]. Therefore, performance of 802.11a receiver is improved with the increase of n . From Fig. 5.3, it is evident that the system performance of 802.11a is not affected by the presence of interfering TH-UWB device for higher values of n . The generalized pulse waveform for $a = 1$ and $b = -1$ gives better performance than that of Gaussian derivatives, $a = 1$ and $b = 0$, for the same values of n . Fig. 5.3 shows that the data rate of UWB system is reduced with the increase of n because of lower bandwidth of the pulse waveform.

The relationship between data rates and ranges of transmission for a UWB communication system is shown in Fig. 5.4 and Fig. 5.5. The simulation assumes G_{tu} and G_{ru} equal to 0 dBi and the receiver has a constant BER of 10^{-6} . The bandwidth is set equal to the bandwidth of the pulse waveform to achieve the maximum SNR [37]. For example,

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Figure 5.3: Performance of 802.11a receiver at different distances (802.11a transmitter and 802.11a receiver) for different values of a , b and n (f_M is center frequency of UWB signal)

Fig. 5.4 shows that the TH-UWB device can have data rate of 100 Mb/s for $n = 1$ and it can operate up to 18 m with a BER performance of 10^{-6} . The data rate is reduced to 40 Mb/s with the same TH-UWB device with same operating distance when n is changed to 3. In general, the data rate reduces with the increase in n . For $b \neq 0$, and $a \neq 0$, the combined pulse waveform satisfies FCC requirements when $n \geq 4$ with fixed value of σ . For all other cases the proposed waveform satisfies FCC requirements when $n \geq 5$ with appropriate value of σ . From Fig. 5.3, Fig. 5.4 and Fig. 5.5, it can be seen that for $a = 1$ and $b = -1$ the system performance of 802.11a for $n = 3$ is slightly better than that of $n = 2$ and $n = 1$. However, for the same values of a and b , the data rates of UWB systems for $n = 1$ is 125 Mbps at distance 15 m, whereas at the same distance the data rate is 30 Mbps for $n = 3$. It implies that UWB interference is reduced gradually with the increase in n , however, system data rate is reduced constantly.

The above discussions lead to the necessity for defining a trade off between interference

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Figure 5.4: Data rate for UWB systems for binary PAM at different distances (UWB transmitter and UWB receiver) for different values of a , b and n (sig is time scaling parameter)

Figure 5.5: Data rate for UWB systems for binary PPM at different distances (UWB transmitter and UWB receiver) for different values of a , b and n ($sig = \sigma$ is time scaling parameter)

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and data rate for this generalized pulse waveform. Value of n can be decided depending on system requirements. Since peak point of ESD moves toward the higher frequency with increase in n , a larger value of n results in lower UWB interference power. However, pulse width increase with increase in n . Therefore, a smaller value of n should be chosen for high data rate, but it should not be less than 5 due to FCC spectral mask requirements. For unit values of a and b , these pulse waveforms are more efficient for reducing the UWB interference at NB systems. In addition, higher order derivatives of Gaussian pulse do not require any carrier signal to step the signal up from the baseband to a transmission bandwidth within the FCC mask. This technique is efficient, since it reduces the required expensive radio frequency (RF) components in the transceiver.

5.7 Summary

In this chapter, PSD analysis and TH-UWB interference on IEEE802.11a system have been investigated. The PSD analysis of PSM, BPSM and OPPM-BPSM modulation schemes have been presented. It has been shown that higher level orthogonal pulse based modulation schemes have higher power efficiency compared to conventional modulation scheme. A generalized combined pulse waveform based on higher-order derivative of the Gaussian pulses has been presented. This combined pulse waveform satisfies the FCC power spectral mask and is able to reduce UWB interference on 802.11a receiver by creating spectral null. Finally, simulation results on the effect of UWB interference on 802.11a systems have been discussed.

Chapter 6

Conclusions and Future Work

6.1 Introduction

This chapter presents a brief summary of the work that was accomplished in this dissertation. Some topics for future work to continue research in this area are also discussed. This chapter is organized as follows: The conclusion of OOK-PSM, OPPM-BPSM and TH-UWB interference with IEEE802.11a is presented in the section 6.2. The possible topics for future works are suggested in section 6.3.

6.2 Conclusions

In this dissertation modulation schemes based on orthogonal pulses have been investigated to reduce the TH-UWB system complexity with high data rates. A combined OOK-PSM scheme has been proposed for N -bit TH-UWB systems. The analytical formulation of MAI, MPI and IPI have been presented in the presence of multiple user and UWB indoor multipath environments. It is observed that MAI can be reduced considerably by assigning different subset of orthogonal pulses for different users. The MPI and IPI are minimum in synchronized systems by assuming perfect channel knowledge. Since orthogonal pulses are more susceptible in the presence of multipath channel, the system performance has been analyzed in the presence of multipath channel for different set of orthogonal pulses. Considerable system performance with higher data rates

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can be achieved by using fewer pulses and receiver correlators than those used in PSM and BPSM schemes. Due to presence of OOK, OOK-PSM is easy to implement in a low cost system design. Since OOK-PSM scheme is used for higher level modulation schemes, the synchronization problem for long sequence of zeroes is reduced compared to the conventional binary OOK scheme. It requires minimum energy and is applicable for energy constrained TH-UWB systems. Although performance of the OOK-PSM is lower than the BPSK-PSM scheme, it plays important role for system design when power and system complexity are constraints. The proposed scheme is downward compatible and hence is useful for adaptive modulation systems to suit different channel conditions and data rates.

It is known that higher level modulation schemes are required for higher data rate systems. However, OOK-PSM modulation scheme gives lower performance for higher level modulation schemes in the presence of multipath. To address this problem, an M -ary OPPM-BPSM modulation scheme has been proposed for higher level modulation scheme and multivariate data rate systems. In order to construct M -ary OPPM-BPSM signal, $L (= 2^l)$ pulse positions and $N (= 2^{k-l-1})$ biorthogonal pulses are required. The number of pulse positions and pulses are decided based on the system performance and the availability of orthogonal pulses with estimable autocorrelation properties. M -ary OPPM-BPSM modulation scheme allows an increase in the number of bits per symbol and consequently, reduces the duration of pulse repetition interval by increasing the number of orthogonal pulses. It achieves higher data rate by introducing more orthogonal pulses in the same pulse repetition interval. The system and computational complexity of this scheme is reduced to half that of M -ary PPM and M -ary PSM schemes by introducing antipodal version of orthogonal pulses. The system model and performance have been analyzed by analytical explanation and numerical results. The channel capacity and PSD of this scheme have also been evaluated to show the efficiency of this modulation scheme in term of system capacity and coexistence capability with other NB systems.

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To show coexistence capability of the nonorthogonal and orthogonal pulses based modulation schemes with other NB receiver, PSD of these modulation schemes have been analyzed. First, PSD of orthogonal pulse based modulation schemes such as PSM, BPSM and OPPM-BPSM have been studied to show the power efficiency of orthogonal pulses. It is shown that PSD of these modulation schemes not only depends on the time dithering but also depends on ESD of orthogonal pulses. Since ESDs of the different orthogonal pulses are different, the selection of order of orthogonal pulses is a crucial role for spectral design. Second, a generalized pulse waveform of Gaussian pulses has been proposed. This generalized pulse waveform is a combination of higher derivatives of Gaussian pulse. It satisfies the FCC spectral requirement without using carrier signal. The corresponding PSD and interference issues with NB system based on PPM and PAM schemes have been presented. UWB interference issues of the proposed pulse waveforms for IEEE802.11a systems are examined in this dissertation. It has been shown that the generalized pulse waveform reduces UWB interference over the IEEE802.11a systems by creating null at the center frequency of the desired NB systems. Hence, generalized pulse waveform enables coexistence of TH-UWB systems with any NB systems by shifting the null to the center frequency of the NB system.

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6.3 Recommended Future Work

In this dissertation two combined modulation schemes based on orthogonal pulses have been proposed for TH-UWB systems. However, these modulation schemes can be implemented for TH-UWB-MIMO, TR-UWB and DS-UWB systems as well. The higher order derivatives of Gaussian pulses and orthogonal pulses can be used to reduce UWB interference on NB systems. The pulse code design for MIMO system is another research area of orthogonal pulse. Details of these future work are described in the following.

6.3.1 OPPM-BPSM Scheme for IR-UWB-MIMO Systems

It is possible to improve transmission rates of the proposed M -ary OPPM-BPSM modulation scheme by employing advanced signal processing algorithms. It has been mentioned that M -ary OPPM-BPSM scheme uses higher level modulation to increase the data rate of the systems. Data rates can be improved further by employing MIMO algorithms in TH-UWB systems. Most of the present work on TH-UWB-MIMO systems is based on binary or 4-ary modulation schemes. By using the proposed M -ary OPPM-BPSM together with TH-UWB-MIMO system, one can obtain high data rates with higher value of M . Since this modulation scheme supports multi mode data rate, it is applicable to adaptive modulation systems by changing its number of positions and orthogonal pulses. However, designing adaptive modulation scheme for MIMO-UWB system is a challenging task in terms of complexity and system design. It also requires complete knowledge on MIMO-UWB channels.

6.3.2 TR-UWB for UWB-MIMO System Based on Orthogonal Pulses

Transmitted reference UWB (TR-UWB) communication system is used for simplicity and robust performance in multipath channel. It is self synchronized and eliminates the

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need for individual pulse synchronization with locally generated templates as in PPM. Despite all of its advantages, data rate of this system is limited due the requirement of double pulse transmission for every data bit. Data rates and performance of IR-UWB system can be improved by using orthogonal pulse based modulation schemes. However, TR-UWB systems perform poorly with a low SNR and in the presence of strong NB interference. Accurate timing of acquisition and tracking of integration window for each received bit is another important factor in the performance of TR-UWB receiver.

On the other hand, UWB channel estimation for IR-UWB-MIMO is a challenging task due to multipath propagation. The synchronization problem for IR-UWB-MIMO is also an important design concern. TR-UWB system with MIMO algorithms can address these issues more effectively. This can relax the channel estimation part in the receiver and utilize the multi-path channel impulse response to decrease the receiver complexity, through a pre-filtering at the transmitter. In addition, it is capable of effectively mitigating ISI and coexisting with the overlapping NB system without significant interference.

6.3.3 Orthogonal Pulses Based Modulation Schemes for DS-UWB Systems

In this dissertation M -ary OOK-PSM and M -ary OPPM-BPSM modulation schemes have been proposed for TH-UWB system. However, this scheme can be implemented for DS-UWB system as well. Due to limitation on the number of spreading codes, increasing the number of users is a major problem for DS-UWB systems. This is more significant for high data rate transmission systems where the number of orthogonal spreading codes is limited due to lower processing gain. It is in this context, one can increase the number of users by introducing several orthogonal pulses for the same spreading code. For instance, if the number of orthogonal pulses for DS-UWB systems is N , N users can use the same spreading code with different order of pulses. The desired data in the receiver can be

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distinguished by using spreading code and then by using orthogonal pulses. This process allows $N \times M$ users by using M spreading codes.

6.3.4 Designing Novel Set of Orthogonal Pulses

The system performance of orthogonal pulse based modulation schemes depends on the correlation properties of orthogonal pulses. To reduce MAI for larger number of users, a large set of orthogonal pulses is required. However, designing a large set of orthogonal pulses with good correlation properties is a challenging task. The MHPs and PSWFs do not give good correlation properties for higher order orthogonal pulses. Therefore, a novel set of orthogonal pulses with good correlation properties is essential for better system performance. Several orthogonal polynomials such as Chebyshev polynomial of the first kind and second kind, Gegenbauer polynomial, Jacobi polynomial, Laguerre polynomial and Legendre polynomial are available in the literature which can be used for developing orthogonal pulses. Like generalized pulse waveform, MHPs are also able to create nulls (notches) in the frequency domain. Therefore, MHPs can be used to reduce UWB interference on NB systems for higher level modulation schemes. However, hardware implementation of orthogonal pulses with sub nanosecond duration is really a challenging task.

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