ULTRA-WIDE BAND IN COMMUNICATIONS: PERFORMANCE ANALYSIS AND ENHANCEMENTS

by

BURAK BERKSOY
B.S. Gonzaga University, 2001

A dissertation submitted in partial fulfillment of the requirements for the degree of Doctor of Philosophy in the School of Electrical Engineering and Computer Science in the College of Engineering and Computer Science at the University of Central Florida Orlando, Florida

Spring Term
2008

Major Professor:
Lei Wei
Abstract

Over the last ten years, Ultra-Wide Band (UWB) technology has attracted tremendous research attention. Frequency allocation of 3.1-10.6 GHz for UWB application by FCC made it apparent that UWB will be the technology for future wireless high speed communication applications.

With the promise of high data rates (high channel capacity), UWB also offers advantages such as communication security, high multi-path resolution, good penetration capability, ability to coexist with other communication schemes in the same band, and finally, circuit simplicity. The theoretical advantages of UWB has made it a great candidate for short distance communications, however, UWB communications have many challenges, for example, sub-nanosecond pulse generation, timing sensitivity of modulation and synchronization, flat antenna performance over a wide bandwidth, effect of existing systems on UWB systems.

In order to experiment with various UWB modulation schemes, and to study transmitter and receiver structures, an accurate channel model need be established. In this dissertation, our first contribution is to evaluate and implement two major statistical channel models. The first model is proposed by AT&T Labs and is in the form of an autoregressive IIR filter. Although this is an accurate channel model to represent UWB behavior, it is proposed before the allocation of 3.1-10.6 GHz frequency band, hence, it could not simulate the correct
frequency spectrum. The second model is proposed by Saleh and Valenzuela, which has been widely accepted by UWB community to be the most accurate channel model for UWB systems. Recently disbanded task group 802.15.3a which was assigned to standardize a UWB communication scheme has also accepted the latter model.

Our second contribution is to derive optimal pulses for PPM signals. Using the accurate channel model in computer simulations, we experimented on various UWB communication schemes. We found that the traditional UWB pulses being used in pulse position modulated UWB systems did not perform optimally. A set of optimized UWB pulses and the methodology to calculate optimal pulses for any modulation index for PPM systems have been proposed in this dissertation. It is found that the optimal pulse can improve the performance of UWB systems by as much as 0.7 dB. With the PPM pulse optimization, the theoretical performance limits of PPM systems are derived.

The third contribution from this dissertation is to design near optimal practical implementable receiver structures. Some of the results obtained from PPM pulse optimization are found to be theoretical and not practical. More practical approach to the receiver structures were needed for industrial interest. We proposed simple sub-optimal receiver structures that are able to perform only a few dB less than the optimal receivers are proposed. These simple, low-cost receiver structures are strong alternatives to the complex traditional optimal receivers.
To my parents, Metin and Nilgn Berksoy for their endless love and support
ACKNOWLEDGMENTS

I would like to express my gratitude for my dissertation advisor and supervisor Dr. Lei Wei for his inspiration, motivation, and counseling. I would like to further expand my gratitude to all the members of my dissertation committee: Dr. Lindwood Jones, Dr. Mainak Chatterjee, Dr. Damla Turgut, Dr. Taskin Kocak, and Dr. Mark Heinrich for their valuable input and help.

I would like to extend my thanks my fellow graduate research assistant Yanxia, Libo, Nilesh, Seth, and Bharath for being great support in the lab and being wonderful people out of the lab. This dissertation has been a result of cooperation and teamwork which would be impossible without the hard working students, faculty and staff of University of Central Florida.

Finally, I would never be able to even start writing this dissertation without the love and support of my family. My mom, whose altruism extends beyond the imaginable. My dad, who worked harder then everyone to provide support for my education and my well-being. My beautiful sister, who has been an angel guiding me through life. For so many reasons that cannot be expressed in words, I dedicate this dissertation to my family.
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>LIST OF FIGURES</td>
<td>xi</td>
</tr>
<tr>
<td>LIST OF TABLES</td>
<td>xviii</td>
</tr>
<tr>
<td>LIST OF ACRONYMS/ABBREVIATIONS</td>
<td>xx</td>
</tr>
<tr>
<td>CHAPTER 1 INTRODUCTION</td>
<td>1</td>
</tr>
<tr>
<td>1.1 History of UWB</td>
<td>2</td>
</tr>
<tr>
<td>1.2 Comparison and Competition in Wireless Personal Area Networks</td>
<td>5</td>
</tr>
<tr>
<td>1.3 Advantages of UWB in Communications</td>
<td>8</td>
</tr>
<tr>
<td>1.4 UWB Challenges</td>
<td>10</td>
</tr>
<tr>
<td>1.5 FCC Regulations for UWB Communications</td>
<td>11</td>
</tr>
<tr>
<td>1.6 Outline of Dissertation</td>
<td>13</td>
</tr>
<tr>
<td>1.7 Contributions</td>
<td>15</td>
</tr>
<tr>
<td>1.8 List of Papers</td>
<td>16</td>
</tr>
<tr>
<td>CHAPTER 2 ULTRA WIDE BAND SIGNALS AND SYSTEMS</td>
<td>18</td>
</tr>
</tbody>
</table>
2.1 MB-UWB with Orthogonal Frequency Division Multiplexing (OFDM) . . . . 18
2.2 DS-UWB with Direct Sequence Spread Spectrum . . . . . . . . . . . . . . 19
2.3 Spread Spectrum Communications . . . . . . . . . . . . . . . . . . . . . . . . 20
2.3.1 Direct-Sequence Spread-Spectrum . . . . . . . . . . . . . . . . . . . . . . . . 21
2.4 UWB Communication Systems . . . . . . . . . . . . . . . . . . . . . . . . . . . . 23
2.4.1 UWB Communications with On-Off Keying . . . . . . . . . . . . . . . . . . . 24
2.4.2 UWB Communications with Pulse Position Modulation . . . . . . . . 29
2.5 UWB Waveforms . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 32
2.6 Multi-User Access in UWB using TDMA . . . . . . . . . . . . . . . . . . . . . . . 33
2.7 Error Performance of an UWB system in AWGN Channel . . . . . . . . 36
2.7.1 Error performance of On-Off Keying in AWGN Channel . . . . . . . . 36
2.7.2 Error performance of Pulse Position Modulation in AWGN Channel . 38

CHAPTER 3 CHANNEL MODELS FOR UWB SIMULATIONS . . . . . . . 41
3.1 Auto-Regressive UWB Channel Model . . . . . . . . . . . . . . . . . . . . . . . . 42
3.1.1 Large Scale Fading Model . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 43
3.1.2 Frequency Domain AR Model . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 46
3.1.3 Characterization of Parameters . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 48
3.1.4 Poles of AR Filter . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 52
3.2 S-V Channel Model ......................................................... 54
  3.2.1 Clustering of Multi-path components ......................... 55
  3.2.2 Channel Creation ..................................................... 56
3.3 Comparison and performance of Channel Models ............... 57

CHAPTER 4 PULSE WAVEFORM OPTIMIZATION FOR UWB SYSTEMS
USING PULSE POSITION MODULATION ................................. 67
  4.1 Pulse optimization for PPM on Gaussian channels ............. 69
  4.2 Pulse Waveform Optimization for Dense Multi-path Channels 73
  4.3 Aspects of Optimal Performance and Optimal Pulses ......... 78
  4.4 A Simple Way to Implement Optimal Pulses for Small Modulation Index 81

CHAPTER 5 SIMPLE ROBUST RECEIVER DESIGN AND PULSE OP-
TIMIZATION FOR UWB CHANNEL ............................... 87
  5.1 Introduction .............................................................. 87
  5.2 Simple Receivers ......................................................... 88
  5.3 Effect of Excess Noise ................................................... 89
  5.4 Pulse Candidates ........................................................ 90
  5.5 Performance Optimization in Additive White Gaussian noise (AWGN) Channels 91
    5.5.1 Receiver Parameter Optimization in AWGN Channels .... 92
5.5.2 Error Performance in AWGN Channels ................. 97

5.6 Performance in Additive Generalized Gaussian Noise (AGGN) Channels . . 98
  5.6.1 Receiver Parameter Optimization in AGGN Channels ................. 98
  5.6.2 Error Performance in AGGN Channels .......................... 102

5.7 Performance in Ultra Wide Band (UWB) Channels ................. 102
  5.7.1 Parameter Optimization for Simple Receivers in UWB Channels . . 104
  5.7.2 Error Performance in UWB Channels ......................... 107

CHAPTER 6 SUMMARY AND FUTURE WORK ......................... 110

  6.1 Summary ......................................................... 110
  6.2 Future Works ................................................... 111

LIST OF REFERENCES .................................................... 113
# LIST OF FIGURES

1.1 Wireless Network Classifications .................................................. 6

1.2 Example of the effect of multi-path on a 1ns pulse and 10ns pulse. Tx-Rx
distance is 1m. There are 4 multi-path components including the LOS initial
pulse with delays 0ns, 0.1001ns, 0.2001ns, 0.2669ns and phases $\pi, \frac{\pi}{0.60}, \frac{\pi}{1.5}, \frac{\pi}{0.89}$ 10

1.3 FCC’s designated emission mask for UWB indoor applications. .......... 12

2.1 BPSK direct-sequence spread-spectrum transmitter .......................... 21

2.2 BPSK direct-sequence spread-spectrum receiver .......................... 22

2.3 Spectrum of a non-spread message signal (a) vs. spread message (b) after
modulation with 50Hz sinusoid .................................................... 23

2.4 Message signal [1 1 0] is spread with code [1 0 1], modulated, demodulated,
and despread with the same code. a) message signal, b) spread message signal
c) modulated signal (transmitted) d)demodulated signal e) Despread message
signal ................................................................. 25
2.5 Message signal [1 1 0] is spread with code [1 0 1], modulated, demodulated, and despread with the code [0 1 1]. (a) message signal, (b) spread message signal, (c) modulated signal (transmitted), (d) demodulated signal, (e) despread message signal. Note that the received message has an invalid shape because of the spread and despread codes are different.

2.6 A Typical UWB Communication System Diagram using OOK

2.7 UWB-OOK Signals and processes. The final value of $r$ determines whether the decision is a "0" or "1".

2.8 A Typical UWB Communication System Diagram using PPM

2.9 Receiver structure for an UWB system using PPM

2.10 Simplified receiver structure for an UWB system using PPM

2.11 UWB-PPM Signals and processes. The final value of $r$ determines whether the decision is a "0" or "1".

2.12 "Traditional Pulse (a)" waveform for UWB systems

2.13 "Traditional Pulse (b)" waveform for UWB systems

2.14 Spectra of traditional pulses (a) & (b) presented in Figs. 2.12 & 2.13 with normalized energy.
2.15 UWB Signal with $N_c = 5$ (a) without time hopping, (b) With time hopping using sequence $[1 \ 3 \ 2 \ 0 \ 2]$ and data bit "0", (c) With time hopping using sequence $[1 \ 3 \ 2 \ 0 \ 2]$ and data bit "1". Notice in figure b pulses are shifted by $T_c e^{(j)}$ and in (c) they are further shifted by $T_n$ corresponding to a "1". 35

2.16 Occurrences and Probabilities of binary OOK signals with $E_b/N_0 = 0$ dB $E_b = 1$. 37

2.17 Occurrences and Probabilities of binary OOK signals with $E_b/N_0 = 0$ dB $E_b = 1$. 39

3.1 Path Loss Model 46

3.2 An example power delay profile generated with parameters given above 47

3.3 Autoregressive UWB Channel Model 48

3.4 Probability density function for magnitudes of filter coefficients $|a_2|$ and $|a_3|$ for LOS 50

3.5 Probability density function for magnitudes of filter coefficients $|a_2|$ and $|a_3|$ for NLOS 51

3.6 Probability density function for phases of filter coefficients $\angle a_2$ and $\angle a_3$ for LOS 52

3.7 Probability density function for phases of filter coefficients $\angle a_2$ and $\angle a_3$ for NLOS 53
3.8 An UWB signal is processed through 3 channels: 1) 6 m. LOS, 2) 10 m. NLOS, 3) 17 m. LOS

3.9 AR filter where one of the poles lie outside of the unit circle and the filter is unstable

3.10 AR filter where both poles lie within the unit circle and the filter is stable with flat response over the entire spectrum

3.11 Transmitted Pulse through UWB channel, pulse width is 3ns which makes its bandwidth near 1.25Ghz (AT&T Channel width)

3.12 Transmitted Pulse Frequency Response, pulse width is 3ns which makes its bandwidth near 1.25Ghz (AT&T Channel width)

3.13 Received Pulse through the AT&T UWB channel model with 1m LOS Tx-Rx separation

3.14 First poles of the 1000 generated channels

3.15 Second poles of the 1000 generated channels

3.16 Cluster Decay and Ray Decay for UWB Channel

3.17 Signal Passing through S-V channel a) Transmitted signal, b) Received signal 1m separation, c) Received signal 10m separation, d) Received signal 60m separation.

3.18 Path Loss vs Tx-Rx Separation for AT&T Model with LOS

3.19 Path Loss vs Tx-Rx Separation for AT&T Model with NLOS
3.20 Path Loss vs Tx-Rx Separation for S-V Model .......................... 66

4.1 Optimal pulse waveforms with $\Delta = \frac{1}{2}$ and $\Delta = \frac{1}{4}$ ............................ 78

4.2 Optimal pulse waveform with $\Delta = \frac{1}{3}$ with and without DC .......................... 78

4.3 Optimal pulse waveforms for multi-path channels with $\Delta = \frac{1}{3}$, (a) $\beta \in (0,1]$, $\tau = \frac{T_w}{3}, \theta \in (-\pi, \pi]$; (b) $\tau \in (0, T_w], \theta \in (-\pi, \pi], \beta = \frac{\sqrt{2}}{2}$ or $\beta \in (0,1]$ . . . 79

4.4 Optimal pulse waveforms for multi-path with $\Delta = \frac{1}{4}, \beta \in [0,1], \tau \in [0, T_w], \theta \in [-\pi, \pi], \Delta_x/\Delta = 1$ (solid), 1/3 (dashed), 1/10 (dotted) ............ 80

4.5 BER Performances of conventional pulse (a) and optimal waveforms given in Fig. 4.4 ..................................................... 81

4.6 Optimal BER degradation for UWB/PPM systems as a function of $N$, where $\frac{1}{N} < \Delta \leq \frac{1}{N+1}$ .................................................. 82

4.7 Comparison of $|X_i|$ and $g(t)$ for $\Delta = 1/20.$ .................................................. 83

4.8 Comparison of the optimal and the approximated waveforms for $\Delta = 1/20.$ . 83

4.9 Comparison of PSDs of the optimal and the approximated waveforms for $\Delta = 1/20.$ .................................................. 84

4.10 Comparison of $|X_i|$ and $g(t)$ for $\Delta = 1/10.$ .................................................. 84

4.11 Comparison of the optimal and the approximated waveforms for $\Delta = 1/10.$ . 85

4.12 Comparison of PSDs of the optimal and the approximated waveforms for $\Delta = 1/10.$ .................................................. 85
4.13 Comparison of the approximated and filtered waveforms for $\Delta = 1/20$.  

4.14 PSD for the optimized filtered UWB waveform overlapped with UWB emission 

mask $\Delta = 1/20$.  

5.1 Block Diagram of a Simple Receiver  

5.2 Four Pulses to be evaluated: a) Traditional UWB Pulse (Gaussian), 2) Half Cosine Pulse, 3) Rectangular Pulse, 4) Sinc Pulse. All Pulses have pulse duration of $T_p = 1$ns  

5.3 4 pulse candidates after the receiver filter with bandwidth $W = [\frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p}]$  

5.4 Probability of Error for 4 pulse candidates at 12dB SNR for $W = [\frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p}]$ with $\beta = 0.5$  

5.5 Probability of Error for 4 pulse candidates at 12dB SNR for $W = [\frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p}]$ with $\beta = 1$  

5.6 Probability of Error for 4 pulse candidates at 12dB SNR for $W = [\frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p}]$ with $\beta = 2$  

5.7 Probability of Error for 4 pulse candidates at 12dB SNR for $W = [\frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p}]$ with $\beta = 4$  

5.8 Probability of Error for Simple Receivers in AWGN channels with Half-Cosine pulse, $\beta = 1, W = \frac{1}{T_p}$
5.9 Bit Error Probability of Simple Receiver with $\beta = 1$ and Half-cosine Pulse in Additive Generalized Gaussian Noise Channels with $\gamma = [0.5, 1, 4]$. . . . . . . 103

5.10 Percentage of the Captured Energy of at the receiver output for different filter bandwidths ($W$) . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 105

5.11 Signal-To-Noise Ratio for pulse candidates at the receiver output for different filter bandwidths ($W$) . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 106

5.12 Probability of Bit Error for two pulse candidates with $W = 1/T_p$ vs receiver Power $\beta$ . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 107

5.13 Probability of Bit Error of Simple receivers using Rectangular Pulse and Half-Cosine Pulse with $W = 1/T_p$ and $\beta = 1$ compared to the optimal OOK performance . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 109
## LIST OF TABLES

3.1 Path Loss Model Parameters .................................................... 45

3.2 Percentage of power captured, number of multipaths, mean excess delay and RMS excess delay averages from AT&T Lab’s measurements .......................... 48

4.1 Optimized $X_i$s for $i = 1, ..., 20$, $\alpha$ and degradation factors ........................................... 77

5.1 Signal component the total signal ratio at the decision block input for $\beta = [0.5, 1, 2, 4]$ and receiver bandwidth $W = [50, 1, 2, 3, 4]$ GHz ........................................... 90

5.2 Energy Capture of Pulses for various filter bandwidths ............................... 92

5.3 Optimal Filter Bandwidth for 4 pulse candidates in AWGN Channel ............... 94

5.4 Probability of Bit Error for filter bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and Filter power $\beta = [0.5, 1, 2, 4]$ for 4 pulse candidates in AGGN Channel Channels with $\gamma = [0.5]$ ................................................................. 99

5.5 Probability of Bit Error for filter bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and Filter power $\beta = [0.5, 1, 2, 4]$ for 4 pulse candidates in AGGN Channel Channels with $\gamma = [1.0]$ ................................................................. 100
5.6 Probability of Bit Error for filter bandwidth \( W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right] \) and Filter power \( \beta = \left[ 0.5, 1, 2, 4 \right] \) for 4 pulse candidates in AGGN Channel Channels with 
\[ \gamma = \left[ 4.0 \right] \]
LIST OF ACRONYMS/ABBREVIATIONS

**AGGN**  Additive Generalized Gaussian Noise

**AR Model**  Auto-Regressive Model

**AWGN**  Additive White Gaussian Noise

**BER**  Bit Error Rate

**BPSK**  Binary Phase Shift Keying

**CDMA2000**  Code Division Multiple Access 2000

**DS-UWB**  Direct Sequence Ultra Wide Band

**FCC**  Federal Communications Commission

**GPRS**  General Packet Radio Service

**GSM**  Global System for Mobile communications

**HDTV**  High-Definition Television

**IIR**  Infinite Impulse Response

**LOS**  Line of Sight

**MAC**  Media Access Control

**MB-UWB**  Multi Band Ultra Wide Band
MPC  Multi Path Component

NLOS  Non Line of Sight

OFDM  Orthogonal Frequency Division Multiplexing

OOK  On-Off Keying

PPM  Pulse Position Modulation

PSD  Power spectral Density

RF  Radio Frequency

RMS  Root Mean Square

Rx  Receiver

S-V Model  Saleh-Valenzuela Model. UWB Channel model proposed by A. A. M. Saleh and R. A. Valenzuela in 1987

TDMA  Time Division Multiple Access

TG4a  Task Group 4 a. Task Group assigned to wireless personal area networks low rate alternative physical layer

Tx  Transmitter

UMTS  Universal Mobile Telecommunication Systems

USAF  United States Air Force
UWB  Ultra-Wide Bandwidth

UWB-PHY  Ultra-Wide Band Physical Layer

WLAN  Wireless Local Area Network

WPAN  Wireless Personal Area Network

WWAN  Wireless Wide Area Network
CHAPTER 1
INTRODUCTION

An Ultra Wide Band device is any device that the fractional bandwidth is greater than 0.25 or occupies 1.5 GHz or more of spectrum. The equation used for calculating the fractional bandwidth is \[ \frac{2(f_H - f_L)}{(f_H + f_L)} \], where \( f_L \) is the lower frequency of the -10 dB emission point and \( f_H \) is the higher frequency of the -10 dB emission point [15]. A typical UWB device transmits burst of pulses of nanosecond durations. Therefore the bandwidth of the transmitted pulse is in the order of several gigahertz.

Ultra wide band (UWB) communications have attracted much of attention recently, especially in the wireless indoor application. The UWB signals can offer high speed, low complexity, and reliable communication solutions [16] [17] [18]. Since the sub-nanosecond pulses effectively spread the energy of the UWB signal over a wide-range of spectrum (typically from zero to a few gigahertz), its link quality is near constant over the almost entire indoor space [34] due to the good penetrating capability of low frequency and diversity over the a few gigahertz bandwidth. This can effectively provide a reliable service to almost any place in the indoor environment, while the conventional approaches may not assure the reliable connection.
UWB communications are intended for short range communications due to the signals large scale fading at high frequencies. Short range communications are classified as wireless personal area networks (WPAN) denoted by FCC as 802.15. Wireless UWB, high speed bluetooth, wireless HDTV, wireless monitor, wireless photo printer, sensor networks are few of the possible applications suitable for UWB communications which has a theoretical data rates as high as 500 Mbits/s.

1.1 History of UWB

First appearance of UWB concept in the literature was in the late 1960’s. Sperry Rand corporation headed by Dr. Gerald F. Ross has developed the initial concepts of UWB, which was referred to, at the time, as carrier-free, baseband, or impulse. By 1970, Ross had published several papers on how to generate nanosecond duration pulses [1][2][3][4][5]. In the meantime, Harmuth published papers suggesting non-sinusoidal communication schemes [8],[9],[10]. Robbins’ US patent dated 1972 on short baseband pulse receiver [6] and Ross’ US patent dated 1973 on transmission and receiver system for short baseband systems [7] were the first landmarks for the UWB communications. Ross is considered one of the major pioneers of UWB communications, however, his patent filed in 1973 did not have any new concepts; it was a combination of concepts that were results of technology dating as early as 1940’s. What was new about Ross’ patent was the assumption that UWB communica-
tion systems could coexist with traditional sinusoidal communication schemes. Ross’ UWB communication system proposal consisted of five major components:

1. Methods to generate pulse trains

2. Methods to modulate pulse trains

3. Methods of switching (creating RF pulses)

4. Methods of detection and receiving

5. Appropriately efficient antennas

The subsystems described in Ross’ patent laid down the basics of UWB communications and remained unchanged for a number of years. Since the early research on UWB communications, the government gained interest in the technology and in 1977 USAF initiated an UWB research program headed by Col. J. D. Taylor. In fact the U.S. government classified UWB research to restrict it from public domain until 1994. Since 1994, the UWB research has been declassified and the research in the area gained momentum.

In the initial proposals UWB, communication systems were carrier-less. Carrier-less pulse transmission has interference issues. How an UWB device would interfere with other devices operating nearby was unclear mainly due to large variety of circuitry and materials used in the circuitry. It was unknown which materials and circuit components would react to short duration pulses.
In February 2002, FCC allocated 3.1-10.6 Ghz band for unlicensed use of UWB. With the allocation of frequency band, it was apparent that the technology would be commercially available in various devices; hence, attracted great deal of research attention. In December 2002, 802.15.3a task-force was issued. The purpose of the task-force was to standardize the physical layer of UWB devices to make them globally compatible. The task-force has narrowed 23 initial UWB-PHY proposals down to two major proposals. DS-UWB and MB-UWB. DS-UWB is known as direct sequence spread spectrum UWB. It is a variation of traditional approach to UWB communications. Today, DS-UWB proposal is backed by "UWB Forum", which is lead by Freescale Semiconductor and Motorola. MB-UWB is a newer approach to the use of FCC allocated UWB frequency. The spectrum is divided into 14 528Mhz bands. Orthogonal frequency division multiplexing is used for MB-UWB systems. Today, "WiMedia Alliance" is backing up the MB-UWB development and standardization, which is lead by Intel, Texas Instruments, Nokia, Hewlett-Packard, Sony, Microsoft, and many more.

802.15.3a task group has disbanded on January 2006, unable to come to a conclusion in standardizing UWB. Both fronts of UWB refused to back down from their proposals, 75% majority could not be reached, hence the task group came to a deadlock. Both proposals are expected to coexist as different standards in the near future.
1.2 Comparison and Competition in Wireless Personal Area Networks

Wireless personal area networks are short range networks typically in indoor environments which allow two or more devices to interconnect without wires at short ranges (typically up to 10m). Fig. 1.1 shows the range and classification of wireless networks. Typical technologies that operate at WPAN range are Bluetooth, UWB, ZigBee, and mmWave. IEEE 802.15 are the collection of standards that apply to wireless personal area networks. IEEE 802.15 has 5 task groups working on standardization of the WPAN technologies.

802.15.1 Bluetooth Task Group 802.15.1 task group along with Bluetooth SIG (Special Interest Group) is working on standardization of Bluetooth technologies. FCC approved Bluetooth 1.1 to be fully compatible with 802.15.1 in December 2002. There has been enhancements in the field with bluetooth 1.2, 2.0, 2.1. Bluetooth 1.1 could operate at 1Mbits/s data rate. Later versions have allowed data rates up to 3 Mbits/s. Bluetooth 3.0 is expected to adopt UWB technology for much higher data rates.

802.15.2 Coexistence Task Group The purpose of this task group is is to ensure that the WPAN can coexist with WLAN without interference.

802.15.3 High Rate Systems Task Group 802.15.3 is a set of standards that apply to MAC and PHY layers of devices that operate at high transmission rates (typically over 20 Mbits/s). 802.15.3 has 3 task groups: 802.15.3a, 802.15.3b, 802.15.3c. 802.15.3a (High Rate Alternative WPAN) was the task group that was assigned to standardize...
UWB technology in WPAN applications. The task group disbanded on January 2006. 802.15.3b is the task group working on the improvements and interoperability of MAC layer. 802.15.3c (Millimeter Wave Alternative WPAN) is the task group (also known as TG3c) formed May 2005 with purpose of standardizing devices that operate at 57-64 unlicensed frequency band.

802.14.4 Low Rate Systems Task Group 802.15.4 are the standards that apply to WPAN devices operating at low data rates, with long battery lives. the first edition of 802.15.4 standards was released in May 2003. ZigBee communication protocols are based on 802.15.4 standards. 802.15.4 has two assigned task groups. 802.14.4a is the low rate alternative PHY. the main interest is to provide reliable communications, high precision ranging location capability. The two candidates for the development of TG4a is pulsed UWB radio, and Chirp spread spectrum.

802.15.5 Mesh Networking 802.15.5 task group standardizes mesh networks, which consist of many nodes interconnected in Ad-Hoc. Mesh networks are not meant to be mobile.

Figure 1.1: Wireless Network Classifications
Wireless Local Area Networks (WLAN) are medium range networks that can extend up to 100m. 802.11 is the set of standards applying to wireless local area networks. Most popular application of WLAN is wireless interconnection of computers and hand-held devices. WLAN technologies evolved starting with 802.11a and 802.11b, then 802.11g. 802.11n is the technology which will be available in the near future. Wireless Wide Area Networks (WWAN) can extend to ranges up to 32 Km. Unlike WLAN, WWAN uses cellular networks. Most popular applications of WWAN is the data transmission through cell phones and hand-held devices. WiMax, UMTS, GPRS, CDMA2000, GSM are some communication technologies in WWAN.

For wireless personal area networks, until mmWave has been proposed, UWB offered the only high speed communication option. UWB technology has been a technology under investigation and has been made commercially available in semiconductor market. At the moment UWB offers the most reliable existing technology for wireless personal area networks. Due to these reasons, some of the older WPAN technologies such as bluetooth and 802.15.4 both see UWB as a candidate for their next generation products. mmWave is a higher frequency, higher data rate alternative to UWB. Although some semiconductor manufacturers have already made mmWave chips available, much like early years of UWB, mmWave has issues that needs to be dealt with.
1.3 Advantages of UWB in Communications

UWB communications has become a strong candidate for short range communications due to its high date rate capabilities. An UWB channel has a very short duration pulses, which results in high symbol rate. In exchange for the high data rate, communication bandwidth is greater and Signal to Noise ratio is reduced (due to FCC radiation limitations). The advantages of UWB in communications can be summarized as:

**High channel capacity** UWB communications have large bandwidth. FCC allocated 7.5 GHz bandwidth for UWB devices. However, the FCC emission mask suggests allows only -41.3dBm/MHz transmission power. According to Shannon’s channel capacity formula \( C = \log_2(1 + SNR) \), the channel capacity increases linearly with bandwidth, it decreases logarithmical with signal to noise ratio. the trade off between the bandwidth and signal to noise ratio allows UWB systems to have high data rates.

**Ability to coexist with traditional communication systems in the frequency spectrum**

The emission limit of -41.3dBm/MHz is considered below the noise floor for narrow-band communications; hence, does not interfere the operation of narrow-band schemes that reside in the same frequency spectrum.

**Communication security** The low average power of UWB transmitters and large bandwidth, the UWB signal is seen as noise to outside recipients. Furthermore, the UWB
signals are modulated with codes unique to designated receivers. Furthermore, the large bandwidth of UWB makes it hard for narrow-band to completely jam the signal.

**High multi-path resolution** Every radiated signal creates an echo of the original signal due to the reflections from surrounding surfaces, this is known as multi-path effect. These multi-path components can create out of phase addition that leads to signal cancelation. In narrow-band systems this can cause up to -40dB degradation in performance. Short duration pulses are more immune to the multi-path cancelation. Fig. 1.2 shows a case of multi-path component on a 1ns pulse transmission and a 10 ns pulse transmission. Example shows 4 received signal components. It is apparent that the 1ns pulses do not interfere since their arrival delay is larger than the pulse duration. In the case of 10ns pulse, the arrival time is smaller than the pulse duration, so the multi-path components interfere with each other. This interference can be constructive or destructive. In this example we assume there are no fading in the received signal. The interference for the 10ns pulse is out-of-phase. Although the transmitted power is 1W, the resulting power of the received signal is 0.3306W. In the case of 1ns pulse there are no interference, hence received power is 1W.

**Superior penetration capability** UWB pulses have very short duration, hence very long wavelengths. Long wavelength signals penetrate through a large variety of materials, including walls. However, the FCC emission limit does not permit the UWB systems to be applied to ranges longer than 10m.
Circuit simplicity The original proposals of UWB systems had carrier-less modulation schemes. Which eliminated the modulation demodulation timing and phase locking issues that followed modulation techniques did not apply to UWB system. However, nowadays, proposed UWB systems include high computational complexity in receivers, especially the MB-UWB which uses high complexity FFT operation.

![Figure 1.2: Example of the effect of multi-path on a 1ns pulse and 10ns pulse. Tx-Rx distance is 1m. There are 4 multi-path components including the LOS initial pulse with delays 0ns, 0.1001ns, 0.2001ns, 0.2669ns and phases $\pi$, $\frac{\pi}{6}$, $\frac{\pi}{5}$, $\frac{\pi}{89}$](image)

1.4 UWB Challenges

There are many challenges to implement and deploy UWB radio systems [17]. In short, they are (a) how to generate a sub-nanosecond pulse. The timing precision requirement is high, especially when the pulse position modulation is used. The timing synchronization
requirement is also high since the matched filter has a very narrow pulse. (b) How to design a ultra wide-band antenna to radiate and catch the energy. Typically, the antenna shall have a flat frequency response over a wide frequency band. (c) How does the UWB affect existing systems? The effect could include increasing the timing or carrier synchronization time or increasing the error rate.

1.5 FCC Regulations for UWB Communications

FCC has agreed on use of UWB communication purposes in compliance with the following regulations:

The equipment must be designed to ensure that operation can only occur indoors or it must consist of hand held devices that may be employed for such activities as peer-to-peer operation.

For indoor operation: Devices operating under this category must demonstrate that the system units will fail to operate if they are removed from the indoor environment. One acceptable procedure may be to show that the transmitting unit requires AC power to function. It is required that 10 dB bandwidth of indoor UWB systems must lie between 3.1 GHz and 10.6 GHz. In the frequency band below 960 MHz these devices are permitted to emit at or below the 15.209 limits, and emissions appearing above 960 MHz will conform to the following emissions mask:
An additional requirement for indoor UWB devices is that they may transmit only when operating with a receiver. A device connected to AC power is not constrained to reduce or conserve power by ceasing transmission, so this restriction will eliminate unnecessary emissions. In addition, if a device is designed to operate pointed downwards in an enclosed structure such as a metal or underground storage tank, it may operate at the levels allowed in this section. Fig. 1.3 shows the FCC’s designated emission mask for indoor UWB communications.

<table>
<thead>
<tr>
<th>Frequency in MHz</th>
<th>EIRP in dBm</th>
</tr>
</thead>
<tbody>
<tr>
<td>960-1610</td>
<td>-75.3</td>
</tr>
<tr>
<td>1610-1990</td>
<td>-53.3</td>
</tr>
<tr>
<td>1990-3100</td>
<td>-53.1</td>
</tr>
<tr>
<td>3100-10600</td>
<td>-41.3</td>
</tr>
<tr>
<td>above 10600</td>
<td>-51.3</td>
</tr>
</tbody>
</table>

Figure 1.3: FCC’s designated emission mask for UWB indoor applications.
Hand held UWB devices are required to operate with a 10 dB bandwidth between 3.1 GHz and 10.6 GHz. FCC is adopting an extremely conservative out of band emission mask to address the concerns of the great majority of commenters. In the frequency band below 960 MHz these devices are permitted to emit at or below the 15.209 limits, and emissions appearing above 960 MHz must conform to the following emissions mask:

<table>
<thead>
<tr>
<th>Frequency in MHz</th>
<th>EIRP in dBm</th>
</tr>
</thead>
<tbody>
<tr>
<td>960-1610</td>
<td>-75.3</td>
</tr>
<tr>
<td>1610-1900</td>
<td>-63.3</td>
</tr>
<tr>
<td>1900-3100</td>
<td>-61.3</td>
</tr>
<tr>
<td>3100-10600</td>
<td>-41.3</td>
</tr>
<tr>
<td>Above 10600</td>
<td>-61.3</td>
</tr>
</tbody>
</table>

### 1.6 Outline of Dissertation

In chapter 1, an introduction to ultra-wide band systems, their advantages and issues have been stated. Furthermore, the reasons behind the pursuit of solutions to the UWB problems have been explained.

The rest of this dissertation is organized as follows:

- In chapter 2, we provide a background of UWB signals and systems. Two main schools of UWB communications, MB-UWB and DS-UWB, are explained. We offer a detailed explanation to commonly used terms such as on-off keying, pulse position modulation,
spread spectrum, time division multiple access, orthogonal frequency division mult-
plexing is offered.

• In chapter 3, we evaluate and implement two major statistical channel models. The first model is proposed by AT&T Labs and is in the form of an autoregressive IIR filter. Although this is an accurate channel model to represent UWB behavior, it is proposed before the allocation of 3.1-10.6 GHz frequency band, hence, it could not simulate the correct frequency spectrum. The second model is proposed by Saleh and Valenzuela, which has been widely accepted by UWB community to be the most accurate channel model for UWB systems. Recently disbanded task group 802.15.3a which was assigned to standardize a UWB communication scheme has also accepted the latter model.

• In chapter 4, we investigate the pulse shapes of UWB systems using PPM and derive a methodology to compute the shape that minimizes bit error rate. Furthermore, we prove that pulses generated with such method yield to optimal pulses and PPM systems using suggested optimal pulses perform at their performance limits. We also quantify the improvement in bit-error rate compared with the UWB/PPM systems using traditional pulses.

• In chapter 5, we introduce simple, low-cost receiver structures. We optimize receiver parameters for minimum bit error rate in Additive white Gaussian channels, additive generalized Gaussian channels and UWB channels. We evaluate the performance of simple receivers for each channel and display the results.
• Finally, in 6, we summarize the work stated in this dissertation and discuss the future goals and experiments.

1.7 Contributions

The major contributions of this dissertation are itemized as follows:

1. Two major statistical UWB channel models; Autoregressive AT&T Channel model and Saleh-Valenzuela UWB channel model, are implemented to simulate UWB indoor environments. Both channel models are evaluated to accurately mimic the characteristics of an UWB channel. An accurate UWB channel is established for further investigation of UWB concepts in simulation.

2. Pulse shape that is used for PPM/UWB systems is optimized according to the modulation index of such system. A method for optimizing pulse shape for any modulation index is introduced. The optimality of the pulse created with above optimization method is proven.

3. By optimizing the pulse shape of PPM systems in order to minimize error performance, a performance limit of a PPM system is derived

4. Three major aspects of optimal pulses for PPM are listed and explained in detail.

5. It is shown that using a small modulation index, such as $\Delta = \frac{1}{10}$ or $\Delta = \frac{1}{20}$, we can approximate the optimal pulses with using gaussian pulse shape with specific values.
This simplifies the generation of pulses. It is also shown that with a simple process, we can modify the shape of the pulse to comply with FCC emission mask without a major performance degradation.

6. Simple, low-cost UWB receiver structures using OOK are introduced. The performance of these receivers are found to be closely related to the parameters of the receivers such as receiver filter bandwidth, receiver power and the pulse shape being used with receiver operation.

7. Parameters of simple receivers are optimized for AWGN channels, generalized gaussian channels, and UWB channels. Optimized receiver performances are evaluated for each channel.

8. Performances of simple receivers are shown to be significantly close to the optimal OOK system performance using matched filtering for every channel investigated. Simple receivers are shown to be promising low-cost, sub-optimal alternative to complex receivers.

### 1.8 List of Papers

Journal papers:
• B. Berksoy and L. Wei, "Pulse Waveform Optimization for UWB/PPM on Gaussian and Dense Multi-path Channels," IEEE Transactions on Vehicular Technology, on second submission.


Conference papers:


CHAPTER 2
ULTRA WIDE BAND SIGNALS AND SYSTEMS

Since the FCC’s allocation of unlicensed band for UWB applications, researchers and manufacturers are divided into two major schools of UWB communications: 1) MB-UWB (multi-band UWB using orthogonal frequency division multiplexing), which divides the allocated spectrum into 14 sub-bands and the information is processed through these sub-channels. 2) DS-UWB (Direct sequence UWB which uses direct sequence spread spectrum), which uses a traditional approach to UWB and the information is processed through a single band with a large bandwidth. The contributions in this document mainly apply to DS-UWB, however, it is useful to explain both schools of interest.

2.1 MB-UWB with Orthogonal Frequency Division Multiplexing (OFDM)

In MB-UWB systems the 3.1 - 10.6 GHz spectrum allocated by the FCC is divided into 14 528 MHz sub-channels. These sub-channels are grouped into 5 bands, which consist of 3 sub-channels except the fifth band, which consist of two sub-channels. Among the five bands, the first band is mandatory, the remaining bands are optional according to the communication mode. Each sub-channel is has 128 sub-carriers with spacing of 4.125 MHz in order to use
orthogonal frequency division multiplexing for multiuser access. Out of 128 sub-carriers, only 122 of them are used for data modulation and pilot signals. MD-UWB uses convolutional coding with $1/3$, $11/32$, $1/2$, $5/8$, $3/4$ for forward error correction. It also uses time frequency coding in order to interleave the coded data into up to three bands. MB-UWB supports data rates of $53.3$ Mbits/s, $80$ Mbits/s, $110$ Mbits/s, $160$ Mbits/s, $200$ Mbits/s, $320$ Mbits/s, $400$ Mbits/s, and $420$ Mbits/s. The support for use of the first band (first three sub-channels) and data rates $53.3$ Mbits/s, $110$ Mbits/s, and $200$ Mbits/s is mandatory. The UWB devices operating on the mandatory band and data rates are called Mode 1 devices. This document does not explain in detail the operation of MB-UWB since the research results stated in this document mostly relates to DS-UWB.

2.2 DS-UWB with Direct Sequence Spread Spectrum

Proposed DS-UWB with DSSS and Time Division Multiple Access (TDMA) supports data rates of $28$, $55$, $110$, $220$, $500$, $660$, $1000$ and $1320$ Mbits/s. It employs binary phase shift keying (BPSK) and quaternary bi-orthogonal keying (4BOK) for data modulation. Similar to MB-UWB, convolutional coding is used for forward error correction with $1/2$ and $3/4$ coding rates. Although it is often refereed to as single band UWB, DS-UWB supports operation in two separate frequency bands $3.1$-$4.85$ GHz (low-band), and $6.2$-$9.7$ GHz (high band). The later band is optional in DS-UWB operation. DS-UWB supports 1-24 chips long spreading code for multi-user operation.
2.3 Spread Spectrum Communications

Although many real world communication channels can be accurately modeled as stationary additive white gaussian noise (AWGN) channels, some of them do not fit in this category. Some channels operate with coexistence of narrow-band noise, some of them operate with existence of signal reflection which results multi-path effects in the communications. There exist a modulation and demodulation technique that can be used as an aid in mitigating the deleterious effects of the types of such interference: Spread Spectrum Communications.

In order for a system to be considered spread-spectrum, it must comply with the following characteristics:

1. The transmitted signal energy must occupy a bandwidth which is larger than the information bit rate (usually much larger) and which is approximately independent of the information bit rate.

2. Demodulation must be accomplished, in part, by correlation of the received signal with a replica of the signal used in the transmitter to spread the information signal.

An UWB system using time-hopping for multiuser access complies with two characteristics of spread-spectrum systems, it is considered to be a direct-sequence spread-spectrum system. Although, frequency hopping is an important spread spectrum technique, we will only explain direct-sequence spread spectrum method since it is the method used in time hopping.
2.3.1 Direct-Sequence Spread-Spectrum

Consider a BPSK signal \( s(t) \) defined by:

\[
s(t) = \sqrt{2P} \cos[\omega_0 t + \theta_d(t)]
\]  \hspace{1cm} (2.3.1)

A spreading sequence \( c(t) \) gets directly multiplied by the BPSK signal to be transmitted as spread spectrum signal:

\[
s_t(t) = \sqrt{2P}c(t)\cos[\omega_0 t + \theta_d(t)]
\]  \hspace{1cm} (2.3.2)

Figure 2.1 shows the direct sequence spread-spectrum transmitter block diagram. To de-

Figure 2.1: BPSK direct-sequence spread-spectrum transmitter

spread a spread-spectrum signal, we multiply the signal with the same spreading code \( c(t) \).

Figure 2.2 shows the BPSK direct-sequence spread-spectrum receiver.

\( c(t - T_d) \) in figure 2.2 is the same spreading code used in transmission shifted for synchro-

nization. A binary message sequence \([1\ 1\ 0]\) is spread using direct-sequence method with the spreading code \([1\ 0\ 1]\) and is modulated using BPSK. Figure 2.4 shows the signals as they

21
get spread, modulated, demodulated, and despread. Figure 2.5 shows the same signal when the despreading code is [0 1 1] (different than spreading code). The output message has an invalid shape is the spread and despread codes are different. Time-hopping multiuser access in UWB is derived from this property of spread spectrum, as each user is assigned a different spreading code, a message spread with one user’s spreading code is only accessible to that user as the codes assigned to each user are orthogonal from each other. In this example, the message signal is spread before being modulated as opposed to the block diagram in 2.1, where, the message signal gets modulated before its spread. The order of spreading and modulation does not matter. Similarly, the order of demodulation and despreading does not matter as well. Figure 2.3 shows the spectrum of modulated message signal with (a) and without (b) spread. It is clear that the spread message occupies more bandwidth, but the peak power is lower than the non-spread signal. Non-spread message is more susceptible to jamming at the modulation frequency.
2.4 UWB Communication Systems

As any other communication system UWB systems consists of a transmitter, a channel, and a receiver. What makes UWB special is that the signal radiated out from the transmitter complies with the requirements explained in section 1.5, that is, the signal has a fractional bandwidth greater than 0.25, or, it has a spectrum of 1.5 GHz or more. In this section, we will demonstrate the operation of two types of modulation schemes suitable for UWB communications: On-off keying and pulse position modulation. The short duration and high-bandwidth signal has different characteristics than a narrowband signal.
2.4.1 UWB Communications with On-Off Keying

Figure 2.6 represents the UWB communication system using OOK. Transmitted signal is either 0 or $w(t)$ corresponding to a binary "0" and "1" respectively. $h_c(t)$ represents the channel effects, $n(t)$ is the noise. The integrator output $r$ is called the decision criterion. $r$ is the integration of the received signal after going through the match filter for one bit duration.

Figure 2.7 shows the UWB signals as they are processed throughout the communication system assuming an AWGN channel. As demonstrated in the first row of figures, $w(t)$ or a 0 is transmitted according to the information bit "1" or "0" respectively. The UWB waveform $w(t)$ given in this example is explained in details in section 2.5 under the name "pulse a". In the second row of figures, noise is added to the transmitted signal. The signal goes through the match filter, therefore it is multiplied (to simplify convolution) by the signal $w(t)$ itself. Finally, in the forth row of figures, $r(t)$ is integrated to decide whether the received signal is "0" or "1". The integrator output $r$ is the decision criterion and if $r < T_h$ decision is "0", if $r > T_h$ decision is "1". $T_h$ is called the decision threshold. The error performance of a UWB OOK system is derived in section 2.7.1.
Figure 2.4: Message signal [1 1 0] is spread with code [1 0 1], modulated, demodulated, and despread with the same code. a) message signal, b) spread message signal c) modulated signal (transmitted) d) demodulated signal e) Despread message signal
Figure 2.5: Message signal $[1\ 1\ 0]$ is spread with code $[1\ 0\ 1]$, modulated, demodulated, and despread with the code $[0\ 1\ 1]$ . (a) message signal, (b) spread message signal, (c) modulated signal (transmitted), (d) demodulated signal, (e) Despread message signal. Note that the received message has an invalid shape because of the spread and despread codes are different.
Figure 2.6: A Typical UWB Communication System Diagram using OOK
Figure 2.7: UWB-OOK Signals and processes. The final value of $r$ determines whether the decision is a "0" or "1"
2.4.2 UWB Communications with Pulse Position Modulation

For a PPM system the transmitted signal $s(t)$ is either $w(t)$ or $w(t - \delta)$ according to the binary information bits "0" and "1" respectively. $w(t)$ is an UWB signal and $\delta$ is called the modulation index. The UWB signal corresponding to a "1" is simply the same signal corresponding to a "0", except, it is delayed by $\delta$.

The PPM receiver for AWGN channel consists of two branches. One of the branches is a match filter matched to the signal corresponding to "0", the second one is matched to
the signal corresponding to "1". The integration outputs $r_0$ and $r_1$ are compared and the
decision is made according to the greater value. In other words, if $r_0$ is greater than $r_1$ the
decision is a "0", and "1" otherwise. This receiver was then simplified to one branch using
the signal $w(t) - w(t - \delta)$ for the match filter impulse response. As it will be shown in the
section 2.7.2, this change did not only simplify the system, it allowed the system to perform
better than the system with two-branched receiver. Figure 2.10 shows the simplified receiver
structure.

![Figure 2.10: Simplified receiver structure for an UWB system using PPM](image)

Figure 2.10 shows the UWB signals corresponding to "0" and "1" for pulse position
modulation in row 1. Row 2 is the transmitted signal corrupted by noise. In row 3, the
corrupted signal gets multiplied by the simplified match receiver $w(t) - w(t - \delta)$. Finally,
row 4 is the integration operation where the final value of $r$ is at the end of the bit duration
(2 ns for this case). For simplified receiver, the sign of $r$ determines the decision. If $r > 0$,
decision is "0". If $r < 0$, decision is "1". The error performance of UWB-PPM systems are
explained in detail in section 2.7.2.
Figure 2.11: UWB-PPM Signals and processes. The final value of $r$ determines whether the decision is a "0" or "1"
2.5 UWB Waveforms

Two of the conventional UWB waveform equations are given and plotted in this section. These two waveforms have been preferred for their desirable spectral properties.

\[
w_a(t) = A \left[ 1 - 4\pi (2tf_c)^2 \right] e^{-2\pi (2tf_c)^2} \tag{2.5.1}
\]

\[
w_b(t) = 3\sqrt{e}Af_c\pi e^{-\left[\pi f_1.5f_c\right]^2} \tag{2.5.2}
\]

where A is the amplitude, \( f_c \) is the inverse of pulse duration \( T_w \). In Figs. 2.12 and 2.13, we have \( A = 1 \) V and \( f_c = 1 \) ns. For future reference the pulse given in 2.12 will be referred as "Traditional Pulse (a)" and 2.13 will be referred as "Traditional Pulse (b)".

![Figure 2.12: "Traditional Pulse (a)" waveform for UWB systems](image)

The spectrum of the two pulses are given in fig. 2.14. Both pulses have a flat power distribution for their given bandwidth. The bandwidth of the pulse is inversely proportional to the pulse duration \( T_w \).
In chapter 4 of this document, we will show that although these two pulses have superior spectral properties, they not the best candidates for communication applications using pulse position modulation (PPM).

2.6 Multi-User Access in UWB using TDMA

One way to use UWB in a multiuser environment is to apply spread spectrum technique using pseudo-random time-hopping codes with the information bits. Consider a pulse position modulated communication system using UWB pulse $w(t)$. One binary bit will be represented by $N_c$ number of subnanosecond pulses. $c^{(j)}$ is the time-hopping sequence for $j^{th}$ user which also corresponds of $N_c$ number of elements (binary numbers for some cases). In other words, $c^{(j)}$ is the $j^{th}$ user’s signature. The signal transmitted that is intended for $j^{th}$ user is $s^{(j)}(t)$ is given with the following formula:

$$s^{(j)}(t) = \sum_{l=0}^{N_c-1} w(t - lT_c - c^{(j)}_l T_d - d^{(j)}_k T_s) \quad (2.6.1)$$
Figure 2.14: Spectra of traditional pulses (a) & (b) presented in Figs. 2.12 & 2.13 with normalized energy.

where $T_c$ is the delay between pulses in the pulse train. $T_s$ is the modulation index. $T_d$ is the delay used for time hopping sequence. $d_k^{(j)}$ is the binary data intended for $j^{th}$ user [22].

In Figure 2.15, (a) shows the UWB pulse transmitted with $N_c = 5$ without spreading. 5 UWB pulses are separated by $T_c$ seconds (2 ns. in the example). In (b), the signal is spread using the time hopping code $[1 \ 3 \ 2 \ 0 \ 2]$. The time hopping code must be a matrix of size $1 \times N_c$. Time spread is applied by shifting each pulse with the $T_d$ time the spreading code corresponding to that pulse. For example, the first pulse gets shifted by $1T_d$, second pulse $3T_d$, and so on ($T_d = 0.4$ ns). If there is no further shift, this signal is a "0" intended for
user with code $[1\ 3\ 2\ 0\ 2]$. In (c) the signal in (b) is shifted by $T_s$ for a "1". This is a binary "1" intended for the same user as in (b).

Figure 2.15: UWB Signal with $N_c = 5$ (a) without time hopping, (b) With time hopping using sequence $[1\ 3\ 2\ 0\ 2]$ and data bit "0", (c) With time hopping using sequence $[1\ 3\ 2\ 0\ 2]$ and data bit "1". Notice in figure b pulses are shifted by $T_c c_1^{(j)}$ and in (c) they are further shifted by $T_s$ corresponding to a "1".

35
2.7 Error Performance of an UWB system in AWGN Channel

UWB performance of a system depends on the modulation scheme used. Since we are interested in the performance of on-off keying and pulse position modulation in this document, we will derive their performances in this section.

2.7.1 Error performance of On-Off Keying in AWGN Channel

As stated in section 2.4, for OOK, there is no pulse transmitted corresponding to a "0", and there is an UWB pulse $w(t)$ transmitted corresponding to a "1". The impulse response of the match filter receiver for this system will be $w(-t - T_w)$. Consider $r$ is the integrator output and the decision criterion. In the absence of noise, $r_0 = 0$ for a transmitted "0". Similarly, $r_1 = 2E_b$ corresponding to a transmitted "1". $E_b$ is the average bit energy and it is given with the following equation:

$$E_b = \frac{1}{2} \int_0^{T_w} w_0(t)^2 dt + \frac{1}{2} \int_0^{T_w} w_1(t)^2 dt$$  \hspace{1cm} (2.7.1)

Since $w_0(t)$ is zero for OOK,

$$E_b = \frac{1}{2} \int_0^{T_w} w_1(t)^2 dt$$  \hspace{1cm} (2.7.2)

In the presence of noise, $r_0 = N$ and $r_1 = 2E_b + N$, where $N$ is the gaussian noise with mean $\mu_n = 0$ and variance $\sigma_n^2 = N_0/2$. 10000 samples of $r_0$ and $r_1$ along with their histogram representing their pdf are displayed on fig.2.16. The optimum threshold that minimizes the
error performance is $E_b$, therefore the error performance can be expressed with:

$$P(e) = \frac{1}{2} \int_{-\infty}^{E_b} P(x|1)dx + \frac{1}{2} \int_{E_b}^\infty P(x|0)dx$$  \hspace{1cm} (2.7.3)

Figure 2.16: Occurrences and Probabilities of binary OOK signals with $E_b/N_0 = 0$ dB $E_b = 1$

Since the probabilities $p(x|0)$ and $p(x|1)$ are normally distributed with $\mu_0 = 0$ and $\mu_0 = 2E_b$ and variances $\sigma_n^2$ for both cases equation 2.7.3 becomes:

$$P(e) = \frac{1}{2} \int_{-\infty}^{E_b} \frac{1}{\sqrt{2\pi}\sigma_n} e^{-\frac{(x-2E_b)^2}{2\sigma_n^2}} dx + \frac{1}{2} \int_{E_b}^\infty \frac{1}{\sqrt{2\pi}\sigma_n} e^{-\frac{(x)^2}{2\sigma_n^2}} dx$$  \hspace{1cm} (2.7.4)

The integration yields equation 2.7.4 to:

$$P(e) = \frac{1}{2} Q\left(\sqrt{\frac{E_b}{N_0}}\right) + \frac{1}{2} Q\left(\sqrt{\frac{E_b}{N_0}}\right)$$  \hspace{1cm} (2.7.5)
Finally, the BER performance of an UWB system using OOK in AWGN channel is given by

\[ P(e) = Q \left( \sqrt{\frac{E_b}{N_0}} \right) \]  

(2.7.6)

### 2.7.2 Error performance of Pulse Position Modulation in AWGN Channel

For PPM, \( p_0(t) = w(t) \) and \( p_1(t) = w(t - \delta) \), where \( \delta \) is the modulation index. The impulse response of the match filter that minimizes the BER performance is:

\[ m(t) = w(t) - w(t - \delta) \]  

(2.7.7)

In absence of noise, the decision criterion \( r \) will be:

\[
\begin{align*}
    r_0 &= \int_0^T w(t)[w(t) - w(t - \delta)] = \int_0^T w(t)^2 + \int_0^T R(\delta) \\
    r_1 &= \int_0^T w(t - \delta)[w(t) - w(t - \delta)] = \int_0^T w(t)^2 - \int_0^T R(\delta)
\end{align*}
\]  

(2.7.8)

(2.7.9)

where, \( R(\tau) \) is the autocorrelation function of \( w(t) \), in other words:

\[ R(\tau) = \int_{-\infty}^{\infty} w(t)w(t - \tau)dt \]  

(2.7.10)

The average bit energy for a PPM system is:

\[ E_b = \int_0^T w(t)^2dt \]  

(2.7.11)

Therefore, in the presence of noise, we can express the decision criteria \( r_0 \) and \( r_1 \) as:

\[ r_0 = E_b - \Gamma + N \]  

(2.7.12)
\[ r_1 = -E_b + \Gamma + N \]  

(2.7.13)

Figure 2.17 shows the occurrence and probability distribution of the decision values \( r_0 \) and \( r_1 \) according to 10000 samples of data. Where \( N \) is the AGWN with variance \( N_0/2 \), and \( \Gamma \)

Figure 2.17: Occurrences and Probabilities of binary OOK signals with \( E_b/N_0 = 0 \) dB \( E_b = 1 \)

is expressed as:

\[
\Gamma = \frac{\int_0^T R(\delta) dt}{E_b}
\]  

(2.7.14)

The decision is ”0” if \( r < 0 \) and ”1” if \( r > 0 \). With the obvious threshold being 0, we can express the error probability of a PPM system with:

\[
P(e) = \frac{1}{2} P(r = 1|0) + \frac{1}{2} P(r = 0|1)
\]  

(2.7.15)
\[ P(e) = \frac{1}{2} \int_{-\infty}^{0} \frac{1}{\sqrt{2\pi}\sigma_n} e^{-\frac{(x-E_b+\Gamma)^2}{2\sigma_n^2}} \, dx + \frac{1}{2} \int_{0}^{\infty} \frac{1}{\sqrt{2\pi}\sigma_n} e^{-\frac{(x+E_b-\Gamma)^2}{2\sigma_n^2}} \, dx \]  
(2.7.16)

\[ P(e) = \frac{1}{2} Q\left(\sqrt{\frac{(1 - \Gamma)E_b}{N_0}}\right) + \frac{1}{2} Q\left(\sqrt{\frac{(1 - \Gamma)E_b}{N_0}}\right) \]  
(2.7.17)

Finally, the probability of error for a UWB system using PPM can be expressed as:

\[ P(e) = Q\left(\sqrt{\frac{(1 - \Gamma)E_b}{N_0}}\right) \]  
(2.7.18)
CHAPTER 3
CHANNEL MODELS FOR UWB SIMULATIONS

Wireless propagation of signals have been investigated for over 50 years. Numerous channel models have been proposed and are available in the literature. Signals that propagate through wireless channels have replicas of the original signal that arrive at the receiver at different time delays due to the travel path of the signal. This phenomenon is known as multipath propagation. It has been widely accepted that the flat Rayleigh-fading channel model is the most accurate channel model for the narrow-band communication schemes. The assumption is that the system bandwidth is small enough to ensure the multipath components of the transmitted signal interfere with each other and the complex envelope of the received signal has a complex gaussian distribution, which results in Rayleigh distribution of the envelope amplitude.

For systems that use larger bandwidths, such as UWB systems, only few multipath components (MPCs) overlap and central limit theorem does not apply to the amplitude of the received signal, hence the Rayleigh fading model is no longer applicable to such channels. For an accurate system analysis, the UWB channel has to be modeled according to the multipath arrival behavior.
There has been many channel model proposals since UWB systems have been established to be a strong candidate for short range, high speed communications. In the original assumption, the UWB was intended to use carrier-less modulation. Measurement campaigns have measured the channel response in such assumption. In February 2002, FCC has announced the authorization of use of UWB technology and allocated the 3.1-10.6 GHz frequency range for the communication applications. The ruling has restricted the application to indoor environments. There has been additional studies performed to measure indoor channel characteristics in the allocated frequency range, and many promising models have been published. 802.15.3a task group has adopted a modified model based on S-V model in [32] as the model for wireless short range UWB personal area network communications.

In this chapter, two channel models are explained in detail. the first model is the autoregressive model suggested in [36]. The second one is the S-V model suggested in [32], which is adopted by the 802.15.3a task group for the UWB short range communications. In chapter 4, results of the earlier experiments are discussed, hence, AT&T model is used for simulations. In chapter 5, experiments were done using S-V model.

### 3.1 Auto-Regressive UWB Channel Model

In order to simulate the UWB channel in software, there has been significant effort to create a model based on channel measurements [34], [36], [37], [38], [39], [41], [40]. The results of the channel sounding measurements done by AT&T Labs in the frequency range 4.375-5.625 GHz
concluded that an indoor UWB channel can be modeled as an autoregressive (AR) filter [36]. Due to the low complexity and high accuracy of the model, we generated a randomized UWB software channel to perform simulations according to their findings in [36]. FCC had allowed the use UWB in the frequencies of suggested model [14], however, recently these frequencies have been reassigned to 3.1 - 10.6 GHz [15]. Although the channel model presented in this chapter is no longer at the frequency of interest, it provides an insight on operation challenges of UWB and the problems associated with it.

### 3.1.1 Large Scale Fading Model

In this subsection, we deal with large scale fading of UWB channels. The parameters described in this subsection do not help us generate an UWB channel, however, it gives us an insight of UWB channel behavior which is important in designing transmitters and receivers for the UWB channel. There are three important parameters to characterize large scale fading in wireless channels. 1) Path Loss with respect to Tx-Rx separation $PL(d)$, 2) mean excess delay $\tau_m$ and 3) RMS excess delay $\tau_{RMS}$. The first parameter characterizes the expected power at the receiver at a certain distance from the transmitter, the later two parameters are important to characterize the multipath delay and UWB pulse "tail" for a particular UWB channel. Generally the path loss can be expressed as:

$$PL(d) = 10\log_{10} \left( \frac{P_{Tx}}{P_{Rx}} \right)$$

(3.1.1)
where \( P_{Tx} \) and \( P_{Rx} \) are the transmitted and received signal power respectively. \( d \) is the Tx-Rx separation.

According to AT&T channel measurements, the path loss for an UWB channel can be modeled by:

\[
PL(d) = PL_0 + \gamma_c 10 \log_{10}(d/d_0) + S
\]  \hspace{1cm} (3.1.2)

where \( d > d_0 = 1 \) m, \( PL_0 \) is the path loss at 1m Tx-Rx separation in dB, \( S(d) \) is the random shadow fading effect due to relative Tx-Rx positioning in dB, and \( \gamma_c \) is the channel path loss exponent, which changes from one location to another.

There are three parameters to model to characterize path loss, and all of these parameters differ from one location to another. Hence, these parameters are statistically modeled according to the measurements. The intercept \( PL_0 \) is a normally distributed random variable with mean and variances given in table 3.1.1. \( \gamma_c \) is the channel fading exponent widely accepted in the literature for all wireless channels. This parameter is also normally distributed with mean and variances given in table 3.1.1. The shadow fading parameter \( S \) is normally distributed with zero mean and normally random standard deviation with mean and variances given in table 3.1.1. The generated path loss samples and mean path loss curves are shown in figure 3.1.

The mean excess delay \( \tau_m \) and RMS excess delay \( \tau_{RMS} \) are important parameters to characterize the time dispersion of the wireless channel. They are given with:

\[
\tau_m = \sum_{i=1}^{L} \tau_i |h(t, \tau_i)|^2
\]  \hspace{1cm} (3.1.3)
<table>
<thead>
<tr>
<th>( P_{L0} )</th>
<th>( \gamma_c )</th>
<th>( \sigma_s )</th>
<th>LOS</th>
<th>Mean</th>
<th>Std. Dev.</th>
<th>NLOS</th>
<th>Mean</th>
<th>Std. Dev.</th>
</tr>
</thead>
<tbody>
<tr>
<td>47</td>
<td>1.7</td>
<td>1.6</td>
<td>47</td>
<td>N/A</td>
<td></td>
<td>51</td>
<td>N/A</td>
<td></td>
</tr>
<tr>
<td>N/A</td>
<td>0.3</td>
<td>0.5</td>
<td>N/A</td>
<td></td>
<td></td>
<td>3.5</td>
<td>0.97</td>
<td></td>
</tr>
<tr>
<td>N/A</td>
<td>3.5</td>
<td>2.7</td>
<td>N/A</td>
<td></td>
<td></td>
<td>0.98</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

\[ \tau_{RMS} = \sqrt{\sum_{i=0}^{L} (\tau_i - \tau_m)^2 |h(t, \tau_i)|^2} \] (3.1.4)

Mean excess delay is the first moment of the power delay profile, and RMS excess delay is the square root of the second moment of the power delay profile. As fig 3.2 shows an example power delay profile which was obtained with the following parameters:

<table>
<thead>
<tr>
<th>( a_2 )</th>
<th>( a_3 )</th>
<th>( w_0 )</th>
<th>( w_1 )</th>
<th>( \sigma_N )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5025</td>
<td>-0.4427</td>
<td>0.00368</td>
<td>-0.00218</td>
<td>1.36e-4</td>
</tr>
<tr>
<td>-0.8017</td>
<td>0.7538</td>
<td>-0.00003</td>
<td>0.00296</td>
<td>N/A</td>
</tr>
</tbody>
</table>

The explanations of these parameters are detailed in subsection 3.1.2

By applying 3.1.3 and 3.1.4, we obtain the mean excess delay and RMS excess delay for the given power delay profile in 3.2 to be: \( \tau_m = 5.3955ns \) and \( \tau_{RMS} = 4.1195ns \).

At&t labs have concluded the following results for the mean excess delay and RMS excess delay measurements in 23 homes and offices with given thresholds:
Another important finding is that $\tau_{RMS}$ increases as $d^{0.26}$ and $d^{0.36}$ for LOS and NLOS respectively.

### 3.1.2 Frequency Domain AR Model

The goal of frequency domain channel model is the develop a statistical representation of the channel with a minimum number of parameters to regenerate the measured channel behavior accurately in computer simulations. According to AT&T measurements, the channel frequency response $H(f, t; d)$ does not exhibit significant variability in time and can be
Figure 3.2: An example power delay profile generated with parameters given above assumed to be stationary. The following is the channel response model expression:

\[ a_i H(f_i, t; d) = n_i - a_2 H(f_i, t : d) - a_3 H(f_{i-2}, t; d) \]  \hspace{1cm} (3.1.5)

where \( n_i \) is a sequence of independent zero mean and identically distributed random variables. \( n_i \) is a sequence of independent Gaussian variables with zero mean and \( \sigma(d) \) standard deviation. The AR model can be implemented using the IIR model shown in Fig.3.3. The frequency response model has four complex parameters; \( a_2, a_3, w_0(d), w_1(d) \), and one real parameter; noise standard deviation \( \sigma(d) \). There are 9 real channel parameters in total.
Table 3.2: Percentage of power captured, number of multipaths, mean excess delay and RMS excess delay averages from AT&T Lab’s measurements

<table>
<thead>
<tr>
<th>Threshold</th>
<th>% Power</th>
<th>L</th>
<th>(\tau_m) (ns)</th>
<th>(\tau_{RMS}) (ns)</th>
<th>% Power</th>
<th>L</th>
<th>(\tau_m) (ns)</th>
<th>(\tau_{RMS}) (ns)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5-dB</td>
<td>46.8</td>
<td>7</td>
<td>1.95</td>
<td>1.52</td>
<td>46.9</td>
<td>8</td>
<td>2.2</td>
<td>1.65</td>
</tr>
<tr>
<td>10-dB</td>
<td>89.2</td>
<td>27</td>
<td>7.1</td>
<td>5.77</td>
<td>86.5</td>
<td>31</td>
<td>8.1</td>
<td>6.7</td>
</tr>
<tr>
<td>15-dB</td>
<td>97.3</td>
<td>39</td>
<td>8.6</td>
<td>7.48</td>
<td>96</td>
<td>48</td>
<td>10.3</td>
<td>9.3</td>
</tr>
<tr>
<td>20-dB</td>
<td>99.4</td>
<td>48</td>
<td>9.87</td>
<td>8.14</td>
<td>99.5</td>
<td>69</td>
<td>12.2</td>
<td>11</td>
</tr>
<tr>
<td>30-dB</td>
<td>99.97</td>
<td>60</td>
<td>10.83</td>
<td>8.43</td>
<td>99.96</td>
<td>82</td>
<td>12.4</td>
<td>11.5</td>
</tr>
</tbody>
</table>

Figure 3.3: Autoregressive UWB Channel Model

3.1.3 Characterization of Parameters

The standard deviation noise component of the AR model \(n_i\) is normally distributed with mean and standard deviations:

\[
E[20\log_{10}[\sigma(d)]]_{LOS} = -10.8\log_{10}(d) - 77.2
\]

\[
\sigma[20\log_{10}[\sigma(d)]]_{LOS} = 3.7
\]

\[
E[20\log_{10}[\sigma(d)]]_{NLOS} = -20.1\log_{10}(d) - 78.5
\]

\[
\sigma[20\log_{10}[\sigma(d)]]_{NLOS} = 4.5
\]
The magnitude of $a_1$ is 1. $|a_2|$ and $|a_3|$ are given with:

$$|a_2|_{LOS} = -0.923W + 0.3836N + 1.7869 \quad (3.1.10)$$

$$|a_2|_{NLOS} = -0.863W + 0.501N + 1.5369 \quad (3.1.11)$$

$$|a_3|_{LOS} = -0.3836W + 0.9235N + 0.9837 \quad (3.1.12)$$

$$|a_3|_{NLOS} = -0.506W + 0.836N + 0.7367 \quad (3.1.13)$$

where $W$ is a weibull distributed random variable with probability distribution function:

$$f(w) = \alpha \beta (W)^{\beta - 1} e^{\alpha W^\beta} \quad (3.1.14)$$

where $\alpha_{LOS} = 15.191$, $\beta_{LOS} = 1.312$, $\alpha_{NLOS} = 8.104$, $\beta_{NLOS} = 1.442$. $N$ is a zero-mean normally distributed random variable with standard deviation $\sigma$ ($\sigma_{LOS} = 0.0202$ and $\sigma_{NLOS} = 0.0234$). Probability density functions for the magnitudes of the filter coefficients are shown in figures 3.4 and 3.5 for LOS and NLOS respectively.

The phases of $a_2$ and $a_3$ are strongly correlated, if one of the phase of $a_2$ is generated, phase of $a_3$ can be calculated using:

$$\angle a_{3,LOS} = -6.4422 + 2.0505 \angle a_{2,LOS} \quad (3.1.15)$$

$$\angle a_{3,NLOS} = -6.5592 + 2.1022 \angle a_{2,NLOS} \quad (3.1.16)$$

where, $\angle a_{2,LOS}$ is Weibull distributed in range (1.95,2.97) radians with $\alpha = 0.3773$, $\beta = 16.5023$, $\angle a_{2,NLOS}$ is also weibull distributed in range (1.24,2.85) radians with $\alpha = 0.001$, $\beta = 7.8896$. Probability density functions for the phases of the filter coefficients are shown in figures 3.6 and 3.7 for LOS and NLOS respectively.
Probability density function for magnitudes of filter coefficients $|a_2|$ and $|a_3|$ for LOS

Initial conditions of the model are denoted as $w_0$ and $w_1$. Magnitudes of initial conditions $w_{0\text{LOS}}$ and $w_{0\text{NLOS}}$ have log-normal distributions with means and standard deviations:

$$E[20\log_{10}|w_0(d)|]_{\text{LOS}} = -16.3\log - 10(d) - 47$$ \hspace{1cm} (3.1.17)

$$STD[20\log_{10}|w_0(d)|]_{\text{LOS}} = 5.6$$ \hspace{1cm} (3.1.18)

$$E[20\log_{10}|w_0(d)|]_{\text{NLOS}} = -36.5\log - 10(d) - 45.9$$ \hspace{1cm} (3.1.19)

$$STD[20\log_{10}|w_0(d)|]_{\text{NLOS}} = 7.8$$ \hspace{1cm} (3.1.20)

Magnitudes of $w_{1\text{LOS}}$ and $w_{1\text{NLOS}}$ are calculated using $w_{0\text{LOS}}$ and $w_{0\text{NLOS}}$ according to following equations:

$$|w_{1\text{LOS}}| = 0.89|w_{0\text{LOS}}|$$ \hspace{1cm} (3.1.21)
Figure 3.5: Probability density function for magnitudes of filter coefficients $|a_2|$ and $|a_3|$ for NLOS

$$|w_{1NLOS}| = 0.93|w_{0NLOS}|$$  \hspace{1cm} (3.1.22)$$

Phases of the initial condition $w_0$ are denoted as $\phi_{0LOS}$ and $\phi_{0NLOS}$ for line of sight and non line of sight; they are uniformly distributed in the interval $(-\pi, \pi)$. Phases of the initial condition $w_1$ is strongly correlated to the ones of $w_1$. They are expressed with the following formula:

$$\phi_{1LOS} = 0.9888\phi_{0LOS} + 2.26 + v_{LOS}$$  \hspace{1cm} (3.1.23)$$

$$\phi_{1NLOS} = 0.9827\phi_{0NLOS} + 2.01 + v_{NLOS}$$  \hspace{1cm} (3.1.24)$$

The noise components in these equations are normally distributed with mean $\mu_{LOS} = 0.0142$ and standard deviation $\sigma_{LOS} = 0.294$, $\mu_{NLOS} = 0.0594$ and $\sigma_{NLOS} = 0.4277$. 

51
With the given complex \((a_2, a_3, w_0, w_1)\) and real \(n\) parameters, an UWB communication channel can be constructed using the equation 3.1.5. We are able to process an UWB signal through the channel based on the above model. Fig. 3.8 shows an UWB pulse (pulse a) received after being processed through three different channels: 1) 6 m. distance LOS, 2) 10 m. distance NLOS, and 3) 17 m. distance LOS.

### 3.1.4 Poles of AR Filter

The stability of the auto-regressive filter must be ensured so that the filter has a flat response over the full UWB frequency spectrum. To ensure stability, we have to investigate the poles
We can express the polynomial for the filter shown in fig. 3.1 as

\[ a_1 X^2 + a_2 X + a_3 = 0 \]  \hspace{1cm} (3.1.25)

then we have two complex roots that satisfy 3.1.25 \( s_1 \) and \( s_2 \). The roots are given with Bhaskaracharya’s widely known quadratic roots formula:

\[ S_{1,2} = \frac{-a_2 \pm \sqrt{(a_2^2 - 4a_1a_3)}}{2a_1} \]  \hspace{1cm} (3.1.26)
When one of the complex roots lies outside of the unit circle, in other words if \( |a_2| > 1 \) or \( |a_3| > 1 \), then the AR filter is no longer stable. The filter energy will be concentrated at the high frequencies, hence the received filter will have more power at high frequency components. An example of such filter is shown in fig. 3.9. When the poles of the channel lie within the unit circle, the filter is stable and has a flat response over the entire frequency spectrum. An example for stable channel is shown in fig. 3.10. With these channel response, we show and compare the transmitted UWB pulse vs. received pulse. In order for the filter to fit the UWB captured data, the magnitude of poles also have to lie within [0.90-1.00]. If the magnitudes of poles are less than 0.90, the filter is stable and the response is flat, however, it does not possess the properties discovered in statistical data.

Using the At&T channel model we generate 1000 channels and we plot the two poles of these channels with the unit circle in figures 3.14 and 3.15. As shown, first poles of the generated channel lie within the [0.90 1] distance from the unit circle. The second poles also lie within the [0.75 1] distance from the unit circle.

### 3.2 S-V Channel Model

S-V Channel model is based on the Saleh Valenzuela paper [32], and is adopted by the 802.15.3a task group for the UWB short range communications. This document includes studies done prior to February 2002, hence, the model used for these studies is the Auto-Regressive UWB channel model explained in section 3.1. For the recent studies, the S-V
model is used. The first main difference between the two models is that the parameters for AR model is applied to the channel impulse response in frequency domain $H_c(f)$. The S-V model statistically estimates the parameters which apply to the time-domain representation of the UWB channel impulse response. For the S-V model the channel impulse response is represented by:

$$h_c(t) = \sum_k \beta_k e^{j\theta_k} \delta(t - \tau_k)$$  \hspace{1cm} (3.2.1)

where $\delta(.)$ is the dirac (impulse) function. This model has been widely accepted for the general multi-path channels. The S-V model is used for statistically estimating the multi-path component amplitude $\beta_k$, phase $\theta_k$, and delay $\tau_k$.

### 3.2.1 Clustering of Multi-path components

Both in S-V model and experiments done by Turin [?] show that the multi-path components (rays) arrive in clusters. So the general multi-path channel model in 3.2.1 can be expanded to:

$$h_c(t) = \sum_{l=0} \sum_{k=0} \beta_{kl} e^{j\theta_{kl}} \delta(t - T_l - \tau_{kl})$$  \hspace{1cm} (3.2.2)

where $\beta_{kl}$, $\theta_{kl}$, and $\tau_{kl}$ are respectively the amplitude, phase, and delay of $k^{th}$ cluster and $l^{th}$ ray. $T_l$ is the $l^{th}$ cluster delay.

According to S-V model, $T_l$ and $\tau_{kl}$ are described by the independent interarrival exponential probability density functions:

$$p(T_l|T_{l-1}) = \Lambda e^{-\Lambda(T_l-T_{l-1})}$$  \hspace{1cm} (3.2.3)
\[ p(\tau_l|\tau_l - 1) = \lambda e^{-\lambda(\tau_{kl} - \tau_{(k-1)l})} \quad (3.2.4) \]

where \( \Lambda \) is the cluster arrival rate, and \( \lambda \) is the ray arrival rate.

\( \beta_{kl} \) are statistically independent positive random variables whose mean squares are monotonically decreasing functions of \( T_l \) and \( \tau_{kl} \). In S-V model,

\[ \overline{\beta^2_{kl}} = \overline{\beta^2_{00}} e^{-T_l/\Gamma} e^{-\tau_{kl}/\gamma} \quad (3.2.5) \]

where \( \Gamma \) is the cluster power delay time constant, \( \gamma \) is the ray power delay time constant, and \( \overline{\beta^2_{00}} \) is the average power gain of the first ray. Fig. 3.16 shows the envelope of mean square amplitude of multi-path components. In S-V model, the clusters often overlap.

### 3.2.2 Channel Creation

After the Clustering of the multi-path components is established, it is appropriate to introduce the creation procedure of the simulation channel. First, the average of the first ray \( \overline{\beta^2_{00}} \) has to be estimated. This quantity is directly related to the multi-path power gain for the room where channel is established \( G(r) \).

\[ G(r) = G_{1m} r^{-\alpha} \]

where \( \alpha = 3 \) to 4 in typical office buildings, but can go up to 6 in dense office buildings. \( G(1m) \) can be calculated using:

\[ G(1m) = G_t G_r [\lambda_0/4\pi]^2 \quad (3.2.6) \]

where \( G_t \) and \( G_r \) (\( \approx 1.6 \)) are transmitter and receiver antenna gains respectively. \( \lambda_0 \) (\( =0.2m \)) is the RF wavelength, for the particular 10ns pulse. The average power of the first ray can
be approximated with:

$$\beta_{00}^2 \approx (\gamma \lambda)^{-1} G(1m) r^{-\alpha}$$

(3.2.7)

Next, the ray arrival times $\tau_{kl}$ must be generated using 3.2.4. Then the amplitudes $\beta_{kl}$ are generated using Rayleigh probability density function:

$$p(\beta_{kl}) = (2\beta_{kl}/\beta_{kl}^2) \exp(-\beta_{kl}^2/\beta_{kl}^2)$$

(3.2.8)

where $\beta_{kl}^2$ can be obtained from 3.2.5. Finally, the phase $\theta_{kl}$ are generated with uniform distribution in the range $[0, 2\pi)$. Typical values to simulate office environment are: $1/\Lambda = 200 - 300\, \text{ns}$, $\gamma = 20\, \text{ns}$, $\Gamma = 60\, \text{ns}$, and $1/\lambda = 5\, \text{ns}$. Fig. 3.17 shows three received signals with receiver separation b) 1m, c) 10m, and d) 60m. The transmitted signal is shown in a). The energy of the transmitted signal is 1. The energy of the received signal in b) is $5.1518 \times 10^{-4}$, which is close to $G(1m)$ given in 3.2.6. Furthermore, the signal loss due to distance is apparent in all the received signals.

### 3.3 Comparison and performance of Channel Models

Both channel models presented in this chapter characterize the UWB channel well enough to give insight for design purposes. The main difference between the two channels is that the AT&T channel model is in frequency domain and S-V model is in time domain. It is possible to obtain frequency domain and time domain representation of each channel model via DFT and IDFT operations. Another difference between the two channel models is that AT&T channel takes into consideration the Tx-Rx line of sight, whereas the S-V channel
does not make such distinction and statistically models the channel for mix of LOS and NLOS positioning. One advantage to the S-V model is the fact that the measurement were taken in the FCC allocated frequencies. However, the channel displays similar properties at both frequencies (3.5GHz and 6.85GHz). In this section we are going to compare the path loss of both channel models.

We investigate the path loss for the generated channels using both models. Figures 3.18 and 3.19 correspond to path loss measurements for the generated channels using AT&T model for LOS and NLOS respectively. Fig. 3.20 displays the path loss using S-V channel using $\alpha = 3$. We can conclude that the S-V channel approximates the path loss model measured by AT&T Labs large scale fading model better than the AT&T channel model itself. First of all the NLOS and LOS path loss models for AT&T model does not show significant differences in Path Loss exponent which is represented by the slope of the least squares line fit. In S-V model this slope is accurately modeled.

Establishing a reliable model for the UWB channel allows us to experiment new ideas that would improve the performance of communication systems in UWB channels. There are numerous good models in the literature that simulate the UWB channel. For our experiments, the two models discussed have been used based on their simplicity and well accepted accuracy. In the next chapters, novel ideas for UWB communication enhancements are introduced, as these ideas are not tested on actual UWB channel, their accuracy relies on the accuracy of these two simulation channels.
Figure 3.8: An UWB signal is processed through 3 channels: 1) 6 m. LOS, 2) 10 m. NLOS, 3) 17 m. LOS
Figure 3.9: AR filter where one of the poles lie outside of the unit circle and the filter is unstable

Figure 3.10: AR filter where both poles lie within the unit circle and the filter is stable with flat response over the entire spectrum
Figure 3.11: Transmitted Pulse through UWB channel, pulse width is 3ns which makes its bandwidth near 1.25Ghz (AT&T Channel width)

Figure 3.12: Transmitted Pulse Frequency Response, pulse width is 3ns which makes its bandwidth near 1.25Ghz (AT&T Channel width)
Figure 3.13: Received Pulse through the AT&T UWB channel model with 1m LOS Tx-Rx separation

Figure 3.14: First poles of the 1000 generated channels
Figure 3.15: Second poles of the 1000 generated channels

Figure 3.16: Cluster Decay and Ray Decay for UWB Channel
Figure 3.17: Signal Passing through S-V channel a) Transmitted signal, b) Received signal 1m separation, c) Received signal 10m separation, d) Received signal 60m separation.
Figure 3.18: Path Loss vs Tx-Rx Separation for AT&T Model with LOS

Figure 3.19: Path Loss vs Tx-Rx Separation for AT&T Model with NLOS
Figure 3.20: Path Loss vs Tx-Rx Separation for S-V Model
CHAPTER 4
PULSE WAVEFORM OPTIMIZATION FOR UWB SYSTEMS USING PULSE POSITION MODULATION

In this chapter, we present several ways to optimize pulse waveforms for ultra wide band (UWB) systems using pulse position modulation (PPM) on Gaussian and dense multi-path channels, which minimize bit error rate (BER) for a given modulation index. We prove several propositions which lead to construct optimal pulses on Gaussian channels. We show that with the same modulation index the optimal pulses can improve the performance by 0.4-0.7 dB over the conventional pulses. On the dense multi-path channels, we present a reduced complexity optimization method. Similar performance improvements can also be obtained.

On Gaussian channels we have $p_m(t) = w(-t) - w(-t - \Delta)$ and $h_c(t) = \delta(t)$. In section 2.7.2, we have shown that probability of bit error in a pulse position modulated UWB communication scheme is:

$$P_b(e) = Q\left(\sqrt{\frac{(1 - \Gamma)E_b}{N_0}}\right)$$  \hspace{1cm} (4.0.1)

where

$$\Gamma = \frac{R(\Delta)}{R(0)}$$  \hspace{1cm} (4.0.2)
\[
Q(x) = \int_{x}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{y^2}{2}} dy \tag{4.0.3}
\]

The autocorrelation function \( R(\tau) \) is given with equation 2.7.10. Equation (4.0.3) clearly shows that we can design a pulse waveform \( w(t) \) to minimize the error rate.

For the multipath case, the channel response can be represented with;

\[
h_c(t) = \sum_{l=1}^{M} \beta_l w(t - \tau_l)e^{j\theta_l} \tag{4.0.4}
\]

where \( M \) is the total number of the multi-path, \( \beta_l \) denotes the magnitude of the \( l^{th} \) path, \( \theta_l \) denotes the phase and \( \tau_l \) denotes the delay. Typically, \( \beta_l, \theta_l, \) and \( \tau_l \) are random variables.

Then, we express the signal at the input of the receiver as

\[
s_i(t) = \sum_{k=-\infty}^{\infty} \sum_{l=1}^{M} \beta_l p_{uk}(t - kT - \tau_l)e^{j\theta_l} \tag{4.0.5}
\]

The matched filter (matched to the transmission pulse and channel) is then

\[
h_m(t) = \sum_{l=1}^{M} \beta_l[w(-t + \tau_l) - w(-t + \tau_l + \Delta)]e^{-j\theta_l} \tag{4.0.6}
\]

The bit error probability is

\[
P_b(e) = \int_{\beta,\tau,\theta} Q\left(\sqrt{\frac{\xi(\beta, \tau, \theta)E_b}{N_0}}\right)p(\beta, \tau, \theta)d\beta d\tau d\theta \tag{4.0.7}
\]

where,

\[
\xi(\beta, \tau, \theta) = \sum_{l=1}^{M} \sum_{q=1}^{M} \beta_l \beta_q e^{j(\theta_l - \theta_q)}[R(\tau_q - \tau_l) - R(\tau_q - \tau_l - \Delta)] \tag{4.0.8}
\]

The pulse waveform design is to find \( w(t) \) to maximize \( P_b(e) \) in (4.0.7). There are a few variations of (4.0.7), for example, given \( \tau \) we can find \( w(t) \) to minimize

\[
P_b(e|\tau) = \int_{\beta,\theta} Q\left(\sqrt{\frac{\xi(\beta, \theta)E_b}{N_0}}\right)p(\beta, \theta)d\beta d\theta \tag{4.0.9}
\]
In the next two sections we will show the methods of pulse optimization and prove several propositions.

4.1 Pulse optimization for PPM on Gaussian channels

In this section, we first prove several propositions and then present the optimization method.

The unconstrained optimization is

\[ w_u(t) = \arg\min_{w(t)} \frac{R(\Delta)}{R(0)} \]  

(4.1.1)

It is desirable to get rid of the DC term in the pulse waveform, the second optimization is then constrained with \( \int_{-\infty}^{\infty} w(t) dt = 0 \). i.e.,

\[ w_c(t) = \arg\min_{w(t)} \left( \frac{R(\Delta)}{R(0)} + \lambda \int_{-\infty}^{\infty} w(t) dt \right) \]  

(4.1.2)

where \( \lambda \) is the Lagrange multiplier. Before we introduce a numerical method, let us study several properties. Let \( \Gamma_u \) and \( \Gamma_c \) denote "Unconstrained" and "Constrained" minimums respectively. Let \( N = \left\lfloor \frac{T_w}{\Delta} \right\rfloor + 1 \).

\[ x_i(t) = \begin{cases} 
  w(t) & (i-1)\Delta \leq t < i\Delta \\
  0 & \text{otherwise}
\end{cases} \text{ for } i = 1, ..., N-1 \]  

(4.1.3)

\[ x_N(t) = \begin{cases} 
  w(t) & N\Delta < t \leq T_w \\
  0 & \text{otherwise}
\end{cases} \]  

(4.1.4)
then, we obtain
\[
\Gamma_{u,\text{min}} = \min_{w(t)} \left( \frac{R(\Delta)}{R(0)} \right) = \min_{x_i} \left[ \frac{\sum_{i=1}^{N-1} \int_0^{T_w} x_i(t - \Delta)x_{i+1}(t)dt}{\sum_{i=1}^N \int_0^{T_w} x_i^2(t)dt} \right]
\]
(4.1.5)

Since,
\[
\int_0^{T_w} x_i(t - \Delta)x_{i+1}(t)dt \geq -\sqrt{\int_0^{T_w} x_i^2(t - \Delta)dt} \sqrt{\int_0^{T_w} x_{i+1}^2(t)dt}
\]
(Cauchy-Schwarz-Buniakowsky inequality), we have
\[
\Gamma_{u,\text{min}} \geq -\min_{X_i} \left[ \frac{\sum_{i=1}^{N-1} X_iX_{i+1}}{\sum_{i=1}^N X_i^2} \right]
\]
(4.1.7)

where,
\[
X_i = \pm \sqrt{\int_0^{T_w} x_i^2(t)dt}
\]
(4.1.8)

**Proposition 1.** If \(N\) is even and if we can find a positive \(\alpha^*\) satisfying the following equations:

\[
\begin{align*}
\alpha_1 &= \alpha \\
\alpha_{i+1} &= \alpha - \frac{1}{4\alpha_i} \quad i = 1, \ldots, \frac{N}{2} \\
\alpha_{N/2} &= \frac{1}{2}
\end{align*}
\]
(4.1.9)

then \(\Gamma_{u,\text{min}} = -\alpha^*\) is the global minimum and

\[
\begin{align*}
X_{i+1} &= -2\alpha_i X_i \quad \text{for } i = 1, \ldots, \frac{N}{2} \\
X_{N-i-1} &= -X_i \quad \text{for } i = 1, \ldots, \frac{N}{2}
\end{align*}
\]
(4.1.10)

**Proof.** Let
\[
A(\alpha) = \sum_{i=1}^{\frac{N}{2}-1} \left[ \left( \sqrt{\alpha_i} X_i + \frac{1}{2\sqrt{\alpha_i}} X_{i+1} \right)^2 + \left( \sqrt{\alpha_{i+1}} X_{N-i} + \frac{1}{2\sqrt{\alpha_{i+1}}} X_{N-i-1} \right)^2 \right] + \left( \sqrt{\alpha_{\frac{N}{2}}} X_{\frac{N}{2}} + \frac{1}{2\sqrt{\alpha_{\frac{N}{2}}}} X_{\frac{N}{2}+1} \right)^2 + \left( \alpha_{\frac{N}{2}} - \frac{1}{4\alpha_{\frac{N}{2}}} \right) X_{\frac{N}{2}+1}^2
\]
(4.1.11)

If \(\alpha^*\) satisfies Eqns. (4.1.9), then there is a set of nonzero \(X_i\), (i.e. (4.1.10)) to have \(A(\alpha^*) = 0\). For any \(\alpha > \alpha^*\), since \(\alpha^*\) is a monotonously increasing function of \(\alpha\), we have \(A(\alpha) = 0\) only when \(X_1 = X_2 = \ldots = X_{\frac{N}{2}} = 0\). By expanding the right hand side of (4.1.11), we obtain
\[
\frac{\sum_{i=1}^{N-1} X_iX_{i+1}}{\sum_{i=1}^N X_i^2} = -\alpha^*
\]
(4.1.12)

Therefore \(\alpha^*\) is the global minimum value of \(\Gamma_{u,\text{min}}\).
It is worth mentioning that for \( N \) even, we have \( \sum_{i=1}^{N} X_i = 0 \). Thus, \( \alpha^* \) is also the global minimum for \( \Gamma_{c,min} \).

**Proposition 2.** If \( N \) is odd and if we can find a positive \( \alpha^* \) satisfying the following equations:

\[
\begin{align*}
\alpha_1 &= \alpha \\
\alpha_{i+1} &= \alpha - \frac{1}{4\alpha_i} & \text{for } i = 1, \ldots, \frac{N-3}{2} \\
\alpha_{\frac{N-3}{2}} &= \frac{1}{2\alpha_{\frac{N-1}{2}}}
\end{align*}
\]

then \( \Gamma_{u,min} \) is the globe minimum and

\[
\begin{align*}
X_{i+1} &= -2\alpha_i X_i & \text{for } i = 1, \ldots, \frac{N-1}{2} \\
X_{N-i} &= -2\alpha_i X_{N-i+1} & \text{for } i = 1, \ldots, \frac{N-2}{2}
\end{align*}
\]

**Proof.** The proof is identical to the proof of 1 except replacing (4.1.11) by

\[
A(\alpha) = \sum_{i=1}^{\frac{N-1}{2}} \left[ \left( \sqrt{\alpha_i} X_i + \frac{1}{2\sqrt{\alpha_i} X_{i+1}} \right)^2 + \left( \sqrt{\alpha_i} X_{N-i+1} + \frac{1}{2\sqrt{\alpha_i} X_{N-i}} \right)^2 \right] + \left( \alpha - \frac{1}{2\alpha_{\frac{N-1}{2}}} \right) X_{\frac{N+1}{2}}^2
\]

\[\square\]

In table 4.1, we list the results for \( N = 2 \) to \( N = 20 \). It is worth mentioning that in order to obtain the best pulse waveform we need to meet the equality in (4.1.6) as well, i.e.,

\[ x_i(t - \Delta) = x_{i+1}(t) \]

Now, let us design the best pulse for several \( N \) values. For \( N=2 \) we have \( X_1 = -X_2 \), that is, \( X_1(t) = -X_2(t - \Delta) \). Many functions satisfy this condition. We give an optimal waveform for \( \Delta = \frac{1}{2} \) in Fig. 4.6.
For \( N=3 \) we have,

\[
X_2 = -\sqrt{2}X_1 \\
X_1 = X_3
\]

\[
x_2(t) = -\sqrt{2}x_1(t - \Delta) \\
x_3(t) = x_1(t - 2\Delta)
\]

For, \( \Delta = \frac{1}{3} \), we construct an optimal waveform given in Fig. 4.2 using half-sine function. It has a DC component.

**Proposition 3.** If \( N \) is odd, we have

\[
\Gamma_{c,\text{min}} = \Gamma_{u,\text{min}}
\]  

(4.1.16)

**Proof.** The proof is simple. If we make each waveform \( x_i(t) \) to have zero mean, then the overall waveform will have zero mean. One way to achieve that is, i.e. \( x_i(t) = x_i(t) + x_i(t) \), where, \( x_i(t) = -x_i(t - \frac{\Delta}{2}) \).

For example, for \( N = 3 \), we have the pulse waveform in Fig. 4.2. However, this may not be good in the real system implementation, since the bandwidth of the signal would be much larger than the pulse in Fig. 4.6.

It is worth mentioning here that if \( \Delta = T_w/N \), then

\[
w(t) = \sum_{i=1}^{N} X_i \delta(t - (i - 1)\Delta)
\]  

(4.1.17)

is also an optimal pulse waveform.

Several interesting aspects have been identified in [?], for example, the BER performance degradation of PPM decreases as \( \Delta \) decreases, but it is a constant over \( \frac{1}{N} < \Delta \leq \frac{1}{N+1} \). In [?], we have further shown a simple way to implement the optimal pulses when \( N \) is large.

Now, we present two numerical optimization methods. These algorithms may not produce global minimum solutions, but they provided some vital insights into the problem.
Method 1: Lagrange discrete optimization

1) Cut a pulse waveform into $N$ jointed sections (as in (4.1.3)).

2) Assume the pulse in each section is a constant, $X_1, X_2, ..., X_N$.

3) Use lagrange optimization method to find $X_1^*, X_2^*, ..., X_N^*$ values which minimizes

$$\Gamma = \frac{\sum_{i=1}^{N} X_i X_{i+1}}{\sum_{i=1}^{N} X_i^2} + \lambda \left( \sum_{i=1}^{N} X_i \right)$$

(4.1.18)

For small $N$, we can apply lagrange optimization to find $X_i$ and $\lambda$. That is,

$$\frac{\partial \Gamma}{\partial X_i} = 0 \text{ and } \frac{\partial \Gamma}{\partial \lambda}$$

(4.1.19)

For a large $N$, we can use Matlab to perform numerical optimization. The following method has also been used for checking our result.

The second search method: Steps 1&2 are identical to lagrange discrete method. Step (3): Search over all $X_i \in [-1, 1]$ numerically.

These two methods can be used to search for pulse waveform with more constraints. For example, if we insist the pulse waveform will have three sections and no DC, we can search over $X_1 = X_3 \in [-1, 0]$, $X_2 \in [0, 1]$, and $X_1 + X_2 + X_3 = 0$.

4.2 Pulse Waveform Optimization for Dense Multi-path Channels

In this section we will first prove that if there is no limitation on what kind of pulse shapes we can select, then the one given in (4.3.1) minimizes the error probability in (4.0.7). Since
the pulse is the sum of several impulse functions, which may not be desirable in real system design. In the second part of this section we will present several waveforms that are numerically optimized and with a limited number of pulse sections.

**Proposition 4.** If $\tau_i - \tau_j \neq k\Delta$ for $i, j = 1, \ldots, M$, and $i \neq j$ and $k$ is a non-zero integer, then the pulse waveform

$$w(t) = \sum_{i=1}^{N} X_i \delta(t - (i-1)\Delta)$$

(4.2.1)

minimizes the error probability in (4.0.7).

**Proof.** The following is a property of the $Q$ function:

$$\frac{dQ(x + k)}{dx} \geq \frac{dQ(x)}{dx} \quad \text{given } k \geq 0 \quad (4.2.2)$$

i.e. derivative of $Q(x)$ is a non-decreasing function. From the inequality (4.2.2) we can conclude that

$$Q(x) \leq \frac{Q(x + z) + Q(x - z)}{2} \quad \text{for any } z \quad (4.2.3)$$

or in other words, if $X$ and $Y$ are two random variables distributed symmetrically around the same mean $\mu$, with variances $\sigma_X^2$ and $\sigma_Y^2$. We can say that

$$\int_{-\infty}^{\infty} Q(X)p_X(x)dx \geq \int_{-\infty}^{\infty} Q(Y)p_Y(y)dy \quad \text{if } \sigma_X^2 \geq \sigma_Y^2 \quad (4.2.4)$$

where $p_X(x)$ and $p_Y(y)$ are probability density functions of $X$ and $Y$ respectively.

Since the error probability is determined by

$$\xi(\beta, \tau, \theta) = \sum_{i=1}^{M} \beta_i^2[R(0) - R(\Delta)] + \psi(\beta, \tau, \theta) \quad (4.2.5)$$

where,

$$\psi(\beta, \tau, \theta) = \sum_{l=1}^{N} \sum_{k=1}^{N} \beta_l \beta_k e^{j(\theta_l - \theta_k)}[R(\tau_k - \tau_l) - R(\tau_k - \tau_l - \Delta)] \quad \text{for } l \neq k \quad (4.2.6)$$

The PDF of $\psi(\beta, \tau, \theta)$ is symmetrical and its mean is zero, if $\theta$ is uniformly distributed over $[-\pi, \pi]$.

If we select the pulse in (4.2.1), we have $\psi = 0$. Thus, we have the minimum error probability as

$$P_b(e) = Q\left(\sqrt{\frac{\sum_{i=1}^{M} \beta_i^2 [1 - \Gamma] E_b}{N_0}}\right) \quad (4.2.7)$$
If $\sum_{i=1}^{M} \beta_i^2 = 1$, (4.0.7) is equal to the error probability on Gaussian channels.

Although the result in the above proposition is nice, the pulse waveforms which form from a set of impulse functions are not desirable in the practice. In the rest of this section, we dedicate to design some good pulses. The pulse design can be performed for fixed $\beta$, $\tau$, $\theta$ or over uniform distributions such as $\beta_x \in (0,1]$ and $\sum_{i=1}^{M} \beta_i^2 = 1$, $\tau_x \in (0, T_w]$, $\theta_x \in (-\pi, \pi]$.

Why $\tau_x \in (0, T_w]$? We study the dense multi-path fading channel in the following way which can significantly reduce the complexity of optimization. It is easy to show that if $|\tau_i - \tau_j| \geq T_w$, $i \neq j$, then the optimal pulse for the multi-path channels is identical to that in Gaussian channels. In dense multi-path channels with delay spread up to $T_d$, we can divide the channel into $\lfloor \frac{T_d}{T_w} \rfloor + 1$ sectors. Letting the $i^{th}$ sector have $M_i$ paths and its $k^{th}$ path have amplitude $\beta_k$, $\tau_k$, and $\theta_k$, then we have $|\tau_k - \tau_q| \in (0, T_w]$. Assuming $M_i = M$, the average BER over the complicated dense multi-path channels can be approximated by the average BER over $M$-path channels defined above. If $M = 2$ and $\lfloor \frac{T_d}{T_w} \rfloor + 1 = 60$, then the dense multi-path channels with 120 paths can be taken into account by two path channels with $\beta_x \in (0,1]$, $\tau_x \in (0, T_w]$, $\theta_x \in (-\pi, \pi]$. We assume $\sum_{i=1}^{M} \beta_i^2 = 1$ for convenience of numerical computation. After the simplification and setting $\theta_1 = 0$, $\tau_1 = 0$, and $\beta_1^2 + \beta_2^2 = 1$, we can then optimize the pulse via three integrations instead of $120 \times 3$ integrations.

The second numerical optimization method is used here. We optimize waveforms to minimize the bit error probability defined in (4.0.7). With $\theta_1 = 0$, $\tau_1 = 0$, and $\beta_1^2 + \beta_2^2 = 1$, 

75
we obtain
\[
\xi_{2\text{path}}(\beta, \tau, \theta) = R(0) - R(\Delta) + \beta_1 \beta_2 R(\tau)(e^{-j\theta_2} + e^{j\theta_2}) - \beta_1 \beta_2 R(\tau - \Delta)e^{-j\theta_2} - \beta_1 \beta_2 R(\tau + \Delta)e^{j\theta_2}
\]

(4.2.8)

In Fig. 4.3 we present the optimal pulses for $\Delta = 1/3$, where $\theta_2$ is uniformly distributed.

In Fig. 4.3 (a) we fixed $\tau_2 = T_w/3$ and let $\beta_2$ be uniformly distributed in $(0, 1]$. In Fig. 4.3 (b), $\tau_2$ is uniformly distributed over $(-\pi, \pi]$ and let $\beta_2$ be either fixed (i.e., $\beta_2 = \beta_1 = \sqrt{2}/2$) or uniformly distributed in $(0, 1]$. For both cases, we have the identical optimal pulse waveform given in Fig. 4.3 (b). In Fig. 4.4 we show the optimal pulse with $\Delta = 1/4$. In Fig. 4.3 and the pulse indicated by the solid line in Fig. 4.4 we construct the waveform using the raised cosine pulse with duration of $\Delta_x = \Delta$. As showed by dashed and dotted lines in Fig. 4.4, we can also construct the pulse waveform using the raised cosine pulse with duration ($\Delta_x$) less than $\Delta$. When $\Delta_x$ decreases, as showed in Fig. 4.5, the performance approaches to that on Gaussian channels. This has been predicted in Proposition 4. Fig. 4.5 presents the simulated BER performance for conventional pulse (a) and optimal pulses given in Fig. 4.4 with $\Delta_x/\Delta = 1, 1/3, \text{and } 1/10$. 76
Table 4.1: Optimized $X_i$s for $i = 1, \ldots, 20$, $\alpha$ and degradation factors

<table>
<thead>
<tr>
<th>N</th>
<th>$\alpha^*$</th>
<th>Dgd(dB)</th>
<th>$X_1$</th>
<th>$X_2$</th>
<th>$X_3$</th>
<th>$X_4$</th>
<th>$X_5$</th>
<th>$X_6$</th>
<th>$X_7$</th>
<th>$X_8$</th>
<th>$X_9$</th>
<th>$X_{10}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>0.500</td>
<td>1.249</td>
<td>0.707</td>
<td>-0.707</td>
<td>0.500</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>0.707</td>
<td>0.688</td>
<td>0.500</td>
<td>-0.707</td>
<td>0.500</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>0.809</td>
<td>0.436</td>
<td>0.372</td>
<td>-0.602</td>
<td>0.602</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>0.866</td>
<td>0.301</td>
<td>0.289</td>
<td>-0.500</td>
<td>0.577</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>0.901</td>
<td>0.220</td>
<td>0.232</td>
<td>-0.418</td>
<td>0.521</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>0.924</td>
<td>0.168</td>
<td>0.191</td>
<td>-0.353</td>
<td>0.462</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>0.940</td>
<td>0.132</td>
<td>0.161</td>
<td>-0.303</td>
<td>0.408</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>0.951</td>
<td>0.108</td>
<td>0.138</td>
<td>-0.263</td>
<td>0.362</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>0.959</td>
<td>0.090</td>
<td>0.120</td>
<td>-0.231</td>
<td>0.323</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>11</td>
<td>0.966</td>
<td>0.074</td>
<td>0.106</td>
<td>-0.204</td>
<td>0.289</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>12</td>
<td>0.971</td>
<td>0.063</td>
<td>0.094</td>
<td>-0.182</td>
<td>0.260</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>13</td>
<td>0.975</td>
<td>0.055</td>
<td>0.084</td>
<td>-0.164</td>
<td>0.236</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>14</td>
<td>0.978</td>
<td>0.048</td>
<td>0.076</td>
<td>-0.149</td>
<td>0.215</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>15</td>
<td>0.981</td>
<td>0.041</td>
<td>0.069</td>
<td>-0.135</td>
<td>0.196</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>16</td>
<td>0.983</td>
<td>0.037</td>
<td>0.063</td>
<td>-0.124</td>
<td>0.181</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>17</td>
<td>0.985</td>
<td>0.033</td>
<td>0.058</td>
<td>-0.114</td>
<td>0.166</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>18</td>
<td>0.986</td>
<td>0.031</td>
<td>0.054</td>
<td>-0.106</td>
<td>0.155</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>19</td>
<td>0.988</td>
<td>0.026</td>
<td>0.049</td>
<td>-0.097</td>
<td>0.143</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>20</td>
<td>0.989</td>
<td>0.024</td>
<td>0.049</td>
<td>-0.091</td>
<td>0.133</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>N</th>
<th>$\alpha^*$</th>
<th>Dgd(dB)</th>
<th>$X_{11}$</th>
<th>$X_{12}$</th>
<th>$X_{13}$</th>
<th>$X_{14}$</th>
<th>$X_{15}$</th>
<th>$X_{16}$</th>
<th>$X_{17}$</th>
<th>$X_{18}$</th>
<th>$X_{19}$</th>
<th>$X_{20}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>11</td>
<td>0.106</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>12</td>
<td>0.182</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>13</td>
<td>0.236</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>14</td>
<td>0.272</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>15</td>
<td>0.294</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>16</td>
<td>0.307</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>17</td>
<td>0.313</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>18</td>
<td>0.314</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>19</td>
<td>0.313</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>20</td>
<td>0.309</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

77
4.3 Aspects of Optimal Performance and Optimal Pulses

In this section, we summarize several aspects of optimal performance and optimal pulses. The completed proof of the optimality can be found in [55]. Some results and numerical methods can be found in [54].

Aspect 1: The BER performance degradation of PPM decreases as $\Delta$ decreases, but it is a constant over $\frac{1}{N} < \Delta \leq \frac{1}{N+1}$. This aspect is illustrated in Fig. 4.6. For example, if
Figure 4.3: Optimal pulse waveforms for multi-path channels with $\Delta = \frac{1}{3}$, (a) $\beta \in (0, 1]$, $\tau = \frac{T_w}{3}$, $\theta \in (-\pi, \pi]$; (b) $\tau \in (0, T_w]$, $\theta \in (-\pi, \pi]$, $\beta = \sqrt{2}/2$ or $\beta \in (0, 1]$.

If $\Delta = 0.4$, then the best achievable BER performance is 0.69 dB away from the performance of BPSK signals. For conventional pulse (a), we have $\Delta = 0.285$ and $(E_b/N_0)_{deg,a} = 0.875dB$. Compared with the optimal achievable performance, which is $(E_b/N_0)_{deg} = 0.436dB$, the gap is 0.439dB.

**Aspect 2:** The optimal performance is in the following form:

$$w(t) = \sum_{i=1}^{N} X_i p(t - (i - 1)\Delta)$$  

(4.3.1)

where $p(t)$ is a time-limited pulse waveform, i.e., $p(t) = 0$ if $t > T_w - (N - 1)\Delta$ and $t < 0$. $X_i$, for $i = 1, ..., N$ needs to be optimized to minimize the bit error rate. $N = \left\lfloor \frac{T_w}{\Delta} \right\rfloor + 1$, where $\lfloor x \rfloor$ denotes the largest integer less than $x$. The optimal pulse could be constructed by conventional pulses with smaller duration (i.e., duration of $T_w - N\Delta$, rather than $T_w$).

The impulse signal can also form an optimal pulse waveform, i.e., $p(t) = \delta(t)$.

How to select $\Delta$? If we select $\frac{1}{N} < \Delta < \frac{1}{N+1}$, then the optimal pulse will return to zero at many regions, i.e., $w(t) = 0$ for $n\Delta - (T_w - (N - 1)\Delta) < t < (n + 1)\Delta$ for $n = 1, 2, ..., N - 1$. 

79
Figure 4.4: Optimal pulse waveforms for multi-path with $\Delta = \frac{1}{4}, \beta \in [0, 1], \tau \in [0, T_w], \theta \in [-\pi, \pi]$, and $\Delta_x/\Delta =$1 (solid), 1/3 (dashed), 1/10 (dotted)

For example, if $\Delta = 0.285$, then $N = 4$ and the optimal pulse must return to zero in the following regions $t \in (0.145, 0.285]$, $t \in (0.43, 0.57]$, and $t \in (0.715, 0.855]$. Fig. 4.2 illustrates the optimal pulses for $\Delta = 0.285$. Clearly, if we do not like the regions returning to zero, then the only choice would be $\Delta = \frac{1}{k}$, where $k$ is an integer.

Aspect 3: The optimal values of $X_i$ have alternative sign, i.e., $\text{sgn}(X_i) = -\text{sgn}(X_{i+1})$.

It is true since the optimal pulse will make each term in Equations (30) and (34) in [55] to be zero. As showed in Table I in [55], for $\Delta = 1/10$, we have $X_1, ..., X_{10} = 0.12, -0.231, 0.323, -0.388, 0.421, -0.421, 0.388, -0.323, 0.231, -0.12$; for $\Delta = 1/20$, we have $X_1, ..., X_{20} = 0.049, -0.091, 0.133, -0.173, 0.209, -0.241, 0.267, -0.287, 0.301, -0.309, 0.309, -0.301, 0.287, -0.267, 0.241, -0.209, 0.173, -0.133, 0.091, -0.049$. These pulses seem difficult to be implemented, but in the next section we will show a simple way to approximate the optimal pulse waveform and its performance degradation.
4.4 A Simple Way to Implement Optimal Pulses for Small Modulation Index

In this section, we will present a simple way to implement the optimal pulses for small modulation index and its performance degradation due to approximation.

When \( \Delta \) is large, \( N \) is a small integer. When \( \Delta \) is small, then \( N \) becomes large, thus there are many regions needed to be considered for generating \( w(t) \). However, we noted that when \( N \) is large, the shape of \( |X_i| \) can be well-approximated by a Gaussian shape.

In Fig. 4.7, we illustrate \( |X_i| \) and \( g(t) \) with \( \sigma = 0.79 \) and \( a = 0.193 \) for \( |t| < 1 \), where

\[
g(t) = \frac{1}{\sqrt{2\pi}\sigma}exp\left(-\frac{t^2}{2\sigma^2}\right) - a \tag{4.4.1}
\]

In Fig. 4.8 (the solid line), we illustrate an optimal pulse waveform with \( p(t) \) as a half-sine waveform. As can see from this figure, the optimal pulse is very close to the product of \( |X_i| \).
Figure 4.6: Optimal BER degradation for UWB/PPM systems as a function of $N$, where $rac{1}{N} < \Delta \leq \frac{1}{N+1}$ and a sine waveform. Therefore, we can approximate the optimal waveform by generating $g(t)$ first, then multiple $g(t)$ by a sine function, as shown in Fig. 4.8 (the dotted line).

The performance degradation for the approximated waveform is 0.0248 dB. Compared with the optimal degradation, the loss is marginal. In Fig. 7, we plot the power spectral density for the optimal and approximated waveforms. Furthermore, we can process the approximated waveforms to comply with the FCC’s UWB emission mask given in Section 1.5. The resulting waveform and the spectrum of it is shown in Figs. 4.13 and 4.14 respectively.

We also go through the procedure for $\Delta = 1/10$. We find the waveform can be approximated by $g(t)$ with $\sigma = 0.74$ and $a = 0.11$. The degradation is 0.0924 dB. Figs. 4.10, 4.11, and 4.12 show the corresponding results for $\Delta = 1/10$. 

82
Figure 4.7: Comparison of $|X_i|$ and $g(t)$ for $\Delta = 1/20$.

Figure 4.8: Comparison of the optimal and the approximated waveforms for $\Delta = 1/20$. 
Figure 4.9: Comparison of PSDs of the optimal and the approximated waveforms for $\Delta = 1/20$.

Figure 4.10: Comparison of $|X_i|$ and $g(t)$ for $\Delta = 1/10$. 
Figure 4.11: Comparison of the optimal and the approximated waveforms for $\Delta = 1/10$.

Figure 4.12: Comparison of PSDs of the optimal and the approximated waveforms for $\Delta = 1/10$. 
Figure 4.13: Comparison of the approximated and filtered waveforms for $\Delta = 1/20$.

Figure 4.14: PSD for the optimized filtered UWB waveform overlapped with UWB emission mask $\Delta = 1/20$. 
CHAPTER 5
SIMPLE ROBUST RECEIVER DESIGN AND
PULSE OPTIMIZATION FOR UWB CHANNEL

5.1 Introduction

Previous work presented in this dissertation has shown that we can achieve in theory, better performance using pulse position modulation (PPM) with optimized pulses compared with commonly used on-off keying systems in UWB communication channels. However, this theory is in fact hard to put in use in practical applications due to time sensitivity and unpredictable behavior of the uwb channel. In this chapter, we design a practical UWB receiver which is simple to build, robust to changes in the communication channel, and performs comparable with the optimal OOK performance. In this chapter, first, we will introduce the simple receivers along with the signal and system diagrams of OOK systems, then we will present the findings of our investigations on performance of these receivers in AWGN channels while optimizing the signals and receiver parameters. Then we will expand our investigation to generalized gaussian channels. Finally we will investigate their performance in UWB channels.
5.2 Simple Receivers

Fig. 5.1 shows the block diagram of a simple receiver. The received signal is first processed through the low-pass filter of bandwidth $W$. The simple receivers use $|.|^\beta$ to process the signal before making a decision. In this operation the excess noise can affect the system performance greatly. The bandwidth of the low-pass filter is critical for simple receivers. The bandwidth of the receiver low-pass filter has to be optimized according the the pulse shape being transmitted and the nature of the noise. The output of the filter is sampled and goes through $|.|^\beta$ operation. The resulting signal is compared to the optimum threshold for a decision for binary ”0” and ”1”. The simple receiver is an ideal substitute to the optimal receiver due to its simple circuitry which consists of a low order butterworth low-pass filter, $|.|^\beta$ operator (which is especially simple to realize with certain values of $\beta$), and a decision comparator.
5.3 Effect of Excess Noise

In simple receivers the transmitted pulses have majority of their power in a limited bandwidth. It is a common sense to filter the noise that exceeds the transmitted signal bandwidth. At the input of the decision block, the signal has a signal component and a noise component. The filter bandwidth has a great effect on the noise component of this signal. Sometimes, it is wise to filter some of the signal energy to keep the noise component to a low value. In order to demonstrate the effect of noise in simple receivers, we investigate an example.

An UWB mono-pulse of duration $T_p = 1\text{ns}$ and Energy $E_p = 1\text{W}$ is transmitted into an AWGN channel. For this example we use signal power $P_s = 1\text{W}$ and Noise power $P_N = 1\text{W}$. At the receiver, we sample the signal at $F_s = 100\text{GHz}$, which corresponds to 100 samples/pulse duration. Due to the sampling frequency, the noise bandwidth is contained in $F_s/2\text{Hz} = 50\text{GHz}$. However, the UWB signal of duration $1\text{ns}$ has most of its power contained in $3\text{Ghz}$. It is common sense to set the filter bandwidth to $3\text{GHz}$ however investigating the signal component to noise component ratio at the decision block input from Table 5.1 reveals that setting the filter bandwidth at $2\text{GHz}$ results in a better performance. For each case of $\beta$, the signal component to total signal ratio at the decision block input is highest when $W = 2\text{GHz}$. This study shows different results for each pulse candidate suggested in section 5.4. Hence, we will optimize the bandwidth of the receiver filter for each pulse candidate in section 5.5.1. One should not compare the $\beta$ values at this point from the table 5.1, in other words, at 2 GHz, $\beta = 0.5$ has greater ratio of signal component to total power.
Table 5.1: Signal component the total signal ratio at the decision block input for
$\beta = [0.5, 1, 2, 4]$ and receiver bandwidth $W = [50, 1, 2, 3, 4]$ GHz

<table>
<thead>
<tr>
<th>W</th>
<th>$\beta = 0.5$</th>
<th>$\beta = 1$</th>
<th>$\beta = 2$</th>
<th>$\beta = 4$</th>
</tr>
</thead>
<tbody>
<tr>
<td>50 GHz</td>
<td>0.8265</td>
<td>0.6924</td>
<td>0.4970</td>
<td>0.2510</td>
</tr>
<tr>
<td>1 GHz</td>
<td>0.9724</td>
<td>0.9579</td>
<td>0.9137</td>
<td>0.7797</td>
</tr>
<tr>
<td>2 GHz</td>
<td>0.9830</td>
<td>0.9725</td>
<td>0.9378</td>
<td>0.8617</td>
</tr>
<tr>
<td>3 GHz</td>
<td>0.9799</td>
<td>0.9663</td>
<td>0.9288</td>
<td>0.8536</td>
</tr>
<tr>
<td>4 GHz</td>
<td>0.9762</td>
<td>0.9563</td>
<td>0.9137</td>
<td>0.8309</td>
</tr>
</tbody>
</table>

signal than $\beta = 1$, however, this does not mean that $\beta = 0.5$ performs better for the given
pulse. There are other variables involved with optimization of $\beta$ parameter. This issue is
addressed in section 5.5

5.4 Pulse Candidates

Fig. 5.2 shows 4 possible pulse candidates suitable for simple receiver applications. In
the next sections, we will evaluate the performance of these 4 pulses, and use the optimal
pulse candidate for our calculation and simulations. In the next sections we will refer to
these pulses as pulse (a): Traditional UWB pulse, pulse (b): Half Cosine pulse, pulse (c):
Rectangular pulse, and pulse (d): Sinc pulse.
Figure 5.2: Four Pulses to be evaluated: a) Traditional UWB Pulse (Gaussian), 2) Half Cosine Pulse, 3) Rectangular Pulse, 4) Sinc Pulse. All Pulses have pulse duration of $T_p = 1\text{ns}$

### 5.5 Performance Optimization in Additive White Gaussian noise (AWGN) Channels

In this section, we first optimize the pulse candidates and the corresponding receiver bandwidth for each pulse candidates given in section 5.4. Then, we will optimize the receiver power $\beta$ for best performance. Finally, we will display the performance of optimized simple receiver in AWGN channels. In AWGN channels the transmitted pulse is distorted with the additive gaussian noise during transmission. The probability density function of the noise can be given with:

$$f(x|\mu, \sigma) = \frac{1}{\sqrt{2\pi\sigma}} e^{-\frac{(x-\mu)^2}{2\sigma^2}}$$  \hspace{1cm} (5.5.1)
Table 5.2: Energy Capture of Pulses for various filter bandwidths

<table>
<thead>
<tr>
<th>Pulse (a)</th>
<th>W = 1/T_p</th>
<th>W = 2/T_p</th>
<th>W = 3/T_p</th>
<th>W = 4/T_p</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pulse (b)</td>
<td>% 17.43</td>
<td>% 63.63</td>
<td>% 87.33</td>
<td>% 95.17</td>
</tr>
<tr>
<td>Pulse (c)</td>
<td>% 83.61</td>
<td>% 98.04</td>
<td>% 99.46</td>
<td>% 99.79</td>
</tr>
<tr>
<td>Pulse (d)</td>
<td>% 72.04</td>
<td>% 86.09</td>
<td>% 90.74</td>
<td>% 93.07</td>
</tr>
<tr>
<td>Pulse (e)</td>
<td>% 55.87</td>
<td>% 88.12</td>
<td>% 96.77</td>
<td>% 98.88</td>
</tr>
</tbody>
</table>

where $\mu = 0$, and $\sigma^2$ is the spectral noise density. In simple receivers, the received signal will go through $|.|^\beta$ operator. The probability density function at the output of the operator can be given as:

$$f(x|x_i, \sigma, \beta) = \frac{1}{\sqrt{2\pi t\sigma\beta}} e^{-\frac{|x_i - x|}{2\sigma^2}} + \frac{1}{\sqrt{2\pi t\sigma\beta}} e^{-\frac{|x_i + x|}{2\sigma^2}}$$  \hspace{1cm} (5.5.2)

where $x_i$ is the $i^{th}$ sample of the transmitted pulse, $\sigma^2$ is the spectral noise density of the noise, and $\beta$ is the receiver power.

### 5.5.1 Receiver Parameter Optimization in AWGN Channels

In section 5.3, we explained the importance of the receiver filter bandwidth and suppression of excess noise. In this section, we will optimize the filter bandwidth $W$ to minimize error performance for different pulses suggested in section 5.4. We will also optimize the filter power $\beta$. First we investigate the energy capture of the pulses at different $W$ values. We will evaluate the optimal bandwidth with respect to the pulse duration $T_p$. In order to keep the receiver circuitry simple, we will assume the filter is a second order butterworth filter with corner frequency $W$. 

92
Figure 5.3: 4 pulse candidates after the receiver filter with bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p} \right]$.

Next we test the effect of the noise filtering. Increasing the filter bandwidth captures more energy of the transmitted pulse, however, it also allows more noise to get into the receiver. We will investigate the optimal bandwidth at $12dB$ signal to noise ratio. By investigating the probability of error calculations with $\beta = 0.5$ (fig. 5.4, $\beta = 1$ (fig. 5.5, $\beta = 2$ (fig. 5.6), and $\beta = 4$ (fig. 5.7), we conclude that the optimal bandwidth for $\beta = 0.5$ is $W = \frac{2}{T_p}$ using the rectangular pulse, and for $\beta = [1, 2, 4]$, it is $W = \frac{1}{T_p}$ using half cosine pulse. The error probabilities corresponding to each pulse at investigated bandwidths is given in table 5.5.1.
Table 5.3: Optimal Filter Bandwidth for 4 pulse candidates in AWGN Channel

<table>
<thead>
<tr>
<th>Pulse</th>
<th>$\beta = 0.5$</th>
<th>$\beta = 1$</th>
<th>$\beta = 2$</th>
<th>$\beta = 4$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$W$</td>
<td>$W$</td>
<td>$W$</td>
<td>$W$</td>
</tr>
<tr>
<td>Pulse (a)</td>
<td>$\frac{3}{T_p}$</td>
<td>$\frac{3}{T_p}$</td>
<td>$\frac{3}{T_p}$</td>
<td>$\frac{3}{T_p}$</td>
</tr>
<tr>
<td>Err. Prob.</td>
<td>0.0035</td>
<td>0.0009</td>
<td>0.0011</td>
<td>0.0054</td>
</tr>
<tr>
<td>Pulse (b)</td>
<td>$\frac{2}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
</tr>
<tr>
<td>Err. Prob.</td>
<td>0.00015</td>
<td>0.0003</td>
<td>0.0002</td>
<td>0.0014</td>
</tr>
<tr>
<td>Pulse (c)</td>
<td>$\frac{2}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
<td>$\frac{1}{T_p}$</td>
</tr>
<tr>
<td>Err. Prob.</td>
<td>0.000065</td>
<td>0.00039</td>
<td>0.0007</td>
<td>0.0039</td>
</tr>
<tr>
<td>Pulse (d)</td>
<td>$\frac{2}{T_p}$</td>
<td>$\frac{2}{T_p}$</td>
<td>$\frac{2}{T_p}$</td>
<td>$\frac{2}{T_p}$</td>
</tr>
<tr>
<td>Err. Prob.</td>
<td>0.0041</td>
<td>0.0005</td>
<td>0.0007</td>
<td>0.0022</td>
</tr>
</tbody>
</table>

Finally, we can also conclude from the table 5.5.1 that the best performing pulse, bandwidth and receiver power combination for AWGN channels is the Half cosine pulse (pulse (b)) with bandwidth $W = 1/T_p$ and receiver power $\beta = 1$. In section 5.5.2, we compute the error performance of simple receivers in AWGN channel with optimized parameters and compare the performance to the optimal OOK performance in AWGN channels using match filter receiver.
Figure 5.4: Probability of Error for 4 pulse candidates at 12dB SNR for $W = \left[ \frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p} \frac{5}{T_p} \right]$ with $\beta = 0.5$

Figure 5.5: Probability of Error for 4 pulse candidates at 12dB SNR for $W = \left[ \frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p} \frac{5}{T_p} \right]$ with $\beta = 1$
Figure 5.6: Probability of Error for 4 pulse candidates at 12dB SNR for $W = \left[ \frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p} \frac{5}{T_p} \right]$ with $\beta = 2$

Figure 5.7: Probability of Error for 4 pulse candidates at 12dB SNR for $W = \left[ \frac{1}{T_p} \frac{2}{T_p} \frac{3}{T_p} \frac{4}{T_p} \frac{5}{T_p} \right]$ with $\beta = 4$
5.5.2 Error Performance in AWGN Channels

In this section, we evaluate the performance of simple receivers in AWGN channels, and compare it to the performance of the optimal match filter receiver. Since $W = \frac{1}{T_p}$, which yields to $u = 2$, where $u = 2WT_p$, we have to use density evolution to get accurate probability density functions at the decision block input. We also use monte carlo simulation to verify the results using a second order butterworth filter with bandwidth $W$ as receiver low-pass filter. The results are shown in fig.5.8. We conclude that using a simple receiver yields to a performance degradation of $1.4dB$, which makes simple receivers a strong candidate in AWGN channels.

![Figure 5.8: Probability of Error for Simple Receivers in AWGN channels with Half-Cosine pulse, $\beta = 1$, $W = \frac{1}{T_p}$](image)
5.6 Performance in Additive Generalized Gaussian Noise (AGGN) Channels

In addition to the AWGN channel, we will optimize and evaluate the performance of simple receivers in Additive Generalized Gaussian Noise Channels. Furthermore, we will test the robustness of the simple receivers in AGGN channels.

Noise in AGGN channels can be statistically modeled with the following pdf:

\[
p(x; A_i) = \frac{\gamma}{2\sigma_i \sqrt{a(\gamma)} \Gamma(1/\gamma)} \exp\left(-\frac{|x|\gamma}{[a(\gamma)]^{\gamma/2}\sigma_i}\right)
\]  

(5.6.1)

where \(\sigma^2\) is the noise spectral density and \(\gamma\) is the noise parameter. When \(\gamma\) is small the noise becomes impulsive. This AGGN noise is identical to AWGN noise when \(\gamma = 2\), and identical to Laplacian Noise when \(\gamma = 1\). We will evaluate the performance of simple receivers for \(\gamma = [0.514]\). For the case of \(\gamma = 2\), the results are already shown in section 5.5. We will also evaluate the performance on a simulated channel that \(\gamma\) varies with time to test robustness of the receiver.

5.6.1 Receiver Parameter Optimization in AGGN Channels

In this section we optimize the receiver parameters \(W, \beta\), and the pulse candidates to find best performing simple receiver in AGGN noise with various \(\gamma\) parameter. We have established the optimal parameters for \(\gamma = 2\) in section 5.6.1. We will perform same type
Table 5.4: Probability of Bit Error for filter bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and Filter power $\beta = [0.5, 1, 2, 4]$ for 4 pulse candidates in AGGN Channel Channels with $\gamma = [0.5]$

<table>
<thead>
<tr>
<th>$\gamma$</th>
<th>$\beta$</th>
<th>$W$</th>
<th>Pulse (a)</th>
<th>Pulse (b)</th>
<th>Pulse (c)</th>
<th>Pulse (d)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>0.5</td>
<td>$\frac{1}{T_p}$</td>
<td>0.090040</td>
<td>0.000485</td>
<td>0.000152</td>
<td>0.008160</td>
</tr>
<tr>
<td>0.5</td>
<td>0.5</td>
<td>$\frac{2}{T_p}$</td>
<td>0.005730</td>
<td>0.000330</td>
<td>0.000119</td>
<td>0.005554</td>
</tr>
<tr>
<td>0.5</td>
<td>0.5</td>
<td>$\frac{3}{T_p}$</td>
<td>0.002908</td>
<td>0.000515</td>
<td>0.000163</td>
<td>0.006246</td>
</tr>
<tr>
<td>0.5</td>
<td>0.5</td>
<td>$\frac{4}{T_p}$</td>
<td>0.003066</td>
<td>0.000716</td>
<td>0.000216</td>
<td>0.007576</td>
</tr>
<tr>
<td>0.5</td>
<td>0.5</td>
<td>$\frac{5}{T_p}$</td>
<td>0.003575</td>
<td>0.000898</td>
<td>0.000261</td>
<td>0.008592</td>
</tr>
<tr>
<td>0.5</td>
<td>1</td>
<td>$\frac{1}{T_p}$</td>
<td>0.063601</td>
<td>0.000062</td>
<td>0.000088</td>
<td>0.001649</td>
</tr>
<tr>
<td>0.5</td>
<td>1</td>
<td>$\frac{2}{T_p}$</td>
<td>0.002197</td>
<td>0.001118</td>
<td>0.000147</td>
<td>0.000912</td>
</tr>
<tr>
<td>0.5</td>
<td>1</td>
<td>$\frac{3}{T_p}$</td>
<td>0.001254</td>
<td>0.000289</td>
<td>0.000259</td>
<td>0.001461</td>
</tr>
<tr>
<td>0.5</td>
<td>1</td>
<td>$\frac{4}{T_p}$</td>
<td>0.001541</td>
<td>0.000527</td>
<td>0.000404</td>
<td>0.002329</td>
</tr>
<tr>
<td>0.5</td>
<td>1</td>
<td>$\frac{5}{T_p}$</td>
<td>0.002092</td>
<td>0.000814</td>
<td>0.000571</td>
<td>0.003301</td>
</tr>
<tr>
<td>0.5</td>
<td>2</td>
<td>$\frac{1}{T_p}$</td>
<td>0.065499</td>
<td>0.000502</td>
<td>0.001318</td>
<td>0.003054</td>
</tr>
<tr>
<td>0.5</td>
<td>2</td>
<td>$\frac{2}{T_p}$</td>
<td>0.003561</td>
<td>0.001046</td>
<td>0.002005</td>
<td>0.001432</td>
</tr>
<tr>
<td>0.5</td>
<td>2</td>
<td>$\frac{3}{T_p}$</td>
<td>0.002659</td>
<td>0.002210</td>
<td>0.003380</td>
<td>0.002255</td>
</tr>
<tr>
<td>0.5</td>
<td>2</td>
<td>$\frac{4}{T_p}$</td>
<td>0.003829</td>
<td>0.004125</td>
<td>0.005572</td>
<td>0.003990</td>
</tr>
<tr>
<td>0.5</td>
<td>2</td>
<td>$\frac{5}{T_p}$</td>
<td>0.006069</td>
<td>0.007003</td>
<td>0.008752</td>
<td>0.006688</td>
</tr>
<tr>
<td>0.5</td>
<td>4</td>
<td>$\frac{1}{T_p}$</td>
<td>0.053079</td>
<td>0.002829</td>
<td>0.005500</td>
<td>0.005152</td>
</tr>
<tr>
<td>0.5</td>
<td>4</td>
<td>$\frac{2}{T_p}$</td>
<td>0.015548</td>
<td>0.010162</td>
<td>0.017329</td>
<td>0.006912</td>
</tr>
<tr>
<td>0.5</td>
<td>4</td>
<td>$\frac{3}{T_p}$</td>
<td>0.019308</td>
<td>0.026043</td>
<td>0.040620</td>
<td>0.015121</td>
</tr>
<tr>
<td>0.5</td>
<td>4</td>
<td>$\frac{4}{T_p}$</td>
<td>0.032868</td>
<td>0.053485</td>
<td>0.078118</td>
<td>0.030544</td>
</tr>
<tr>
<td>0.5</td>
<td>4</td>
<td>$\frac{5}{T_p}$</td>
<td>0.055134</td>
<td>0.092253</td>
<td>0.126253</td>
<td>0.054374</td>
</tr>
</tbody>
</table>

of analysis for $\gamma = [0.514]$. We calculate the error probability of simple receiver for each $\gamma$ value for four pulse candidates and $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and $\beta = [0.5124]$.

We can conclude from the tables 5.6.1, 5.6.1, and 5.6.1 that the half cosine pulse (pulse b) performs best with bandwidth $W = \frac{1}{T_p}$ and receiver power $\beta = 1.0$. This result is consistent with the case of AWGN channel.
Table 5.5: Probability of Bit Error for filter bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and Filter power $\beta = [0.5, 1, 2, 4]$ for 4 pulse candidates in AGGN Channel Channels with $\gamma = [1.0]$

<table>
<thead>
<tr>
<th>$\gamma$</th>
<th>$\beta$</th>
<th>$W$</th>
<th>Pulse (a)</th>
<th>Pulse (b)</th>
<th>Pulse (c)</th>
<th>Pulse (d)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.5</td>
<td>$\frac{1}{T_p}$</td>
<td>0.096377</td>
<td>0.000447</td>
<td>0.000136</td>
<td>0.007935</td>
</tr>
<tr>
<td>1</td>
<td>0.5</td>
<td>$\frac{2}{T_p}$</td>
<td>0.006153</td>
<td>0.000305</td>
<td>0.000108</td>
<td>0.005800</td>
</tr>
<tr>
<td>1</td>
<td>0.5</td>
<td>$\frac{3}{T_p}$</td>
<td>0.003283</td>
<td>0.000520</td>
<td>0.000170</td>
<td>0.006905</td>
</tr>
<tr>
<td>1</td>
<td>0.5</td>
<td>$\frac{4}{T_p}$</td>
<td>0.003608</td>
<td>0.000801</td>
<td>0.000272</td>
<td>0.008475</td>
</tr>
<tr>
<td>1</td>
<td>0.5</td>
<td>$\frac{5}{T_p}$</td>
<td>0.004500</td>
<td>0.001138</td>
<td>0.000422</td>
<td>0.009806</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\frac{1}{T_p}$</td>
<td>0.057386</td>
<td>0.000054</td>
<td>0.000078</td>
<td>0.001262</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\frac{2}{T_p}$</td>
<td>0.001593</td>
<td>0.000087</td>
<td>0.000116</td>
<td>0.000582</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\frac{3}{T_p}$</td>
<td>0.000872</td>
<td>0.000222</td>
<td>0.000232</td>
<td>0.000954</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\frac{4}{T_p}$</td>
<td>0.001095</td>
<td>0.000432</td>
<td>0.000393</td>
<td>0.001560</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\frac{5}{T_p}$</td>
<td>0.001557</td>
<td>0.000719</td>
<td>0.000627</td>
<td>0.002265</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>$\frac{1}{T_p}$</td>
<td>0.050514</td>
<td>0.000345</td>
<td>0.000940</td>
<td>0.002173</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>$\frac{2}{T_p}$</td>
<td>0.002146</td>
<td>0.000623</td>
<td>0.001264</td>
<td>0.000818</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>$\frac{3}{T_p}$</td>
<td>0.001413</td>
<td>0.001133</td>
<td>0.001859</td>
<td>0.001113</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>$\frac{4}{T_p}$</td>
<td>0.001734</td>
<td>0.001800</td>
<td>0.002627</td>
<td>0.001678</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>$\frac{5}{T_p}$</td>
<td>0.002345</td>
<td>0.002631</td>
<td>0.003567</td>
<td>0.002420</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>$\frac{1}{T_p}$</td>
<td>0.043057</td>
<td>0.002166</td>
<td>0.004211</td>
<td>0.004008</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>$\frac{2}{T_p}$</td>
<td>0.008229</td>
<td>0.005146</td>
<td>0.008292</td>
<td>0.003697</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>$\frac{3}{T_p}$</td>
<td>0.007577</td>
<td>0.009189</td>
<td>0.013243</td>
<td>0.005989</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>$\frac{4}{T_p}$</td>
<td>0.010010</td>
<td>0.014100</td>
<td>0.019262</td>
<td>0.009260</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>$\frac{5}{T_p}$</td>
<td>0.013649</td>
<td>0.019943</td>
<td>0.026460</td>
<td>0.013224</td>
</tr>
</tbody>
</table>
Table 5.6: Probability of Bit Error for filter bandwidth $W = \left[ \frac{1}{T_p}, \frac{2}{T_p}, \frac{3}{T_p}, \frac{4}{T_p}, \frac{5}{T_p} \right]$ and Filter power $\beta = [0.5, 1, 2, 4]$ for 4 pulse candidates in AGGN Channel Channels with $\gamma = [4.0]$

<table>
<thead>
<tr>
<th>$\gamma$</th>
<th>$\beta$</th>
<th>$W$</th>
<th>Pulse (a)</th>
<th>Pulse (b)</th>
<th>Pulse (c)</th>
<th>Pulse (d)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>0.5</td>
<td>$\frac{1}{T_p}$</td>
<td>0.088498</td>
<td>0.000340</td>
<td>0.000101</td>
<td>0.006576</td>
</tr>
<tr>
<td>4</td>
<td>0.5</td>
<td>$\frac{2}{T_p}$</td>
<td>0.005400</td>
<td>0.000233</td>
<td>0.000082</td>
<td>0.004733</td>
</tr>
<tr>
<td>4</td>
<td>0.5</td>
<td>$\frac{3}{T_p}$</td>
<td>0.002921</td>
<td>0.000422</td>
<td>0.000140</td>
<td>0.005800</td>
</tr>
<tr>
<td>4</td>
<td>0.5</td>
<td>$\frac{4}{T_p}$</td>
<td>0.003457</td>
<td>0.000698</td>
<td>0.000265</td>
<td>0.007410</td>
</tr>
<tr>
<td>4</td>
<td>0.5</td>
<td>$\frac{5}{T_p}$</td>
<td>0.004600</td>
<td>0.001053</td>
<td>0.000480</td>
<td>0.008813</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>$\frac{1}{T_p}$</td>
<td>0.060920</td>
<td>0.000065</td>
<td>0.000091</td>
<td>0.001430</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>$\frac{2}{T_p}$</td>
<td>0.001643</td>
<td>0.000094</td>
<td>0.000122</td>
<td>0.000618</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>$\frac{3}{T_p}$</td>
<td>0.000851</td>
<td>0.000221</td>
<td>0.000221</td>
<td>0.000943</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>$\frac{4}{T_p}$</td>
<td>0.001035</td>
<td>0.000420</td>
<td>0.000391</td>
<td>0.001467</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>$\frac{5}{T_p}$</td>
<td>0.001467</td>
<td>0.000709</td>
<td>0.000652</td>
<td>0.002076</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>$\frac{1}{T_p}$</td>
<td>0.045787</td>
<td>0.000300</td>
<td>0.000824</td>
<td>0.001947</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>$\frac{2}{T_p}$</td>
<td>0.001635</td>
<td>0.000505</td>
<td>0.000997</td>
<td>0.000678</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>$\frac{3}{T_p}$</td>
<td>0.000950</td>
<td>0.000830</td>
<td>0.001311</td>
<td>0.000832</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>$\frac{4}{T_p}$</td>
<td>0.001058</td>
<td>0.001209</td>
<td>0.001699</td>
<td>0.001144</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>$\frac{5}{T_p}$</td>
<td>0.001317</td>
<td>0.001633</td>
<td>0.002136</td>
<td>0.001519</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>$\frac{1}{T_p}$</td>
<td>0.039046</td>
<td>0.002013</td>
<td>0.003931</td>
<td>0.003690</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>$\frac{2}{T_p}$</td>
<td>0.006466</td>
<td>0.004112</td>
<td>0.006603</td>
<td>0.002948</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>$\frac{3}{T_p}$</td>
<td>0.005178</td>
<td>0.006227</td>
<td>0.008763</td>
<td>0.004141</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>$\frac{4}{T_p}$</td>
<td>0.006042</td>
<td>0.008109</td>
<td>0.010601</td>
<td>0.005598</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>$\frac{5}{T_p}$</td>
<td>0.007320</td>
<td>0.009760</td>
<td>0.012130</td>
<td>0.007011</td>
</tr>
</tbody>
</table>
5.6.2 Error Performance in AGGN Channels

In this section, we will calculate the error probability of simple receiver in Additive Generalized Gaussian Noise Channels with the optimized parameters from the previous section.

Figure 5.9 shows the probability of bit error for simple receivers in AGGN channels with $\gamma = [0.5, 1, 4]$, where $\gamma = 1$ corresponds to Laplacian Noise. The performance is similar to the performance in AWGN channels. This result is not surprising since the AGGN noise samples with $\gamma = [0.5, 1, 4]$ has identical means and variances as $\gamma = 2$ (AWGN) noise. Hence, the sum of $N$ samples will be distributed with Normal distribution due to the Central Limit Theorem when $N$ is large. However, the impulsive property of the additive noise when $\gamma$ is small effects the noise that passes through the filter. As a result, for low SNR situations, simple receivers perform better in impulsive noise. For high SNR situations, simple receiver performs better in distributed noise (high $\gamma$).

5.7 Performance in Ultra Wide Band (UWB) Channels

The simple receivers are designed to eliminate the timing issues of UWB channel and simplify the popular complex rake receivers. If we can optimize the simple receivers parameters to perform comparable to the rake receivers, then simple receivers can be a strong low-cost candidate against the popular complex receivers.
In the analysis of simpler receiver performance for both AWGN and AGGN channels, we assumed that the channel impulse response was $\delta(t)$, which does not have an effect on signal shape during the transmission. In UWB channel, the channel impulse response is no longer $\delta(t)$; therefore, the received signal, in the absence of noise, is no longer identical to the transmitted signal. There are several publications in modeling of the channel impulse response for the UWB channel. We have covered two of them in a previous Chapter. For the
analysis of the simple receiver performance in UWB channel, we will use the Saleh-Valenzuela model which is accepted widely to be accurate in modeling the UWB channel.

Since we are using a statistically modeled UWB channel, we cannot use mathematical analysis optimize parameters, hence cannot mathematically calculate the error performance. Therefore, we are using numerical methods to optimize the filter parameters and to evaluate receiver performance. For all the analysis done in this section, we use the following parameters for UWB Channel Modeling: $\alpha = 3$, which represents open space office environment with few multi-path creating surfaces. Ray arrival rate $\lambda = 2.1\text{ns}$, Cluster arrival rate $\Lambda = 0.0667\text{ns}$, ray power decal constant $\gamma_r = 7.9 \times 10^{-9}$, cluster power decay constant $\Gamma_C = 14 \times 10^{-9}$. We set the transmission distance to $3m$. To maintain simplicity we limit the receiver filter to second order butterworth filter.

5.7.1 Parameter Optimization for Simple Receivers in UWB Channels

First, similar to the analysis of AWGN and AGGN Channels, we investigate the effect of receiver filter on the received signal. We investigate the percentage of the signal energy that passes through the receiver filter for various filter bandwidths. Fig. 5.10 shows the percentage of the captured energy for the received signal when the signal passes through the receiver filter in the absence of noise.
Figure 5.10: Percentage of the Captured Energy of at the receiver output for different filter bandwidths ($W$)

From Fig. 5.10, we see that the Half-Cosine and Rectangular Pulses both have a significant energy passing through the filter at low receiver bandwidths. In order to find optimal filter bandwidth, we have to investigate the signal-to-noise ratio of these pulse candidates for different filter bandwidths. Fig. 5.11 shows the signal-to-noise ratio of four pulse candidates at various filter bandwidths. From this analysis, we conclude that the rectangular pulse as the highest signal-to-noise ratio among all the pulses at $W = 1/T_p$, however, the half-cosine pulse’s signal-to-noise ratio at $W = 1/T_p$ is comparable to the rectangular pulse. We note this distinction because the rectangular pulse of a nanosecond duration may not be realizable in real world applications. It is likely that in real world applications, the rectangular pulse of one nanosecond duration will have some rise and fall round-off and the pulse will resemble a half-cosine pulse.
Figure 5.11: Signal-To-Noise Ratio for pulse candidates at the receiver output for different filter bandwidths ($W$)

The last parameter we have to optimize is the receiver power $\beta$. For this optimization we will only analyze the rectangular and half-cosine pulse at $W = 1/T_p$ since these are the optimized parameters from our previous analysis. Fig 5.12 shows the probability of bit error for values of $\beta = [0.5, 1, 2, 4]$ for the rectangular pulse and half-cosine pulse at $W = 1/T_p$. As a result, we observe that both pulse candidates have their best error performance when receiver power $\beta = 1$. This is a desirable result since the absolute value operation $|.|$ is a simple low-cost operation when applied in circuitry.
Figure 5.12: Probability of Bit Error for two pulse candidates with $W = 1/T_p$ vs receiver Power $\beta$

5.7.2 Error Performance in UWB Channels

In the previous section we have optimized the receiver parameters for best error performance in UWB channels. In this section, we apply these parameters and obtain error performance of simple receivers in UWB channels. For this analysis we normalize the received signal energy at the receiver. The threshold calculation for simple receivers in UWB channels depends on the signal-to-noise ratio, furthermore, the probability density for the received energy when "0" and "1" is transmitted are different. For the match filter receivers the threshold is simply the average of the means of the probability density functions; however, due to the differences of shape and variances of the two distributions, this argument is not true for simple receivers. In UWB Channels, we numerically calculate the optimal threshold
for the simple receiver with previously given parameters. Furthermore, we use curve fitting method to come up with an equation to simplify the calculation of decision threshold.

\[
10 \log \left( \frac{E_b}{T_{h_{UWB}}} \right) = 0.44 \left( 10 \log \left( \frac{E_b}{N_0} \right) \right) + 10 \log (E_b) + 31
\]  

(5.7.1)

Fig. 5.13 shows the bit error probability of simple receivers using rectangular pulse and the half-cosine pulse. These calculations are performed using monte carlo simulation over 2000 statistically generated UWB channels. In the previous section we concluded that the performance of half-cosine pulse and rectangular pulse are close and comparable, therefore, we include the error performance of both pulses in Fig. 5.13. We can say that the rectangular pulse represents a theoretical optimal pulse.

In conclusion, the simple receivers have 2.5dB performance degradation over the optimal OOK receiver. Considering the complexity of the optimal receiver, and the fact that although most UWB channels are modeled as time-invariant, they in fact change slowly over time, the simple receiver is a very strong candidate over the complex alternatives. For most applications 2.5dB degradation is an acceptable loss considering the gain on cost of the receiver.
Figure 5.13: Probability of Bit Error of Simple receivers using Rectangular Pulse and Half–Cosine Pulse with $W = 1/T_p$ and $\beta = 1$ compared to the optimal OOK performance.
CHAPTER 6
SUMMARY AND FUTURE WORK

6.1 Summary

The need for higher data rates in communications is growing by day. Naturally, high theoretical data rate of UWB systems makes it a strong candidate for short range indoor applications. Although it has many advantages, UWB systems have challenges.

In this dissertation, firstly, the definition and history of UWB is introduced. The advantages, challenges and competition of UWB in the WPAN market is stated. In the second chapter, the background of some concepts involving UWB is explained. These concepts include: UWB with orthogonal frequency division multiplexing (MB-UWB), spread spectrum and UWB with direct sequence spread spectrum (DS-UWB), on-off keying, pulse position modulation, time division multiple access, traditional UWB pulses and the error performance of OOK and PPM in additive white gaussian noise.

The third chapter evaluates and describes implementation of two statistical UWB channel models for experimenting in simulation. The two models investigated are autoregressive AT&T channel model [36] and widely accepted Saleh-Valenzuela [32]. Methods for generating UWB simulation channels are discussed and performance of simulated channels are evaluated in this chapter.
In chapter four, pulse waveform shapes are optimized to minimize bit error rate of pulse position modulated UWB systems. A proof of optimality of these pulses are shown, and a method to calculate the optimal pulse shapes for any modulation index are suggested. Furthermore, aspects of the optimal pulses are discussed. It is shown that pulses suggested by Motorola and supported by UWB forum for standardization of 802.15.3a communications is a close estimation of the optimal pulse.

In chapter five, we suggest several simple, low-cost receivers for OOK UWB that perform below optimum. The receiver parameters are optimized to achieve best performance. The best performance obtained with these parameters are shown to be comparable to the optimal performance, hence, making these receivers simple, low-cost alternatives to the complex optimal receivers.

### 6.2 Future Works

1. The coefficients calculated and derived in PPM pulse shape optimization have important significance. They result in the minimum value of the autocorrelation of any pulse shape. A further study to investigate the performance of a pulse train with amplitudes corresponding to such coefficients needs to be done.

2. A real life implementation of simple, low-cost optimized receiver should be constructed and evaluated with narrowband pulses and UWB pulses. The performance should be compared to theoretical findings.
3. In most UWB simulations the channel is assumed to be time invariant. In reality, the channel is slowly changing with time. An interpolation of channel variance could be implemented in simulations to improve accuracy of channel model.

4. All of the experiments and findings presented in this dissertation applies to DS-UWB. A collection of experiments and studies in MB-OFDM could prove to be useful in order to cover UWB completely.
LIST OF REFERENCES


115


