Performance of UWB Receivers Using Stored Channel Responses

By

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Abstract

Ultra Wideband (UWB) technology possesses several characteristics that are required for the next generation Wireless Personal Area Network (WPAN) applications: simple implementation, high data rate, coexistence with existing systems. Furthermore, due to its wideband nature, UWB signals have higher penetration capability and better performance in indoor multipath environment.

This thesis presents performance evaluations of the Multiband OFDM (MB-OFDM) UWB system based on simulations using the indoor UWB channel measurements collected by Yingzhi Chen in [1]. Comparison of system performance (in terms of bit error rate) using different system parameter settings is presented. The parameters considered include Cyclic Prefix vs. Zero-Padded Suffix being used in the OFDM transmitter and receiver, length of the cyclic prefix, different combinations of modulation, coding and interleaving allowed for data rate of 200 Mbps vs. 53.3 Mbps.

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Acronyms

BER	Bit Error Rate
CE	Channel Estimation
CP	Cyclic Prefix
dB	Decibel
ECMA	European Computer Manufacturers Association
EIRP	Effective Isotropic Radiated Power
FCC	Federal Communication Commission
FFT / IFFT	Fast Fourier Transform / Inverse Fast Fourier Transform
IDFT	Inverse Discrete Fourier Transform
IEEE	The Institute of Electrical and Electronics Engineers
LOS	Line of sight
MAC	Media Access Control Layer
MB-OFDM	Multiband Orthogonal Frequency Division Multiplexing

NLOS	Non line of sight
OFDM	Orthogonal Frequency Division Multiplexing
PHY	Physical Layer
PLCP	Physical Layer Convergence Protocol
PPDU	PLCP Protocol Data Unit
PSD	Power Spectral Density
PSDU	PLCP Service Data Unit
QPSK	Quadrature Phase Shift Keying
SNR	Signal-to-Noise Ratio
T-R	Transmitter-Receiver
UWB	Ultra Wideband
VNA	Vector Network Analyzer
WPAN	Wireless Personal Area Network
ZPS	Zero-Padded Suffix

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Chapter 1 Introduction

This chapter first presents the objective of this thesis and an overview of findings. Then, existing literatures examining the UWB system performance under various wireless environments are reviewed. Finally, the organisation of this thesis is given.

1.1 Thesis Objective

Over the past several years, the interest in Ultra Wideband (UWB) technology has been continuously increasing since its exposure to the public. UWB can achieve high data rates by utilizing the wide bandwidth made available, and has become one of the most favourable indoor short-range wireless technologies in development for Wireless Personal Area Network (WPAN) applications. As of today, the industry is divided into two separate groups of supporters for different UWB proposals: Multiband Orthogonal Frequency Division Multiplexing (MB-OFDM) and Direct-Sequence UWB (DS-UWB). In this thesis, MB-OFDM UWB technology will be studied. While DS-UWB keeps the original UWB pulsed nature by sending a short pulse utilizing all the available bandwidth, MB-OFDM UWB divides the whole 7.1G bandwidth into frequency bands of 528 MHz each and uses OFDM modulation in each frequency band. Each of the two proposals was being supported by a different group of companies.

As a strong candidate for the physical layer of the next generation WPAN, it is important to thoroughly examine the UWB channel characteristics and obtain a better

understanding of its system performance in different environments. The main question to be considered in this thesis is:

What is the performance of MB-OFDM based UWB receivers in a realistic indoor environment?

This thesis is intended to help answer that question by providing various performance analyses of MB-OFDM UWB receivers using stored indoor channel impulse responses measured in a typical lab room at Carleton University by Yingzhi Chen (as presented in [1]). A Matlab simulation model was built for this thesis based on the MB-OFDM UWB design proposed to IEEE task group 802.15.3a [2], which later became a European Computer Manufacturers Association (ECMA) standard [3]. Since indoor channels are comparatively static, measured channel statistics can be used to help investigate the performance of UWB receivers in the field.

The main contribution of this thesis is to provide additional views of UWB indoor channel characteristics and to help develop a better understanding of the performance of MB-OFDM UWB system in those indoor multipath environments. Performance comparison using different system parameter settings such as Cyclic Prefix length and data rate configurations is also performed.

1.2 Overview of Findings

In this thesis, the system performance in terms of bit error rate (BER) in different indoor environments is examined using a Matlab model created to utilize the measured indoor channel responses from [1] and simulate the MB-OFDM UWB physical layer processing. In general, three sets of channel response measurements are used in this thesis: Line-of-Sight (LOS), Non-Line-of-Sight (NLOS) in the same room behind bookshelves, and NLOS in the hallway behind the wall. System configuration for data rate of 200 Mbps (highest mandatory data rate) is used in most simulations, while data rate of 53.3 Mbps (lowest mandatory data rate) is also examined for comparison purpose [3].

First of all, with no added noise, the resulting BER using Cyclic Prefix (CP) is compared with BER using Zero-Padded Suffix (ZPS) instead in the OFDM transmitter and receiver. An overlap-and-add algorithm is used in the receiver for ZPS processing. Slightly better performance (i.e. lower BER) is observed with CP in most scenarios. For that reason, all other simulations in this thesis use CP.

Secondly, the system performance degradation with additive Gaussian noise is examined. Plots of BER as a function of transmitter-receiver (T-R) distance for each set of channel measurements are presented. The path loss assessment based on channel measurements in [1] is used to derive the amount of noise added at various T-R distances in different channel environments. The findings indicate that BER values can sometimes approach 50% in NLOS environments, which appears to be very difficult to operate with such a high BER at data rate of 200 Mbps with CP length of 32 samples (i.e. 60.6 ns). When

the CP length is increased from the default of 32 samples to 96 samples (i.e. 181.8 ns), BER decreases slightly for the lower T-R distance NLOS channels (i.e. less than 4 meters). As an attempt to further lower the BER, simulations using a lower data rate of 53.3 Mbps are performed, which uses higher coding rate, higher overall spreading gain, and hence better system performance are to be expected and is confirmed by the simulation results.

Third, the system performance with no added noise using three variations of Cyclic Prefix (CP) length is evaluated: 32 (default), 64 (2x default) and 96 (3x default) samples. The simulation results show that BER is much lower when longer CP length is used.

Fourth, system performance comparison is made using modulation, coding and interleaving combinations allowed for 200 Mbps and for 53.3 Mbps. CP length of 32 samples is used. In addition, results were compared with data extracted from several existing publications, which helped to further validate the simulation results in this thesis. The simulation results quantify the extent of performance improvement by using a lower data rate configuration instead of using the highest mandatory data rate of 200 Mbps.

Finally, BER vs. SNR plots for different T-R distances at 200 Mbps data rate with CP length of 96 samples (i.e. 3x default) are generated, to show the relative improvement as compared with performance using default CP length.

1.3 Existing Literatures Review

In recent years, there have been many publications evaluating the performance of MB-OFDM based UWB with different indoor channel environments, which are believed to be the most common communication application environment for UWB. In general, most existing studies used the IEEE UWB channel models CM1 – 4 [4], which were proposed by a subgroup in IEEE 802.15.3a for WPAN applications. It offers a common channel model among different performance studies and proposals, and captures the path loss and multipath characteristics of a typical environment where UWB will be used. After comparing various channel models and proposals with measurements, the IEEE subgroup found Saleh-Valenzuela (S-V) model to be the best fit with the measured UWB indoor channel characteristics. The S-V model is a statistical description of an indoor channel environment based on measurements done by Adel A. M. Saleh and Reinaldo A. Valenzuela in 1987 [5]. The final models proposed (i.e. CM1-4) were actually the S-V model with some modifications after comparing with some UWB channel measurements [4]. CM1 and CM2 model the LOS and NLOS scenario respectively in a typical residential area with size comparable to a small office room. On the other hand, CM3 and CM4 characterize the NLOS scenario in a typical office or laboratory setup with higher number of multipaths.

The performance analysis of the MB-OFDM UWB using IEEE UWB channel models CM1, CM2, CM3, and CM4 at various data rates (i.e. 55, 110, and 200 Mbps) and the corresponding BER vs. E_s/N_0 (symbol energy to noise ratio) plots can be found in [6].

BER vs. E_b/N_0 (bit energy to noise ratio) plot for 200 Mbps data rate with channel model CM2 is published in [7], while analysis on data rates of 55, 160, and 480 Mbps using channel models CM1 and CM4 are presented in [8].

In some publications such as [6], BER is presented as a function of either E_s/N_0 or E_b/N_0 . Hence, when comparing the results with the findings in this thesis, they need to be transformed according to the following relationship: [9]

$$\frac{E_s}{N_0} (dB) = \frac{E_b}{N_0} (dB) + 10 \log_{10}(k) \quad (1)$$

Where k is the number of information bits per symbol and is equal to the product of the convolutional code rate ($R = 5/8$) and the number of coded bits per modulated symbol (2 for QPSK). Hence, $k = (5/8)*(2) = 5/4$ for data rate of 200 Mbps in MB-OFDM UWB.

1.4 Thesis Organisation

Chapter 2 gives the history and an overview of UWB technologies. A general introduction of WPAN and OFDM technology will be given. In addition, some essential wireless radio channel characteristics are discussed. Chapter 3 provides the description and the pre-processing steps required for the channel response measurements used in this thesis. Plots of all the processed channel response measurements will also be included. Chapter 4 first provides a high level overview of the MB-OFDM UWB Matlab model

developed for this thesis, followed by the detailed description of each functional block in the model. Chapter 5 presents all the simulation results and analyses carried out for this thesis. Finally, conclusions and future work are given in Chapter 6.

Chapter 2 Background

First of all, this chapter gives a summary of the development history of Ultra-Wideband (UWB) technology and an overview of the Multiband Orthogonal Frequency Division Multiplexing (MB-OFDM) proposal. Then, a description of Wireless Personal Area Network (WPAN) and an introduction of the OFDM design are given. Furthermore, some essential indoor wireless radio channel properties are discussed.

2.1 UWB Technology Overview

2.1.1 History

UWB technology applies to the physical radio layer, which was first developed for military use due to its flexible, high penetrating capability, and difficult-to-intercept nature. The development of UWB theory started as a military application dating back from 1960s. The UWB technology was more generally known as the baseband, carrier-free or impulse radio technology [10]. Due to its low interceptability and fine ranging resolution ability, UWB technology grew its popularity in military and radar applications, and has been restricted to military programs such as highly secured communications until 1990s [11]. In 1989, the term “ultra wideband” was publicized by the U.S. Department of Defence, after the technology had been under development for almost 30 years. Since then, the interest in UWB from the commercial industry has continued to rise. On February 14, 2002, the Federal Communication Commission (FCC) approved the Report and Order to authorize the use of UWB in the unlicensed 3.1

GHz to 10.6 GHz frequency band (bandwidth of 7.5 GHz) for commercial purposes. The definition of UWB today, according to FCC, requires the occupied bandwidth to be at least 20% of the center frequency, or minimum of 500MHz [3]. The Effective Isotropic Radiated Power (EIRP) is limited to -41.3 dBm per MHz, which is the maximum power available from the transmitter. As shown in Figure 1, the transmit Power Spectral Density (PSD) mask limits the emissions level at 0 dBr (dB relative to the maximum spectral density of the signal) from -260 MHz to 260 MHz around the centre frequency, at -12 dBr at 285 MHz frequency offset, and at -20 dBr at 330 MHz frequency offset and above [3].

This transmit-PSD-mask is copied from the ECMA-368 standard [3], which describes the limitation on the relative emission power level in frequency domain for UWB signals.

Figure 1 Transmit power spectral density mask

The main advantages of UWB include its capability to:

- Offer higher channel capacity due to the large bandwidth as limited by the channel capacity theorem, which shows a linear relationship between bandwidth and maximum achievable capacity of a system [10]:

$$C = B \times \log_2 \left(1 + \frac{S}{N} \right) \quad (2)$$

Where:

C = maximum channel capacity (bits per second)

B = channel bandwidth (Hz)

S = signal power (Watts)

N = noise power (Watts)

- Increase robustness to multipath fading with shorter signal duration, which makes it easier to resolve all the received multipath signals, hence reduces fading commonly caused by random multipath reflections that are received out of phase for narrowband signals [1].
- Utilize wideband spectrum instead of specifically allocated frequency bands, which allows UWB to be deployed around the world with minimum concern on its interference on any particular frequency band, and helps to resolve the problem of non-sufficient frequency band available for allocation.
- Operate at a low power consumption level as limited by the regulations.

Since UWB can transmit data at high rates for near-distance wireless communications, it became good candidate for replacing the high-speed data cables and the next generation high performance Wireless Personal Area Network (WPAN) applications in homes and offices, such as wirelessly connecting desktop peripherals with personal devices, file transfer and streaming video [12]. In January 2003, a task group was formed by The Institute of Electrical and Electronics Engineers (IEEE), named 802.15.3a, to pursuit the standardization of UWB as an alternative physical layer design for WPAN high rate applications.

After numerous meetings and debates, two distinctive proposals remained: Multiband OFDM (MB-OFDM) and Direct-Sequence (DS) UWB. While DS-UWB keeps the original UWB pulsed nature by sending a short pulse utilizing all the available bandwidth, MB-OFDM UWB divides the whole 7.1G bandwidth into frequency bands of 528 MHz each and uses OFDM modulation in each frequency band.

In January 2006, the IEEE 802.15.3a committee still could not reach an agreement among the group, and supporters for each proposal finally voted to withdraw from the committee. Both opposing groups of companies have decided to carry on with their own designs and leave the decision to the consumer market. Supporters for MB-OFDM have formed an organization called Multiband OFDM Alliance (MBOA), which eventually became the WiMedia Alliance [12]. It is a global non-profit organization that promotes the standardization of UWB for high-speed wireless connectivity in personal devices. WiMedia also helps to foster an open forum for the industry to establish requirements and

specifications for MB-OFDM UWB. The specification was then later adopted as a European Computer Manufacturers Association (ECMA) Standard [3]. This Geneva-based non-profit industry association is dedicated to fast-tracking of standardization of information and communication systems.

2.1.2 Spectrum Allocation for MB-OFDM UWB

Since FCC issued the Report and Order in 2002, development on UWB did not take it as a particular modulation technique of short duration pulses, but rather the available spectrum. Based on this assumption, the idea of multiband UWB signals were being developed, which can efficiently utilize the spectrum by transmitting multiple UWB signals at the same time via different frequency slots. MB-OFDM UWB allows signals either all to be transmitted at the same time to achieve high data rate for one user, or to be transmitted independently to allow multiple-access with each user using different frequency band. This multiband approach makes UWB applications more adaptive and scalable. In addition, multi-band applications can satisfy region-dependent spectral emission masks by disabling certain bands and avoid unwanted interference to other licensed bands.

The entire allocated spectrum (3.1 – 10.6 GHz) is divided into 14 bands spaced at 528 MHz from each other, and are grouped into 5 band groups [13]. Band group #1 is proposed to be used in the initial deployment of UWB devices, while it is optional for

band groups #2 and #3 or they can be added in later versions.

Figure 2 Spectrum allocation for MB-OFDM UWB

2.1.3 Overview of MB-OFDM UWB Physical Layer

According to the ECMA-368 standard for UWB Physical layer (PHY) for WPAN, it is mandatory to support transmitting and receiving at data rates of 53.3, 106.7 and 200 Mbps. Furthermore, the standard defines the method of transmitting and receiving data wirelessly at the PHY layer and how it should interface with the MAC layer. In this thesis, the Physical Layer Convergence Protocol (PLCP) sublayer for both the lowest (53.3 Mbps) and the highest (200 Mbps) mandatory data rate will be investigated.

The PLCP sublayer converts a PLCP Service Data Unit (PSDU) into a PLCP Protocol Data Unit (PPDU) before sending to the MAC layer. Each PPDU consists of three main parts: PLCP preamble, PLCP header, and PSDU [3].

Figure 3 PLCP Protocol Data Unit (PPDU) component breakdown

The PLCP preamble is the first component to be sent, which includes the deterministic packet / frame synchronization sequence (24 symbols) and the channel estimation sequence (6 symbols) for packet detection and piconet¹ identification.

Following PLCP preamble, PLCP header (12 symbols) is inserted for every PPDU and contains PHY and MAC related information such as header check sequence and parity bits for proper error correction decoding at the receiver. Finally, the PSDU consists of the frame payload, frame check sequence, tail bits and pad bits.

There are two types of diversity spreading available for MB-OFDM UWB for enhancing the robustness of transmitted signals against frequency and time selective errors in the channel [14]:

- Frequency domain spreading – This is only available for data rates 53.3 and 80

¹ A piconet is an ad-hoc computer network linking a user group of devices using wireless technology protocols to allow one master device to interconnect with other devices. A piconet typically has range of about 10 meters.

Mbps. A spreading factor of 2 is achieved by using the conjugate symmetric input to IFFT, which maps 50 modulated data subcarriers symmetrically to 100 IFFT inputs of an OFDM symbol.

- Time domain spreading – This is available for data rates up to 200 Mbps. A spreading factor of 2 is achieved by sending the same data OFDM symbol twice consecutively.

The table below is extracted from ECMA-368 standard [3], listing the recommended diversity settings and the coding rate for different data rates.

Table 1 Data rate dependent PSDU parameters

Data Rate (Mbps)	Coding Rate (R)	Frequency Diversity Spread	Time Diversity Spread
53.3	1/3	YES	YES
80	1/2	YES	YES
106.7	1/3	NO	YES
160	1/2	NO	YES
200	5/8	NO	YES
320	1/2	NO	NO
400	5/8	NO	NO
480	3/4	NO	NO

2.2 Wireless Personal Area Network (WPAN)

A Personal Area Network (PAN) is a collection of personal devices that are

inter-connected as a network within short-range. PANs can be used for communicating among the personal devices themselves or for connecting to a higher level network and the Internet. A Wireless Personal Area Network (WPAN) connects wirelessly with network technologies such as Bluetooth and UWB. Typically it has a short transmission range of less than 10 meters.

A working group is formed in the Institute of Electrical and Electronics Engineers (IEEE) specializing in WPAN standard, named IEEE 802.15. The objective is to standardize the WPAN technologies such that different manufacturers can create interoperable networking equipment. Within the working group, there are a number of task groups working on various WPAN technologies standardization, while Task group #3 focuses on High Rate WPAN. The goal is to create a standard for the MAC and PHY of wirelessly-connected, high-speed, multimedia-capable consumer devices [15]. Three different projects were created under IEEE 802.15.3 Task Group [16]:

- *802.15.3a – WPAN High Rate Alternative PHY*

In 2003, UWB was proposed to be the alternative PHY design for high rate WPAN, and task group 3a was formed in an attempt to establish a standard for it. However, consensus could not be reached by the supporters of two remaining proposals (MB-OFDM UWB vs. DS-UWB), and the task group was dismissed in 2006. Development for the two proposals has been taking place in separate forums outside of IEEE since then.

- *802.15.3b – MAC Amendment*

This group was formed to improve implementation and interoperability of the existing MAC for existing 802.15.3 WPAN standard, which includes minor optimizations while preserving backward compatibility [16].

- *802.15.3c – WPAN Millimetre Wave Alternative PHY*

This group was formed in March 2005 to develop a millimetre-wave-based alternative PHY for the existing 802.15.3 WPAN standard. This mmWave WPAN will operate in the new and clear band including 57 – 64 GHz unlicensed band defined by FCC 47 CFR 15.255. It will allow high coexistence (close physical spacing) with all other microwave systems in the 802.15 family of WPANs. It will also support high data rate at least 1 Gbps applications such as high speed internet access and streaming content download [16].

An example of potential UWB future application is home theatre systems, where a set of wireless consumer electronic devices such as monitors and audio speakers are involved in a room. The estimated information data rate requirement is roughly 50 Mbps with high definition (HD) video system [17].

2.3 Orthogonal Frequency Division Multiplexing

In each frequency band for MB-OFDM UWB, the Orthogonal Frequency Division Multiplexing (OFDM) is used as the modulation scheme. OFDM is a multiple carrier modulation technique which can capture the multipath energy efficiently and help to

achieve high data rate in multipath environments of broadband network. The concept of OFDM was first patented by Chang in 1960's [18], when he was trying to develop a solution for the multipath problem without reducing the data rate. Basically, the bandwidth is divided into a number of sub-bands, where each sub-band is modulated individually with modulation scheme such as Phase Shift Keying (PSK). Therefore, parallel data streams can be transmitted concurrently on separate sub-carriers. Since the signal power is spread among a wide frequency range, interference to the wireless systems in all frequency bands is set to minimal. Other attractive properties of OFDM include its resistance to narrowband interference and spectral flexibility.

Figure 4 Constellation mapping diagram for QPSK

According to the ECMA standard for MB-OFDM [3], modulation scheme Quadrature Phase Shift Keying (QPSK) is used to modulate coded data bits in every sub-band (or subcarrier) for data rate from 53.3 Mbps to 200 Mbps. QPSK is a popular digital modulation scheme that transforms a signal into the phase of a reference signal and increases the resistance to noise during transmission. Each pair of data bits being modulated will become a QPSK constellation data symbol according to the fixed

constellation diagram as shown in Figure 4 [19]. For example, if the input coded bits are $\{1, 1\}$, the modulated signal will be $\{(1 + j) / \sqrt{2}\}$.

This diagram shows the constellation mapping for Dual-Carrier Modulation (DCM), which is used for high data rates MB-OFDM UWB (i.e. 320 – 480 Mbps [3]).

Figure 5 Constellation mapping diagram for DCM

Dual-Carrier Modulation (DCM) scheme is used instead of QPSK for higher data rates (i.e. 320, 400, and 480 Mbps), to improve transmission range by providing additional diversity without reducing the data rate. With DCM, every same group of 4 coded bits are mapped onto two different 16-point symbols using the two fixed mappings as shown in Figure 5, which are very similar to 16-QAM constellation but with different mapping. The two DCM symbols are then mapped onto two frequency subcarriers that are 50 subcarriers apart, to improve frequency diversity in order to combat frequency selective fading within the transmission bandwidth.

In theory, a bank of matched filters is needed for OFDM transceivers in order to transmit simultaneously at a number of frequency bands. In fact, those operations can also be achieved by using an Inverse Fast Fourier Transform (IFFT) and Fast Fourier Transform (FFT) operation, given the number of frequency bands can be represented as a power of two. This helps to reduce the complexity and the cost of the receivers. The IFFT operation basically transforms the signal from frequency domain to time domain at the transmitter, while the FFT operation does the reverse at the receiver.

Many existing wireless networks have already adopted the OFDM technology as their modulation scheme to enable higher data rates and better system performance. Examples include IEEE 802.11a/g Wireless Local Area Network (WLAN) standard and IEEE 802.16 standard for broadband wireless access system commonly known as WiMAX (Worldwide Interoperability for Microwave Access). OFDM is also chosen for the 4G cellular networks physical layer.

2.4 Indoor Wireless Radio Channels Properties

The objective of this chapter is to provide an overview of the important aspects of indoor wireless radio channels. In the following sub-chapters, three indoor wireless channel characteristics are discussed: path loss, multipath propagation, and impulse response.

2.4.1 Path Loss

When radio waves travel through a medium, attenuations to the signal causing large scale propagation losses are called path losses. The amount of signal power decay is dependent on the distance traveled, i.e. distance between the transmitter and the receiver. A considerable portion of the link budget is cost due to propagation loss [18].

The simplest way to predict the amount of pass loss is by assuming free space propagation. Free space loss is defined as the decay of signal power between two isotropic antennas in obstacle-free environment. An isotropic antenna is an ideal antenna which emits power uniformly in all directions. The free space power received can be described by the Friis free space equation [1]:

$$\frac{P_R}{P_T} = \frac{G_T G_R \lambda^2}{(4\pi)^2 d^2} \quad (3)$$

Where P_T , P_R is the transmitted and received power, and G_T , G_R is antenna gain² of the transmitter and receiver respectively. λ is the transmission wavelength and d is the link distance.

For severe multipath environments, field channel measurements are required to properly determine the path loss equation [18]. Since the power levels can have a large variation,

² Antenna gain is the amount of power increase introduced by the antenna at transmitter and receiver.

the path loss is often expressed in decibel (dB)³ form [1]:

$$PL(d) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) \quad (4)$$

Where d_0 is the reference distance used in measurements and d is the transmitter-receiver (T-R) distance. $PL(d)$ and $PL(d_0)$ are the path loss at distance d and d_0 respectively in dB, while n is the path loss exponent which specifies the amount of path loss increase depending on the d/d_0 ratio. The values of $PL(d_0)$ and n vary with different environments, and can be determined based on field measurements.

2.4.2 Multipath Propagation

Most broadband wireless services are required to operate in challenging NLOS conditions. The transmission of a wireless signal can be reflected by obstacles in the room such as walls and furniture, hence causing multiple delayed and weakened duplicates of the transmitted signal being detected at the receiver. Those signals are referred to as multipath signals. Due to the misalignment of their arrival times, those duplicates can be out-of-phase with the primary received signal. They can generate inter-symbol interference to the received signal, which could degrade the system

³ Decibel is the logarithmic expression of a value, and is commonly used for power. A value in dB is derived by $10 \cdot \log(x)$.

performance.⁴

A multipath channel is usually characterized by its power delay profile, which is represented by a plot of relative received power as a function of time delay relative to the first detectable signal arriving at the receiver [1].

Three essential multipath channel parameters are derived from the power delay profile:

- Mean excess delay – the first moment (i.e. weighted average) of all received multipath signals in the power delay profile.
- RMS (root-mean-square) delay spread – the square root of the second central moment of the power delay profile, which is also the standard deviation of the mean excess delay.
- Maximum excess delay – the time delay when the last multipath signal within a certain power threshold (relative to the most significant path) is received relative to the first received signal.

Indoor wireless transmission can be generalized into two categories: Line-Of-Sight (LOS) and Non Line-Of-Sight (NLOS). In LOS environments, since there is no obstruction between the transmitter and the receiver, the amplitude of the primary received signal is much larger than the multipath signals, therefore can be easily overcome. However, in NLOS environments, since all received signals are either reflected or diffracted (i.e.

⁴ The performance of wireless systems is commonly limited by the signal-to-noise ratio (SNR)

non-direct path), the amplitudes of the primary and multipath signals are more comparable, hence more difficult to differentiate and could cause higher interference level.

2.4.3 *Impulse Response*

From a high level perspective, an impulse response of a radio channel is the expected output at the receiver when an impulse signal is sent. An impulse is usually represented by the Dirac delta function $\delta(t)$, a pulse made in a very short period of time with infinitely large amplitude such that the total integral is equal to one. The complex impulse response $h(t)$ of a static time-invariant multipath channel can be modelled by the following equation [18]:

$$h(t) = \sum_{k=0}^{N-1} a_k e^{j\theta_k} \delta(t - \tau_k) \quad (5)$$

Where N is the number of multipath components received, a_k is the gain per path, θ_k is the phase shift per path, and τ_k is the path delay. The equation also reveals that an impulse response is simply the sum of the primary and all other multipath signals received. The Fourier transform of an impulse response is the frequency response $H(f)$ of the channel, and vice versa.

Chapter 3 UWB Channel Measurements

This chapter describes the channel measurement method and setup used in collecting the data as described in [1]. The link budget parameters derived based on the measurements are also discussed. Then, the pre-processing required to transform the channel measurements into proper format for simulations is explained. Lastly, plots of all processed channel response measurements are shown.

3.1 Channel Measurement Method Description

All measurements used in this thesis were extracted from [1]. Measurements were performed at Carleton University Minto Building room 4050, a typical lab/office environment in a campus building with big glass windows on two sides and metallic racks by the wall. There are a number of equipments on the rack and inside the room. The transmitting antenna was fixed in the same position while the receiver was moved to various locations both inside and outside of the lab room.

Three types of channel measurements were made:

1. Line of Sight (LOS), with receiver in the same room as the transmitter
2. Non Line of Sight (NLOS), with receiver in the same room behind bookshelves
3. Non Line of Sight (NLOS), with receiver in the hallway behind wall

The following map is copied from [1], showing the various locations of where the receiver was placed in Carleton University MINTO room 4050 for different sets of channel measurements.

Figure 6 Indoor locations of channel measurements

The frequency band measured was from 2 GHz to 5 GHz, with center frequency at 3.5 GHz and bandwidth of 3 GHz. This frequency band was chosen to be measured based on equipment limitation and the consideration to reduce measurement uncertainty [1].

The frequency domain channel sounding method was used to collect the channel impulse response measurements, with its system block diagram shown in Figure 7. A Vector Network Analyzer (VNA) scans a particular frequency band by stepping through discrete frequencies and the S-parameter test set measures the frequency response $S_{21}(\omega)$, which is the transfer function $H(\omega)$ of the channel. The channel transfer function $H(\omega)$ is then converted to the time domain $h(t)$ by using Inverse Discrete Fourier Transform (IDFT) processing.

The following map is copied from [1], showing the system block diagram for measurements using Frequency Domain Channel Measurement method.

Figure 7 Frequency Domain Channel Measurement System

The advantage of this channel measurement method is that it can measure both the amplitude and phase information of each individual multipath component. However, this measurement system requires careful calibration and a cable between receiver and VNA. Therefore, it is practical for very close measurements such as indoor channel sounding. Another limitation with this measurement technique is that scanning the whole frequency band will take some time so the measurement is not real-time. If the channel is a rapid changing time-variant channel, this system may give an erroneous impulse response. To mitigate this problem, the total sweep time interval was kept as small as possible [1].

Since the indoor environment being measured is relatively static, the measured response

is taken to be essentially time-invariant. Every measurement contains 801 received complex tones, with frequency step of 3.75 MHz, which translates to time domain duration of 267 ns. For each location, 10 channel response measurements were made at slightly different time instances. All impulse responses are being used in simulations, so that for every location there are 10 simulation results available. In total, there are 17 different locations inside and outside of the room being measured. Therefore, total (17 locations) x (10 impulse response measurements per location) = 170 impulse response measurements were examined in this thesis.

Table 2 Summary of Channel Measurements Used

Type of measurement		Distance between TX & RX	Number of sets of measurements
1	Line of Sight (LOS)	{1 – 7 meters}	7
2	Non Line of Sight (NLOS), with receiver in the same room	{2 – 7 meters}	6
3	Non Line of Sight (NLOS), with receiver in the hallway	{6.5 – 9.7 meters}	4
<i>Total number of sets of measurements examined</i>			<i>17</i>

3.2 Link Budget Parameters for the Measured Channels

The link budget for the UWB radio channel can be determined by finding out the total gains and losses of the signal power through the transmission path. Unity antenna gain

is assumed at the transmitter and the receiver for this thesis, which means 0 dB with no increase or decrease to the signal power. On the other hand, noise figure (N_F) is the degradation of the signal-to-noise ratio due to certain RF components along the transmission path, and can vary with different receivers. In this thesis, N_F of 6 dB is assumed [2].

The FCC Report and Order for UWB sets a limit on the Effective Isotropic Radiated Power (EIRP) to -41.3 dBm per MHz. Given that the bandwidth (B) of MB-OFDM UWB is 528 MHz, the maximum average transmitted power is:

$$P_T(dBm) = -41 + 10 * \log(B) = -13.8 dBm \quad (6)$$

While the average noise power per bit is generally known to be:

$$N_{avg}(dBm) = -174 + 10 * \log(B) = -86.8 dBm \quad (7)$$

Another major component in a link budget is the path loss, which constitutes a significant part of the signal power loss during transmission. The values of path loss for the measured LOS and NLOS channels are determined in [1] as follows (in dB):

$$\begin{aligned} \text{LOS: } PL(d) &= 43.654 + 10 * 1.70 * \log\left(\frac{d}{d_0}\right) \\ \text{NLOS: } PL(d) &= 44.791 + 10 * 3.17 * \log\left(\frac{d}{d_0}\right) \end{aligned} \quad (8)$$

Where d is the T-R distance and d_0 is the reference distance of 1 meter. Hence, the amount of additive noise (SNR) can be estimated by the following equation:

$$\text{channel dependent additive noise (SNR)} = P_T (\text{dBm}) - N_{\text{avg}} (\text{dBm}) - N_F (\text{dB}) - PL (\text{dB}) \quad (9)$$

3.3 Pre-processing of Channel Measurement for simulations

In order to convolve the measured channel impulse response with the transmitted signal in simulations, the following pre-processing is required on each set of measurements before being input to the simulations. For demonstration purposes, a LOS measurement data at T-R distance of 7 meters is shown below.

1. Using the measured magnitude and phase to construct the complex channel impulse response in frequency domain

(Note: All raw measurement files contain channel magnitude in dB, and phase in degree, which need to be converted to real values and radian respectively).

2. Perform IFFT to obtain the time-domain channel response.

3. Interpolate the sampling rate from $\{T_s = 267 \text{ ns} / 801 \text{ samples} = 0.333 \text{ ns}\}$ to $\{T_s = 312.5 \text{ ns} / 165 \text{ samples} = 1.894 \text{ ns}\}$ to align with the sampling rate in simulations.

4. Pad with zeros at the end to match the length of transmitted signal to be

convolved with in simulations, which is 165-sample.

3.4 Plots of Channel Impulse Response Measurements

In this chapter, some of the measured channel responses used in this thesis are shown. Pre-processing to convert the measured data into time domain impulse responses for our simulations is performed on all measurements (as described in Chapter 3.3). While there are total of 17 different channels and 10 measurements available per channel, Figure 8 shows the increase in multipaths when T-R distance is increased in each of the three measured channel types:

- LOS: 3 & 6 meters
- NLOS same room behind bookshelves: 3 & 6 meters
- NLOS in hallway behind wall: 6.5 & 9.7 meters

LOS – 3 meter **LOS – 6 meter** **NLOS behind bookshelves – 3 meter** **NLOS behind bookshelves – 6 meter** **NLOS in hallway behind wall – 6.5 meter** **NLOS in hallway behind wall – 9.7 meter**

Figure 8 Samples of channel impulse response measurements

Chapter 4 MB-OFDM UWB System Modeling Details

In this chapter, an overview of the MB-OFDM UWB model built for this thesis is given. Then, a detailed breakdown of each functional block in the model is provided. Finally, an example of simulation results is shown.

4.1 MB-OFDM UWB Matlab Model Overview

For this thesis, we have simulated the MB-OFDM UWB physical layer transmission as proposed in IEEE 802.15 Task Group 3a [2] and later defined in EMCA-368 standard [3]. Figure 9 illustrates the high-level block diagram of the simulation model⁵.

Figure 9 Matlab UWB model high-level block diagram

⁵ Matlab model [20] was used as the reference in building this model.

Our simulation model incorporates the measured channel responses and determines the corresponding BER with various system parameter settings. The flow of the simulation is described below:

- (1) Simulations start with generation of random data bits.
- (2) Data bits are sent to the convolutional encoder of rate $R = 1/3$; the encoded bits will go through puncturing where certain bits are discarded based on a predefined pattern in order to achieve $R = 5/8$ for 200 Mbps data rate.
- (3) Encoded bits are grouped and interleaved.
- (4) Every two interleaved bits are modulated to become one QPSK constellation data symbol.
- (5) OFDM modulation is applied through a 128-point IFFT, then Cyclic prefix (CP) or Zero-Padded Suffix (ZPS) and guard band are added to the signal.
- (6) Additive White Gaussian Noise (AWGN) is added to the transmitted signal when needed.
- (7) A 128-point FFT and channel estimation are performed inside the OFDM receiver block.
- (8) QPSK demodulation is done to retrieve the coded data bits.
- (9) De-interleaving is done accordingly.
- (10) Null bits are inserted into the data bits based on the same predefined pattern as used in puncturing procedure at the transmitter side for 200 Mbps data rate. Then, Viterbi decoding is used to recover the original data bits.
- (11) Finally, the average bit error rate is determined by comparing those retrieved data bits with their corresponding values at the input of the transmitter and

recording the total number of data bits and error bits received.

Matlab is a high-level programming tool for numerical computation and data analysis. Simulink is a Matlab-based graphical simulation tool that provides a block-diagram environment for developing simulations and a library of built-in functional blocks for signal processing and communications analyses [19].

In this thesis, the system configurations for both the lowest (53.3 Mbps) and the highest mandatory data rate (200 Mbps) are examined. In addition, since perfect synchronization is assumed, no packet/frame synchronization sequence will be included in the PLCP preamble. Therefore, preamble only contains the channel estimation sequence. In addition, the PLCP header is not included since the interaction between the PHY and MAC layer is out of the scope of this thesis.

The following chapters will provide the details in each functional block in the simulation model.

4.2 Random Bits Generator

To simulate the information bits being transmitted, random binary numbers are generated using Bernoulli distribution, with 50% probability of generating a zero (i.e. equal chance of sending either a one or a zero). Groups of binary bits are generated periodically at a fixed time interval.

The number of bits being generated periodically is equivalent to the amount of information bits required for an OFDM symbol at the transmitter output. For example, with 200 Mbps data rate, 125 information bits are generated, which will be transformed into a single OFDM symbol.

Figure 10 Transform of data bits in the transmitter for 200 Mbps

The goal is to achieve a sampling frequency of 528 MHz at the output of the transmitter. Hence, the sending interval at the bits generator has to include all overhead cost in time that could affect the bandwidth efficiency. There are three kinds of overhead cost that need to be accounted for:

- Convolutional encoding: 8 coded bits are generated for every 5 information bits sent
- In each OFDM signal, there are 65 overhead tones out of the total 165 tones transmitted: 12 Pilots, 10 Guard subcarriers, 5 zero padding subcarriers, 32 Cyclic

Prefix samples, and 5 guard band samples.

- In the OFDM transmitter, 6 CE sequences are sent in additional to the data symbols in time domain per PSDU frame. The total number of data symbols (after time diversity spreading) per PSDU is defined by: [3]

$$\#symbols\ per\ PSDU = 6 * \left(\frac{8 * LENGTH + 38}{N_{IBP6S}} \right) \quad (10)$$

Where *LENGTH* is the number of bytes in payload, N_{IBP6S} is the number of information bits per six “on-air” OFDM symbols since PHY defines an interleave depth of 6 symbols, and the value 38 represents the number of bits of frame check sequence and tail bits section when length of payload is non-zero [3]. In our simulations with 200 Mbps data rate, payload size of 1024 octets⁶ is used, and N_{IBP6S} is equal to $3 * 125 = 375$. This is because every 125 information bits are encoded and modulated into a single OFDM symbol, which will be sent twice with 2x time diversity. Therefore, there are 375 information bits per 6 “on-air” OFDM symbols. Hence, 132 symbols are being sent per PSDU frame:

$$\#symbols\ per\ PSDU = 6 * \left(\frac{8 * 1024 + 38}{375} \right) = 132 \quad (11)$$

Therefore, for simulations of 200 Mbps data rate, information bits are being generated

⁶ The payload size of 1024 octets was chosen because it was used in a number of published studies.

every 5.23 ns as derived below:

bit sending interval at the generator

$$\begin{aligned}
 &= \left\{ \frac{1}{BW} \right\} * \left\{ \frac{\# \text{ coded bits}}{\# \text{ data bits}} \right\} * \left\{ \frac{\# \text{ data} + \text{overhead tones}}{\# \text{ data tones}} \right\} * \left\{ \frac{\left(\frac{\# \text{ data}}{\text{symbols}} \right) + \left(\frac{CE}{\text{symbols}} \right)}{\left(\frac{\# \text{ data}}{\text{symbol}} \right)} \right\} \quad (12) \\
 &= \left\{ \frac{1}{528 \text{ MHz}} \right\} * \left\{ \frac{1}{5/8} \right\} * \left\{ \frac{165}{100} \right\} * \left\{ \frac{132 + 6}{132} \right\} \\
 &= 5.23 \text{ ns}
 \end{aligned}$$

4.3 Convolutional Encoding with Puncturing

Convolutional encoding is a type of error correction method that transforms every k input bits into a symbol of n coded bits. The code rate is recognized as $R = k/n$, where $n \geq k$. Also, the encoded output symbol will be dependent on the last K symbols, which is known as the constraint length of the code. The convolutional encoder for MB-OFDM UWB applications has code rate of $R = 1/3$, and consists of a shift register of length $K = 7$ (called the constraint length) [2]. Another parameter used to characterize the convolutional encoder is the free distance, which is defined as the minimum Hamming distance between any two encoded sequences. It represents the error-correcting capability of a Trellis code. The convolutional encoder used in this study has free distance of 15, which is determined using Matlab based on the known constraint length of 7 and the generator polynomials of [133 165 171].

For data rate of 200 Mbps, code rate of $R = 5/8$ is wanted, which can be generated from

the $R = 1/3$ convolutional encoder output by puncturing [2]. Puncturing is a procedure that removes some of the encoded bits at the transmitter side (i.e. reducing the number of transmitted bits) to achieve the desired number of output bits. In the simulations, the puncture vector of [1 1 0 0 0 1 1 1 0 0 0 1 1 1 0] is used according to [3], where “0” represents bit to be removed. Therefore, 7 of the 15 coded bits from the $R = 1/3$ convolutional encoder will be removed. For data rate of 53.3 Mbps, puncture vector with all 1’s is used to achieve $R = 1/3$ coding rate. That is, no coded bit will be removed.

Figure 11 Demonstration of encoding and puncturing for 200 Mbps

4.4 Interleaving and Modulation

The coded bits output from the encoder are then interleaved to provide robustness against burst errors. An interleaver permutes symbols according to a known pattern, and is used

to protect the data transmission against burst errors [19]. A corresponding deinterleaver in the receiver uses the inverse mapping to restore the original sequence of symbols. Three stages for bit interleaving are performed [3]:

Figure 12 Three stages of bits interleaving

1. **Inter-symbol interleaving** – This is done by first grouping coded bits for 6 “on-air” OFDM symbols, and permuting the grouped bits using a block interleaver of size {Row x Column}, which denotes the number of rows and columns of the block, and is dependent on the data rate configured.

Figure 13 Stage 1 Bit interleaving: Inter-symbol interleaving for 200 Mbps

For 200 Mbps, size {200 x 3} is used, as demonstrated in Figure 13. The

interleaver will fill the input symbols row by row first, and output the interleaved symbols column by column. This provides frequency diversity within the band group [3].

2. **Intra-symbol interleaving** – Similar to Inter-symbol interleaving, intra-symbol interleaving permutes coded bits in groups of an OFDM symbol by a block interleaver of size {Row x Column}, which is also data rate dependent. For 200 Mbps, the block size is {20 x 10}, as demonstrated in Figure 14. This provides frequency diversity across data tones within an OFDM symbol and increases the robustness against narrowband interferers [3].

Figure 14 Stage 2 Bit interleaving: Intra-symbol interleaving for 200 Mbps

3. **Cyclic shifts** – At the last stage of interleaving, coded bits that belong to 6 “on-air” OFDM symbols are grouped and cyclically shifted by deterministic amounts, as demonstrated in Figure 15. This will help to better exploit frequency diversity [3].

Figure 15 Stage 3 Bit interleaving: Cyclic shifter for 200 Mbps

After 3 stages of interleaving, each pair of data bits is modulated with Quadrature Phase Shift Keying (QPSK) method, where serial input of bits is divided into groups of two and then converted to complex values according to QPSK constellation as shown in Figure 4 in Chapter 2.1.3. Since QPSK modulates 2 bits per symbol, 200 coded bits will be modulated into 100 QPSK symbols.

4.5 OFDM Transmitter

The OFDM modulation is applied through a 128-point Inverse Fast Fourier transform (IFFT), with the 128 subcarriers distribution as shown in Figure 16:

- 100 Data subcarriers – Each data subcarrier contains a QPSK constellation data symbol.
 - 12 Pilot subcarriers – Inserted at pre-defined locations among the data subcarriers for robust coherent detection against frequency offsets and phase noise [14].
- Note that in our simulations, information contained in the pilot subcarriers will

not be processed or used.

- 10 Guard subcarriers – Copy of the first 5 and last 5 data carriers, placed at the beginning and at the end respectively. These guard subcarriers are used to mitigate inter-symbol interference (ISI). It is also optional to reduce the power of the guard subcarriers in order to relax the specifications of transmit and receive filters and possible performance improvements [3].
- 1 Null subcarrier – Also known as the Direct Current (DC) subcarrier, located at the center frequency (i.e. 0th subcarrier). It is not used (or is set to null value) in OFDM PHY layers in order to simplify the Digital-to-Analogue and Analogue-to-Digital Converter operations in the RF system [2].
- 5 Zero padding subcarriers – Padded at the end of the symbol to make up 128 carriers for the 128-point IFFT

Figure 16 Frequency subcarriers distribution per OFDM symbol

IFFT is then applied on the frequency subcarriers input and transform the signal into time

domain OFDM symbol with duration of 242.4 ns (i.e. 128 subcarriers / 528 MHz).

Afterwards, Cyclic Prefix (CP) is added to the beginning of the symbol by copying the last 60.6 ns of the signal. The CP converts the linear convolution with channel response into a cyclic convolution, where a cyclic convolution in time domain is equivalent to a scalar multiplication in frequency domain. Hence, most conventional wireless OFDM systems use CP to minimize inter-symbol interference (ISI) and to capture multipath energy efficiently.

Figure 17 OFDM symbol in time domain with CP

An alternative of using CP is to use Zero-Padded Suffix (ZPS), where the same amount of duration (i.e. 60.6 ns) of zero padding is added to the end of the OFDM symbol instead. ZPS is actually recommended in ECMA standard because studies have shown that the effect of using ZPS is similar to using CP, while ZPS can avoid the power back-off required if CP is used, which is intended for compensating the ripples caused to the transmitted power level [21]. In this thesis, system performance using CP and ZPS is investigated.

Finally, duration of 9.5 ns with null values are added at the end of the symbol as guard band to allow sufficient time for the transmitter and receiver to switch between the different center frequencies.

Figure 18 OFDM symbol in time domain with ZPS

For each transmission of PLCP Service Data Unit (PSDU) in the PHY layer, there are always 6 repeated Channel Estimation (CE) sequences sent at the beginning of the PSDU before sending the data symbols. The CE sequence is defined in the ECMA-368 standard and contains 122 identifiable QPSK constellation data bits values [3], which are known by both the transmitter and the receiver. It is used to estimate the channel frequency response at the receiver. At the transmitter, the 122 CE sequence QPSK constellation bits will be mapped onto the IFFT subcarriers in the same way as other data symbols: Null DC subcarrier at the center frequency and 5 zero padding subcarriers at the end to make up the 128 subcarriers needed for IFFT, then followed by CP at the beginning (or ZPS at the end) and guard band at the end of the signal.

Following the 6 repeated CE sequence symbols, 66 data symbols per PSDU are sent for

200 Mbps data rate (see Chapter 4.2 for details). With the time diversity spreading, each data symbol is sent twice, one after the other. Hence, $66 \times 2 = 132$ data symbols are sent per PSDU frame.

Figure 19 Composite of a PSDU frame

4.6 Convolution with Measured Channel Response

At this stage, each transmitted OFDM signal is convolved with the measured channel impulse response (as discussed in Chapter 3.3). This is done by the Matlab built-in convolution functional block, which mathematically convolve the input signal and the measured channel impulse response. Since both inputs have complex value elements, the output from convolution is also complex-valued.

4.7 Additive White Gaussian Noise

For some of our simulations where additional noise is introduced, Additive White Gaussian Noise (AWGN) is added to transmitted signal after convolving with the measured channel impulse response. Gaussian noise is commonly caused by natural sources such as thermal noise and radiation from the earth, which can generally be described statistically by a Gaussian distribution. The AWGN in simulations is generated by a Matlab built-in functional block, which linearly adds complex-valued white noise at a specified signal-to-noise (SNR) level, with a Gaussian distribution of amplitude, to the complex input signal [19].

4.8 OFDM Receiver

First of all, the guard band (i.e. the last 9.5 ns) for each received signal is removed. If Cyclic Prefix (CP) is used at the OFDM transmitter, the first 60.6 ns duration of the signal is removed.

On the other hand, if Zero-Padded Suffix (ZPS) is used, an overlap-and-add method is needed at the receiver before removing the last 60.6 ns duration of the signal, in order to obtain the circular convolution property [21].

After the prefix or suffix is removed, Fast Fourier transform (FFT) is performed to obtain the frequency domain response of the signal. Then, 5 zero padding subcarriers at the

end and the DC subcarrier in the center are removed, the original subcarrier settings with 100 data, 10 guard and 12 pilot subcarriers are now retrieved.

This diagram shows the ZPS specific processing required at the OFDM receiver (as compare to using CP). The operation is known as overlap-and-add. Also shown is signal duration at different stages along the path.

Figure 20 ZPS processing at OFDM receiver

A simple channel estimation and frequency domain equalization algorithm is used in our simulations. The equalization algorithm assumes that each frequency subcarrier will experience flat fading. Therefore, each subcarrier will have its own complex ratio determined using the expected and received CE sequence, to compensate the flat fading in magnitude and phase of the received data symbol.

As described in Chapter 4.5, 6 CE sequence symbols are sent at the beginning of every

PSDU frame. When they arrive at the receiver, the guard band and cyclic prefix are first removed and FFT is performed on each of them. After removing the Null DC and zero padding subcarriers, the average of those 6 received CE sequences is determined. The ratio of the expected CE sequence to the averaged received CE sequence is then calculated and multiplied to each of the succeeding received data OFDM symbols within the same PSDU frame. As a result, the original OFDM symbols with 122 frequency subcarriers are recovered.

After channel estimation, the 10 guard and 12 pilot subcarriers are removed with the 100 data subcarriers remaining. The time domain de-spreading is done by adding the succeeding repeated symbol to the OFDM symbol received, and then divided by two to obtain the average.

4.9 Demodulation and De-interleaving

The signal is then demodulated with QPSK according to the Gray mapping as discussed in the modulation chapter earlier.

The demodulated coded bits are then sent to 3 stages of de-interleaving, corresponding to the 3 stages of interleaving performed in the transmitter:

Figure 21 Three stages of bits deinterleaving

1. **Cyclic reverse shifts** – grouping bits that belong to 6 “on-air” OFDM symbols and reverse cyclic shifting

Figure 22 Stage 1 Bit deinterleaving: Cyclic reverse shifter for 200 Mbps

2. **Intra-symbol deinterleaving** – grouping bits that belong to the same OFDM symbol and performing block deinterleaving, with the same block size as in intra-symbol interleaver

Figure 23 Stage 2 Bit deinterleaving: Intra-symbol deinterleaving for 200 Mbps

3. **Inter-symbol deinterleaving** – grouping bits that belong to 6 “on-air” OFDM symbols and performing block deinterleaving, with the same block size as in inter-symbol interleaver

Figure 24 Stage 3 Bit deinterleaving: Inter-symbol deinterleaving for 200 Mbps

4.10 Viterbi Decoding

Before decoding, in order to invert the puncturing done at the transmitter, dummy zeros are inserted into the received coded bits at the receiver according to the puncture vector as defined in the standard [3].

To retrieve the data bits, Viterbi decoding algorithm is recommended by ECMA-368 standard for decoding symbols in the MB-OFDM UWB receiver [19]. Viterbi decoding was developed by Andrew J. Viterbi in 1967, and is commonly used for decoding the convolutionally encoded data bits. With code rate of $R = k / n$ used in the encoder, the corresponding Viterbi decoder will transform every n encoded bits input into k data bits output. In our simulations, $R = 1/3$ is used so that one data bit will be retrieved from every 3 encoded input bits in the receiver.

Figure 25 Demonstration of Viberbi decoding for 200 Mbps

The decision type used in the decoder is hard decision, which means that the decoder will

accept any real number as input. A positive real value input will be identified as a logical zero output, and a negative real value input as a logical one output.

4.11 Error Rate Calculation

The data bits output from the receiver (i.e. after leaving the Viterbi Decoder) will then be compared with its original value at the input of the transmitter (i.e. output from the random bit generator). This can be done by knowing the expected delay per information bit between the output of the bit generator and the output of the decoder in the model. As a result, the total number of information bits received and the total number of error bits are recorded. The bit error rate (BER) is basically the ratio of the number of error bits to the total number of bits received.

4.12 Example of simulation results

In this chapter, an example of simulation results using channel measurement of LOS at 7 meters is shown. Coding rate and interleaving combinations for data rate of 200 Mbps with Cyclic Prefix is used, with no Gaussian noise added.

Figure 26 A sample of channel measurements for location LOS – 7m

Before the simulation starts, each channel measurement data is first transformed into a time domain channel response as shown in Chapter 3.3. There are total 10 impulse response measurements available for location LOS at 7 meters, with one of them shown in Figure 26. Simulations were run using each of those 10 channel measurements.

Then, a Matlab script is used to import each of those 10 channel measurements for location LOS – 7m and run the simulation with the selected configurations (e.g. coding rate of 5/8, cyclic prefix in the OFDM transmitter and receiver). The simulation duration is set to 0.02 seconds of simulation time, which generates more than 3 million data bits. When simulations using all 10 channel impulse responses are completed, the total number of information data bits received and the number of error bits in each

simulation run are exported into a text file. An example of the recorded values is shown in Figure 27.

# of error bits	Total # of bits received	bit error rate
25503	3807716	0.67%
2509	3807716	0.07%
2441	3807716	0.06%
8945	3807716	0.23%
1476	3807716	0.04%
20528	3807716	0.54%
14226	3807716	0.37%
11222	3807716	0.29%
2886	3807716	0.08%
4145	3807716	0.11%

Figure 27 An example of simulation records for LOS-7m

The average, maximum and minimum bit error rate (BER) are then determined and plotted in graphs, with the maximum and minimum BER appear as the upper and lower point of the vertical bar on each data point as demonstrated below.

Figure 28 An example of BER result interpretation for LOS-7m

Chapter 5 Simulation Results and Analyses

This chapter presents all the simulation results and analyses obtained for this thesis.

Five groups of results and analyses are shown in this chapter:

- i. Compare the system performance when Cyclic Prefix (CP) is used in the OFDM transmitter and receiver, versus when Zero-Padded Suffix (ZPS) is used instead
- ii. Examine the system performance with the presence of additive Gaussian noise, which includes path loss based on measurements
- iii. Compare the performance using 3 variations of CP length (i.e. 32, 64, and 96 samples) in the OFDM transmitter and receiver
- iv. Compare the performance using different combinations of modulation, coding and interleaving allowed for data rate of 200 Mbps vs. 53.3 Mbps
- v. Investigate the sensitivity of system performance using 3x default CP length

There are 10 impulse response measurements available for each measurement location being examined, and all measurements were used in our simulations. For all the simulation results presented in this chapter, each BER value shown in the plots actually represent the average of 10 BER values resulting from the 10 simulations ran for the same measurement location. For each measurement location, the maximum and minimum of the 10 BER results are also shown in the plots as the upper and lower limit of the vertical error bar on each data point (as demonstrated in Figure 28 in Chapter 4.12). Three types of channel measurements are being examined: LOS, NLOS behind bookshelves, and NLOS behind wall.

5.1 Comparison of Performance with Cyclic Prefix vs. Zero-Padded Suffix

In ECMA-368 standard for MB-OFDM UWB [3], Zero-Padded Suffix (ZPS) in time domain OFDM symbol is recommended instead of Cyclic Prefix (CP), because the effect of using ZPS is expected to be the same as using CP. Also, using ZPS can help to reduce the RF transmission power required as compared to CP. The objective of this chapter is to examine the system performance difference between using CP and ZPS.

This plot shows the BER as a function of added noise (SNR) using CP vs. using ZPS in LOS channels. Data rate of 200 Mbps and CP or ZPS length of 32 samples was used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 29 Comparison of performance using CP vs. ZPS – LOS

Same length of 32 samples (or 60.6 ns) is used for both CP and ZPS. Combination of modulation, coding and interleaving for 200 Mbps is used. Figure 29 shows that the resulting BER using CP is relatively lower than using ZPS in LOS channels, hence better system performance. Results for NLOS behind bookshelves and NLOS behind wall channels are shown in Figure 30 and Figure 32 respectively. In general, the system performance is slightly improved with an average of approximately 27% relative decrease in BER when CP is used instead of ZPS.

This plot shows the BER as a function of added noise (SNR) using CP vs. using ZPS in NLOS channels in the same room behind bookshelves. Data rate of 200 Mbps and CP or ZPS length of 32 samples was used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 30 Comparison of Performance using CP vs. ZPS – NLOS in same room behind bookshelves

For NLOS behind bookshelves channels starting at T-R distance of 4 meters, the BER performance became much worse compare to T-R distance of 2-3 meters (as shown in Figure 30). One of the reasons is the increase in delay spread from T-R distance of 3 meters to 4 meters. This is also shown in the plots of impulse response measurements for those two locations (Figure 31). The measured multipath signals at 4 meters are more significant compared to the multipath signals at 3 meters.

Figure 31 Comparison of channel impulse response measurements for NLOS behind bookshelves at 3 meters and at 4 meters

Results for NLOS behind wall channels in Figure 32 showed some performance

improvement when CP is used instead of ZPS. Similar to NLOS behind bookshelves channels, the BER curve for 7.5 meters is much worse than the BER curve for 6.5 meters due to the increase in the measured delay spread.

This plot shows the BER as a function of added noise (SNR) using CP vs. using ZPS in NLOS channels in hallway behind wall. Data rate of 200 Mbps and CP or ZPS length of 32 samples was used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 32 Comparison of Performance using CP vs. ZPS – NLOS in hallway behind wall

It is observed that there are relatively less curvature for some of the BER curves of certain NLOS channels, where they appear to be reaching a plateau at higher SNR (e.g. NLOS behind wall 7.5 – 9.7 meters channels). In other words, the rate of BER decreasing slows down when SNR is further increases. One of the reasons is because

the Cyclic Prefix length used in those simulations is the default length of 32 samples, which is smaller than the maximum delay spread for some of the measured channels. Further discussion will be given in Chapter 5.3, where the performance sensitivity to CP length will be examined.

5.2 System Performance with Additive Gaussian Noise

The objective of this exercise is to examine the UWB MB-OFDM performance with added Gaussian noise to the transmitted signal. The amount of added noise takes into account the maximum allowed transmit power (as restricted by FCC), the T-R distance dependent path loss based on measurements [1], and the average noise power within the bandwidth.

Table 3 shows the link budget calculations used (as described in Chapter 3.2) for determining the noise level or SNR to be added for each measurement location.

The corresponding Path Loss and SNR determined for each measurement location are listed in Table 4. Notice that for NLOS behind wall channels, SNR values are negative. It means that the radio conditions for similar indoor channel environments are expected to be poor and difficult to operate at 200 Mbps.

Table 3 Link Budget Parameters

Information data rate (R_b)	200 Mbps
Bandwidth (B)	528 MHz
Total average transmitted power $P_T = -41 + 10 \cdot \log(B \text{ in MHz})$	-13.8 dBm
Average noise power within the bandwidth $N_{avg} = -174 + 10 \cdot \log(B \text{ in Hz})$	-86.8 dBm
Rx Noise Figure – antenna terminal [2] N_F	6.6 dB
Noise Power $P_N = N_{avg} + N_F$	-80.2 dBm
Free space path loss at d meters, derived in [1] based on measurements⁷ $PL(d) = PL(d_0) + 10n \log(d/d_0)$	<ul style="list-style-type: none"> • LOS: $PL(d_0) = 43.654 ; n = 1.70$ • NLOS: $PL(d_0) = 44.791 ; n = 3.17$
Noise to be added $SNR = P_T - PL(d) - P_N$	66.4 – PL(d)

⁷ d_0 is the reference T-R distance used in 0, and is chosen to be 1 meter.

Table 4 Additive noise for each measurement location

Distance (meter)	Path Loss (dB)	SNR (dB)
<i>LOS</i>		
1	43.7	22.7
2	48.8	17.6
3	51.8	14.6
4	53.9	12.5
5	55.5	10.9
6	56.9	9.5
7	58.0	8.4
<i>NLOS - same room</i>		
2	54.3	12.1
3	59.9	6.5
4	63.9	2.5
5	66.9	-0.5
6	69.4	-3.0
7	71.6	-5.2
<i>NLOS - in hallway</i>		
6.5	70.5	-4.1
7.5	72.5	-6.1
8.5	74.2	-7.8
9.7	76.0	-9.6

The simulation results of BER as a function of T-R distance with added noise is summarized in Figure 33.

The following diagram shows the BER as a function of T-R distance, with different amount of added noise (SNR) per measurement location according to Table 4. Data rate of 200 Mbps and CP length of 32 samples were configured. The error bars represent the minimum and maximum BER values from the 10 simulation results per location. Notice that BER values for some locations are not displayed in the graph, since their values are zeros and cannot be shown in this semi-log graph.

Figure 33 Performance with added noise for 200 Mbps with CP32

As demonstrated in Figure 33, BER is very high (close to 50%) in NLOS channels at 200 Mbps for T-R distance ≥ 5 meters. This is mostly due to the high noise level introduced (with negative SNR values) for those channels as listed in Table 4. As a result, two methods are attempted in this chapter to improve the performance in the measured NLOS environment:

Evaluation #1: Increase CP length from 32 to 96 samples – as shown in Figure 34, there are some improvements in the shorter distance range (T-R distance < 4 meters)

for NLOS channels. It is because longer CP length allows more multipath energy to be captured. However, the error rate is still too high for distance ≥ 5 meters.

The following diagram shows the BER as a function of T-R distance, with different amount of added noise (SNR) per measurement location according to Table 4. Data rate of 200 Mbps and CP length of 96 samples (3x default) were configured. The error bars represent the minimum and maximum BER values from the 10 simulation results per location. Notice that BER values for some locations are not displayed in the graph, since their values are zeros and cannot be shown in this semi-log graph.

Figure 34 Performance with added noise for 200 Mbps with CP96

Evaluation #2: Decrease the data rate from 200 Mbps to 53.3 Mbps (i.e. the lowest mandatory data rate) – This implies higher coding rate and spreading factor. CP length of 32 samples was configured in both scenarios.

This diagram shows the BER as a function of T-R distance, with different amount of added noise (SNR) per measurement location according to Table 4. Simulation results for both 200 Mbps and 53.3 Mbps were shown for comparison. The error bars represent the minimum and maximum BER values from the 10 simulation results per location. Notice that BER values for some locations are not displayed in the graph, since their values are zeros and cannot be shown in this semi-log graph.

Figure 35 Performance with added noise for 200 Mbps vs. for 53.3 Mbps

As shown in Figure 35, when the data rate is reduced from the highest mandatory rate of 200 Mbps to the lowest mandatory rate of 53.3 Mbps, the system performance is improved significantly. For NLOS behind wall channel type at T-R distance of 9.7 meter, the resulting BER with added noise is reduced from almost 5×10^{-1} at 200 Mbps to approximately 2×10^{-6} at 53.3 Mbps. The reasons for the improvement include higher encoding rate (1/3 vs. 5/8) and higher overall spreading gain (4x vs. 2x) with the lower data rate. Notice that the resulting values of BER for most channel responses are zero for 53.3 Mbps, hence cannot be plotted in the semi-log graph. Therefore, the simulation

results have demonstrated that data rate lower than 200 Mbps will provide better system performance in a typical NLOS indoor environment with added noise.

5.3 Evaluation of Various Cyclic Prefix Lengths

When CP length of 32 samples is used, bit errors can still be significant at larger T-R distance range due to the short CP length compare to the channel multipath delay. In this chapter, the effect of using different lengths of CP is examined by simulating 32, 64, and 96 samples of CP (i.e. 60.6 ns, 121.2 ns, and 181.8 ns of CP duration respectively). Data rate of 200 Mbps is used in all simulations in this chapter.

Figure 36 Measured multipath delay for LOS, NLOS behind bookshelves, and NLOS in hallway behind wall

First of all, the multipath maximum, mean and RMS delay spread values for each set of

channel response measurement as derived in [1] are shown in Figure 36. In some cases, maximum delay does not increase as distance and BER increase (e.g. NLOS in the same room with T-R distance at 5 meters). One possible explanation is that the maximum delay spread was derived by using a fixed -40dB threshold, instead of a relative threshold from the strongest signal received. This approach was taken in [1] since all resolvable multipath components were required in that study.

This plot shows the resulting BER at each location using CP length of 32, 64, 96 samples. Results for all 3 channel environments are shown: LOS, NLOS behind bookshelves, and NLOS in hallway behind wall. All simulations were run using configurations for data rate of 200 Mbps. No noise is added.

Figure 37 Performance with various Cyclic Prefix lengths for LOS, NLOS behind bookshelves, and NLOS in hallway behind wall

In order to reduce inter-symbol interference, the length of CP should be larger than the

maximum delay spread of the channel. The average measured values of Maximum Excess Delay are 65.37 ns and 63.9 ns for LOS and NLOS respectively [1], which are both larger than the default Cyclic Prefix length used (i.e. 60.6 ns). As a result, more significant bit errors can be observed when default CP length of 32 samples is used.

The simulation results of BER as a function of T-R distance using CP length of 32, 64, and 96 samples are shown in Figure 37. As expected, the error rate is visibly reduced when longer CP duration is used in all 3 types of measured channels: LOS, NLOS behind bookshelves, and NLOS behind wall. The resulting BER is reduced by up to an order of magnitude when using 2x of the default CP length, and up to 3 orders of magnitude when using 3x of the default CP length.

Note that although longer CP length can help to reduce transmission error, it will also decrease the bandwidth efficiency. Thus, more detailed analysis is needed to determine the optimum CP length that can improve better system performance while minimize the impact on bandwidth efficiency.

5.4 Performance Comparison at different data rates

This chapter compares the system performance using data rate of 200 Mbps vs. 53.3 Mbps. Cyclic Prefix of 32 samples is used. Table 5 lists some of the system parameter settings configured for 200 Mbps vs. 53.3 Mbps.

Table 5 System parameter settings for 53.3 Mbps & 200 Mbps data rate

System Parameters	53.3 Mbps	200 Mbps
Coding rate (R)	1/3	5/8
Modulation scheme	QPSK	QPSK
Frequency diversity spread	YES (2 x)	NO (1 x)
Time diversity spread	YES (2 x)	YES (2 x)
Overall spreading gain	4 x	2 x

In addition to the simulation results, data from several existing publications using IEEE UWB channel models CM1 – CM4 are also shown in this chapter for comparison. Table 6 shows the values of the Mean Excess Delay and RMS delay spread for CM1 – CM4 [21], and are compared with the channel measurements used in this thesis [1].

Table 6 Parameters for different UWB Channel Models

Channel Responses	Measurements used in this thesis		Residential		Office	
	LOS	NLOS	CM1 (LOS)	CM2 (NLOS)	CM3 (NLOS)	CM4 (NLOS)
Distance (m)	1 – 7 m	2 – 9.7 m	0 – 4 m	0 – 4 m	4 – 10 m	N/A
Mean Excess delay (ns)	9.21	4.61	5.05	10.38	14.18	N/A
RMS delay (ns)	12.15	12.8	5.28	8.03	14.28	25

CM1 and CM2 model the LOS and NLOS scenario respectively in a typical residential area with size comparable to a small office room. On the other hand, CM3 and CM4

characterize the NLOS scenario in a typical office or laboratory setup with higher number of multipaths. The resulting BER as a function of added noise (SNR) in simulations for LOS, NLOS behind bookshelves, and NLOS behind wall channels are shown in Figure 38, Figure 39, and Figure 40 respectively. The dotted curves are data extracted from some existing publications for comparison and verification. Note that the resulting BER curve for AWGN channel has better performance than an uncoded QPSK BER curve, because of the coding gain introduced by the convolutional encoding used in this study. The amount of coding gain is further verified by [23], which shows 3.7 dB of coding gain at $\text{BER} = 10^{-5}$.

This graph shows the BER performance using 200 Mbps vs. using 53.3 Mbps configuration for LOS channels. CP of 32 samples was configured in both scenarios. Data extracted from Ref [7] and Ref [22] is also shown for channel model CM2 for comparison and verification. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 38 Comparison of performance for 200 Mbps vs. 53.3 Mbps – LOS

When comparing the CM1 channel model (i.e. LOS with RMS delay spread of 5.28 ns) with our measured LOS channels, our measurements have more multipath components captured with longer RMS delay spread of 12.15 ns. This is due to the fact that there are lots of equipment and metal racks in the lab reflecting the UWB signal when the measurements were collected [1]. Therefore, the simulation results for LOS are compared with CM2 simulation result from [7] instead of CM1 in the LOS results in Figure 38. Figure 39 and Figure 40 show that the simulation results for NLOS have similar characteristics as results for CM3 and CM4 from references.

This graph shows the BER performance using 200 Mbps vs. using 53.3 Mbps configuration for NLOS in the same room behind bookshelves channels. CP of 32 samples was configured in both scenarios. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 39 Comparison of performance for 200 Mbps vs. 53.3 Mbps – NLOS behind bookshelves

This graph shows the BER performance using 200 Mbps vs. using 53.3 Mbps configuration for NLOS in the hallway behind wall channels. CP of 32 samples was configured in both scenarios. Data extracted from Ref [6] is also shown for channel model CM3 and CM4 for comparison. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 40 Comparison of performance for 200 Mbps vs. 53.3 Mbps – NLOS in hallway behind wall

Also, notice that the BER vs. SNR curves for the following measured channels exhibit a slow down of their rate of BER decreasing when SNR is increasing, i.e. the end of the curves appear to be flatter than other curves of shorter T-R distance.

- LOS: 6 & 7 meter
- NLOS behind bookshelves: 4 to 7 meters

- NLOS in hallway behind wall: 7.5 to 9.7 meters

The slow down of performance improvement for certain channels can be explained by the fact that those measured channels have higher multipath energy than other channels, hence becoming more difficult to reduce the BER by increasing the SNR.

When the data rate is reduced from the highest mandatory rate of 200 Mbps to the lowest mandatory rate of 53.3 Mbps, the system performance is improved significantly. The reasons for the improvement include higher encoding rate ($5/8$ vs. $1/3$) and higher overall spreading gain (2x vs. 4x) with the lower data rate. The simulation results show that using a lower data rate can be an alternative to improve the system performance in a typical NLOS indoor environment.

5.5 Performance at Various Noise Levels with 3x Default CP length

As discussed in Chapter 5.3, the default CP length of 32 samples is generally shorter than the measured multipath maximum delay used in this thesis, hence could have a negative impact on the system performance. To further investigate this behaviour, simulations using CP length of 96 samples instead of 32 samples were carried out (i.e. 3x default length).

The objective of this chapter is to show the magnitude of performance improvement that is possible by increasing the CP length by 3x, which is greater than the average maximum

delay spread of the measured channels. As shown in the simulation results plotted in Figure 41, Figure 42, and Figure 43, BER decreases as CP length is increased.

As discussed in Chapter 5.3, although longer CP length can improve the transmission quality, it will also reduce the bandwidth efficiency at the same time. Therefore, more detailed investigation is needed to determine the optimum CP length.

This graph shows the BER performance using CP length 32 vs. 96 samples in LOS channels. Data rate of 200 Mbps is used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 41 Performance for 200 Mbps with CP96 – LOS

This graph shows BER performance using CP length 32 vs. 96 samples in NLOS behind bookshelves channels. Data rate of 200 Mbps is used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 42 Performance for 200 Mbps with CP96 – NLOS same room

This graph shows the BER performance using CP length 32 vs. 96 samples in NLOS behind wall channels. Data rate of 200 Mbps is used. The error bars represent the minimum and maximum BER values from the 10 simulation results per location.

Figure 43 Performance for 200 Mbps with CP96 – NLOS in hallway

Chapter 6 Conclusion and Future Work

6.1 Conclusion

In this thesis, the system performance of MB-OFDM UWB in an indoor multipath environment is examined using channel measurements made in [1]. Based on the simulation results, it is concluded that the system performance of MB-OFDM UWB is tolerable in short range LOS channels. However, performance will be weak in NLOS environments with the highest mandatory data rate (200 Mbps) while using default system settings as suggested in ECMA-368 [3]. One of the reasons is because the default length of Cyclic Prefix is determined based on RMS delay spread. While in certain NLOS channels when they have larger values of delay spread (i.e. more multipaths), inter-symbol interference starts to introduce error to the transmitted signal. In addition, there is no frequency diversity at 200 Mbps data rate configuration. Therefore, system settings are adjusted in order to improve the performance in the simulations. Below is a summary of findings after comparing the resulting performance with various system parameters adjustments:

- When Cyclic Prefix is used instead of Zero-Padded Suffix (of the same length) on OFDM symbol, the system performance is slightly improved in general with an average of approximately 27% relative decrease in BER at 200 Mbps data rate with noise added.

- When CP length is increased to 2x or 3x of the default length such that it is larger than the maximum delay spread of the measured channels, the performance is enhanced accordingly. With no added noise at 200 Mbps data rate configuration, the resulting BER is reduced by up to an order of magnitude when using 2x CP length, and reduced by up to 3 orders of magnitude when using 3x CP length.
- When the data rate is reduced from 200 Mbps to 53.3 Mbps, BER is significantly reduced and the signal transmission quality is improved in NLOS indoor environments. For NLOS behind wall channel type at T-R distance of 9.7 meter, the resulting BER with added noise is reduced from almost 5×10^{-1} at 200 Mbps to approximately 2×10^{-6} at 53.3 Mbps.

By utilizing the channel impulse response measurements made by Yingzhi Chen, this thesis demonstrates the system performance of MB-OFDM UWB in an actual measured indoor environment. Better understanding of the performance in different channel environments can help to improve the design of future UWB products.

6.2 Future Work

- Research to optimize CP length in order to achieve both good system performance and bandwidth efficiency.
- Perform channel measurements in a typical family room with wooden structure and dry walls surrounding (i.e. fewer multipaths).

Chapter 7 References

The following references are sorted in the order of appearance in this thesis.

- [1] Yingzhi Chen, “Experimental Characterization of an Ultra-Wideband Indoor Wireless Channel”, M.Sc. thesis, Department of Systems and Computer Engineering, Carleton University, Ottawa, Ontario, Canada, January, 2006.
- [2] “MultiBand OFDM Physical Layer Proposal for IEEE 802.15 Task Group 3a”, *MultiBand OFDM Alliance SIG*, September 14, 2004. Available (last visited on February 10, 2009): http://www.wimedia.org/imwp/idms/popups/pop_download.asp?ContentID=6516
- [3] Standard ECMA-368, *High Rate Ultra Wideband PHY and MAC standard*, 2nd Edition, December 2007. Available: www.ecma-international.org
- [4] A. F. Molisch, J. R. Foerster, and M. Pendergrass, “Channel models for ultrawideband personal area networks,” *IEEE Personal Communications Magazine*, Volume 10, Page 14–21, December 2003.
- [5] A. Saleh, R. Valenzuela, “A Statistical Model for Indoor Multipath Propagation”, *IEEE Journal on Selected Areas in Communications*, Volume 5, Issue 2, Page 128 – 137, February 1987.
- [6] Y. Zhang, J. Zhao, “Performance Simulation of Fixed Point for MB-OFDM UWB system”, *Proceedings of Wireless Communications, Networking and Mobile Computing*, Volume 1, Page 292 – 295, 23-26 Sept. 2005.
- [7] T. Higashi, E. Taira, S. Kinjo, H. Ochi, “Performance Evaluation of MBOA UWB System Under a Multipath Channel”, *The 7th International Conference on Advanced Communication Technology (ICACT)*, Volume 2, Page(s) 1261 – 1264, February 2005.
- [8] S. M. S. Sadough, A. Mahmood, E. Jaffrot, P. Duhamel, “Performance Evaluation of IEEE 802.15.3a Physical Layer Proposal Based on Multiband-OFDM”,

- International Symposium on Telecommunications*, September 2005. Available (last visited on February 10, 2009): http://www.lss.supelec.fr/~publi/U2V5ZWQtTW9oYW1tYWQtU2FqYWQgU0FET1VHSA==799_Full%20paper%20IST%202005_sadough.pdf
- [9] D. Minoli, *Satellite Systems Engineering in an IPv6 Environment*, Auerbach Publications, Taylor & Francis Group, LLC, U.S.A., 2009. ISBN 420078682 (Section 4.1.5, Page 119)
- [10] J. M. Wilson, *Ultra-Wideband / a Disruptive RF Technology?*, Version 1.3, Intel Research & Development, September 10, 2002. Available (last visited on February 10, 2009): http://www.intel.com/technology/comms/uwb/download/Ultra-Wideband_Technology.pdf
- [11] F. Nekoogar, *Ultra-Wideband Communications: Fundamentals and Applications*, Prentice Hall, 2005. ISBN-10: 0-13-146326-8.
- [12] Homepage for WiMedia Alliance. Available: www.wimedia.org
- [13] M.B. Shoemake, *MultiBand OFDM Update and Overview*, WiQuest Communications, Inc., September 12, 2004.
- [14] B. H. Walke, S. Mangold, L. Berlemann, *IEEE 802 Wireless Systems: Protocols, Multi-Hop Mesh/Relaying, Performance and Spectrum Coexistence*, John Wiley and Sons, 2006. ISBN 978-0-470-01439-4.
- [15] P. K. Gilb, *Wireless Multimedia, A Guide to the IEEE 802.15.3 Standard*, The Institute of Electrical and Electronics Engineers, Inc., U.S.A., 2004. ISBN 0-7381-3668-9
- [16] Homepage for IEEE 802.15 Working Group for Wireless Personal Area Networks (WPANs) (last visited on May 25, 2009): <http://ieee802.org/15/index.html>
- [17] J. Ellis, K. Siwiak, R. Roberts, "TG3a Technical Requirements, IEEE P802.15 Working Group for Wireless Personal Area Networks (WPANS)", IEEE P802.15-03/030r0, December 27, 2002.

- [18] M. Engels, *Wireless OFDM Systems: How to make them work*, Kluwer Academic Publishers, 2002. ISBN 1-4020-7116-7.
- [19] Homepage for Matlab online technical documentation. Available: www.mathworks.com
- [20] M. Clark, M. Mulligan, *UWB Fixed-Point Model (Multiband OFDM)*, 2004. Available (last visited on February 10, 2009): <http://www.mathworks.com/matlabcentral/fileexchange/loadFile.do?objectId=4577>
- [21] A. Batra, J. Balakrishnan, G.R. Aiello, J.R. Foerster, A. Dabak, "Design of a multiband OFDM system for realistic UWB channel environments", *IEEE Transactions on Microwave Theory and Techniques*, vol. 52, no. 9, pp. 2123-2138, September 2004.
- [22] I. Ramachandran, Y. Nakache, P. Orlik, A. F. Molisch, J. Zhang, "Symbol Spreading for Ultrawideband Systems Based on Multiband OFDM", *15th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, Volume 2, Page 1204 – 1209, September 2004.
- [23] S. Choi, S. Choi, H. Lee, "A FEC architecture for UWB system", *IEEE 64th Vehicular Technology Conference 2006*, Page 25 – 28, September 2006.
- [24] B. Razavi, T. Aytur, C. Lam, F. Yang, R. Yan, H. Kang, C. Hsu, C. Lee, "Multiband UWB Transceivers", *IEEE Custom Integrated Circuits Conference*, Page 141 – 148, September 2005.
- [25] A. Batra, J. Balakrishnan, A. Dabak, "Multi-Band OFDM: A New Approach for UWB", *Proceedings of the 2004 International Symposium on Circuits and Systems*, Volume 5, Page V-365 – V-368, May 2004.
- [26] B. Muquet, Z. Wang, G. B. Giannakis, M. Courville, P. Duhamel, "Cyclic Prefixing or Zero Padding for Wireless Multicarrier Transmissions?", *IEEE Transactions on Communications*, Volume 50, No. 12, Page 2136 – 2148, December 2002.
- [27] B. P. Lathi, *Modern Digital and Analog Communication Systems*, Second Edition, Oxford University Press, Inc., 1989. ISBN 0195110099.

- [28] C. Li, G. Zheng, M. Wang, "Performance Analysis of Constellation Mapping and Channel Estimation in MB-OFDM UWB Systems", *IEEE International Conference on Wireless Communications, Networking and Mobile Computing*, Pages 1 – 4, September 2006.
- [29] D. C. Popescu, P. Yaddanapudi, "A Simulation Study of Interference Suppressing OFDM for UWB Systems with Multiple Users", *IEEE Radio and Wireless Symposium*, Page 359 – 362, January 2006.
- [30] D. Gerakoulis, P. Salmi, "Link performance of an ultra wide bandwidth wireless in-home network", *Proceedings of the Seventh International Symposium on Computers and Communications (ISCC'02)*, Page 699 – 704, July, 2002.
- [31] H. Lai, K. Liu, "Performance Analysis of Multiband OFDM UWB Systems With Imperfect Synchronization and Intersymbol Interference", *IEEE Journal of Selected Topics in Signal Processing*, Volume 1, NO. 3, Page 521 – 533, October 2007.
- [32] H. Luediger, B. Kull, S. Zeisberg, A. Finger, "An Ultra-Wideband Indoor NLOS Radio Channel Amplitude Probability Density Distribution", *IEEE 7th Int. Symp. On Spread-Spectrum Tech. & Appl.*, Prague, Czech Republic, Page 68 – 72, September, 2003.
- [33] H. Viittala, M. Hamalainen, J. Iinatti, "Comparative Studies of MB-OFDM and DS-UWB with Co-Existing Systems in AWGN Channel", *IEEE 17th International Symposium on Personal, Indoor and Mobile Radio Communications*, Page 1 – 5, September 2006.
- [34] H. Zhang, T. Udagawa, T. Arita, M. Nakagawa, "Home Entertainment Network: Combination of IEEE 1394 and Ultra Wideband Solutions", *IEEE Conference on Ultra Wideband Systems and Technologies*, Page 141 – 145, 2002.
- [35] J. G. Proakis, *Digital Communications* (Third Edition), McGraw-Hill, Inc., New York, 1995. ISBN 0-07-051726-6.
- [36] J. R. Foerster, V. Somayazulu, S. Roy, "A Multibanded system architecture for ultra-wideband communications", *IEEE Military Communications Conference*, Volume 2, Page 13 – 16, October 2003.

- [37] M. Hamalainen, V. Hovinen, R. Tesi, J. H. J. Iinatti, M. Latva-aho, "On the UWB System Coexistence With GSN900, UMTS/WCDMA, and GPS", *IEEE Journal on Selected Areas in Communications*, Volume 20, No. 9, Page 1712 – 1721, December 2002.
- [38] M. Nakagawa, H. Zhang, H. Sato, "Ubiquitous Homelinks Based on IEEE 1394 and Ultra Wideband Solutions", *IEEE Communications Magazine*, Volume 41, Issue 4, Page 74 – 82, April 2003.
- [39] Q. Zou, Al Tarighat, "Performance Analysis of Multiband OFDM UWB Communications With Application to Range Improvement", *IEEE Transactions on Vehicular Technology*, Volume 56, No. 6, Page 3864 – 3878, November 2007.
- [40] R. A. Scholtz, R. Weaver, E. Homier, J. Lee, P. Hilmes, A. Taha, R. Wilson, "UWB Radio Deployment Challenges", *The 11th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, Volume 1, Page 18 – 21, September 2000.
- [41] R. J. Cramer, R. A. Scholtz, M. Z. Win, "Evaluation of an Ultra-Wide-Band Propagation Channel", *IEEE Transactions on Antennas and Propagation*, Volume 50, Issue 5, Page 561 – 570, May 2002.
- [42] T. S. Rappaport, *Wireless Communications Principles & Practice*, Prentice Hall, Inc, 1996. ISBN 0-13-375536-3.
- [43] Z. Iranhhauteen, H. Nikookar, G.J.M. Janssen, "An Overview of Ultra Wide Band Indoor Channel Measurements and Modeling", *IEEE Microwave and Wireless Components Letters*, Volume 14, No. 8, Page 386 – 388, August 2004.