Experimental Characterization of an Ultra-Wideband Indoor Wireless Channel

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Abstract

The Ultra-Wideband (UWB) technology is currently developing very fast and attracts more and more global attention. A lot of research work has been carried out on the UWB technology and one of the most important tasks is to characterize the UWB wireless channels. Recently several statistical models for characterizing the large-scale path loss and the multipath delay profile for UWB propagation have been proposed both in IEEE 802.15 standards contributions and in published papers.

This thesis presents frequency domain channel measurements and the statistical modeling efforts on an indoor 2GHz – 5GHz UWB wireless channel in a typical Lab/office environment. A statistical model for the prediction of large-scale path loss is discussed for both LOS and NLOS scenarios. Time domain analysis is also performed and several statistical models are discussed to characterize the multipath delay spread and the Multipath Intensity Profile of this UWB channel.

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List of Acronyms

CDF	Cumulative Distribution Function
CFR	Code of Federal Regulations
EIRP	Effective Isotropic Radiated Power
FCC	Federal Communications Committee
GPS	Global Positioning System
GUI	Graphic User Interface
IDFT	Inverse Discrete Fourier Transform
IEEE	Institute of Electrical and Electronics Engineers
ISI	Inter Symbol Interference
LOS	Line-of-Sight
MBOA	Multi-Band OFDM Alliance
MIP	Multipath Intensity Profile
NLOS	Non-Line-of-Sight
OFDM	Orthogonal Frequency Division Multiplexing
PCS	Personal Communication Systems
PDP	Power Delay Profile
PL	Path Loss
PN	Pseudo Noise
RBW	Resolution Bandwidth
RF	Radio Frequency

RMS	Root-Mean-Square
SNR	Signal to Noise Ratio
SIR	Signal to Interference Ratio
T-R	Transmitter to Receiver
UWB	Ultra Wideband
VNA	Vector Network Analyzer
WLAN	Wireless Local Area Networks
WPAN	Wireless Personal Area Networks

List of Symbols

$H(\omega)$	the channel transfer function
$h(t)$	the channel impulse response
P_t	the transmitted power
P_r	the received power
G_t	the transmitter antenna gain
G_r	the receiver antenna gain
d	the T-R separation distance in meters
d_0	the reference T-R distance
λ	the signal wavelength in meters
d_f	the far-field distance
D	the largest physical linear dimension of antenna
n	the path loss exponent
X_σ	the path loss shadowing factor in dB with standard deviation σ dB
$PL(d_0)$	the path loss at the reference point
$a_i(t; \tau; d)$	the amplitude of the i_{th} multipath signal
$\theta_i(t; \tau; d)$	the phase of the i_{th} multipath signal
τ_i	the excess delay of the i_{th} multipath signal

N	the number of resolvable multipath components above a certain threshold
$ h(\tau) $	the amplitude of channel impulse response
p_i	the relative power of i th multipath signal
MIP	the multipath intensity profile
$MIP_{Individual}$	the individual MIP for each measurement data
$\overline{MIP}$	the time averaged MIP
PDP	the power delay profile
$\tilde{p}_i$	the spatial averaged multipath power
G	the path gain
$\hat{a}_i$	the normalized multipath amplitude
τ_m	the mean excess delay
τ_{rms}	the RMS delay spread
τ_{max}	the maximum excess delay
τ_{bb}	the pulse width
f_c	the carrier frequency
$\Delta\tau$	the time resolution
T_c	the chip duration
R_c	the chip rate
$S_{21}(\omega)$	the VNA response from port 1 to port 2
τ	the excess delay

PL_0	the reference path loss
γ	the path loss exponent
α	the path loss exponent
σ	the standard deviation of shadowing factor
k	the scale factor in linear form
τ_i	the excess delay in ns for the i_{th} multipath component.
$s(\tau_i)$	a noise-like variation with a mean value = 1
$\hat{\alpha}$	a dimensionless exponential decay constant.
$\overline{\tau_{rms}}$	the global average of τ_{rms} over all NLOS locations
W	the UWB channel bandwidth,
P_i	the dB value of $ a_i ^2$
K	the dB value of the scale factor k
α	a dimensionless exponential decay constant for the dB form
S_i	the deviation between measured P_i and calculated P_i
σ_s	the standard deviation of S_i
A	the direct arrive signal power
α_0	the exponent α at reference distance d_0
ϵ	a zero mean Gaussian random process over d
σ_ϵ	the standard deviation of ϵ

Chapter 1

Introduction

This chapter starts by introducing the motivation of this thesis work, followed by a review of relevant literatures. Then the thesis contributions are summarized and finally, an outline of the thesis is presented.

1.1 Thesis Motivation

A new wireless communications technology named as Ultra-Wideband (UWB) is currently developing very fast and attracts more and more global attention. UWB transmitters work by sending very low-powered short-pulse signals across a very wide spectrum of frequencies up to several GHz in bandwidth. UWB technology can provide high-speed data rate of several hundred Mbps within distance of tens of meters and with relatively low power consumption in milliwatts range. Also because of its low power and wide bandwidth, the UWB signal has very low spectrum density so that it has very little interference into other communication systems. This makes it possible for UWB systems to share the already assigned spectrum resources with existing wireless communication systems.

In February 2002, the FCC issued a Report & Order giving users permission to deploy low powered indoor UWB systems within the 3.1 to 10.6GHz spectrum [1]. These

guidelines make it possible to use UWB technology for short-range indoors applications such as 802.15.3a Wireless Personal Area Networks (WPAN), or as a replacement of Bluetooth or WLAN technology with higher data rate and lower cost.

In order to make such indoors UWB systems practical, the research work to characterize indoor UWB signal propagation environment and to establish statistical channel models has become very important. We know that the indoor signal propagation environment is very complicated and full of multipath signals. The low power UWB receivers usually need to exploit multipath energy by using techniques such as rake receivers and adaptive combiners. This requires detailed characterization of the UWB channel including path loss and the multipath delay parameters.

Therefore, the statistical channel modeling efforts based on the channel sounding measurements are very useful for the UWB receiver design and simulation or to evaluate the performance of UWB communication systems. This is the main motivation for the research work described in this thesis to characterize an indoor UWB wireless channel.

1.2 Review of Relevant Literature

A literature search was first conducted at the early stages of this thesis to look for the recent efforts on the channel measurements and modeling of indoor UWB channels. Some of these literatures are introduced below.

Similar frequency domain channel sounding measurements as this thesis work were performed by J. Keignart and N. Daniele and their measurement setup and methodology are presented in [2].

For indoor UWB communications, several statistical models for the UWB signal path loss and multipath delay profile have been proposed, both in IEEE 802.15 standards contributions [3, 4] and in published papers [5] – [12].

Particularly, Saeed S. Ghassemzadeh and his colleagues have conducted extensive channel measurements in residential environments and statistical models for evaluating the UWB path loss and multipath delay parameters in residential environments are presented in [5]. Also a method for measuring UWB frequency responses and then for evaluation of the path loss and impulse response of the UWB indoor channel is described in [6].

The effects of antenna directivity on path loss and multipath propagation in the UWB indoor channel are analyzed by Jason A. Dabin and Nan Ni and a channel model is then presented in [7].

The models to characterize the UWB channel multipath decaying profile for indoor environments are also analyzed and presented by S Saeed S. Ghassemzadeh and his colleagues in [8] and [9]. He also performed channel measurements in commercial buildings and a large-scale path loss model is presented in [10].

Also, Dajana Cassioli, Moe Z. Win and Andreas F. Molisch have performed both statistical modeling and simulation work and established a statistical model for the UWB indoor channels. Their work presented in [11] was based on an extensive measurement campaign in a typical modern office building with 2-ns delay resolution. Moe Z. Win and Robert A. Scholtz also carried out further research of UWB signal propagation in a typical modern laboratory/office building and a communication theoretic view is then presented in [12].

Currently, significant research work is still ongoing to gather more experience with UWB signal propagation and collect more useful data to be used in UWB system simulation and receiver design.

1.3 Thesis Contribution

The research work described in this thesis is concentrated on the experimental characterization of a 2GHz to 5GHz indoor UWB channel in a typical lab/office campus building. Channel sounding measurements were carried out at 45 different receiving locations under both Line-of-Sight (LOS) and Non-Line-of-Sight (NLOS) scenarios. Statistical analysis of several channel models are then performed based on the measurement data.

The UWB channel measurements were performed by using the frequency domain channel sounding techniques. The channel transfer function is measured at each receiving location and then the channel impulse response is calculated from the frequency domain

raw data by using Inverse Discrete Fourier Transform (IDFT) method. From the channel impulse response we can then obtain the Multipath Intensity Profile (MIP) and the Power Delay Profile (PDP) for this UWB channel. A statistical model for the prediction of large-scale path loss is discussed for both LOS and NLOS scenarios. Time domain analysis is also performed and several statistical models are discussed to characterize the multipath delay spread and the Multipath Intensity Profile for this UWB channel.

The channel models discussed in this thesis can be used for the simulation and design work of indoor UWB communication systems.

1.4 Thesis Organization

This thesis is organized in eight chapters. The first chapter is an introduction of the thesis work. The motivation for the UWB channel measurements is presented first then followed by a review of relevant literatures. After that the engineering contribution of this thesis is summarized.

Chapter 2 gives an overview of the UWB technology and some fundamentals of channel characterization. A few popular channel models are introduced followed by a brief description of different channel sounding techniques.

Chapter 3 describes the frequency domain channel sounding measurement setup and the measurement plan for this thesis work. The equipment used in the setup and the system calibration procedures are detailed and the measurement location plan is described.

Chapter 4 discusses the post processing of the measurement data. The process of obtaining the path loss and multipath delay parameters such as the Mean Excess Delay τ_m , the RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} are detailed. An example was given to better describe the data processing procedure.

Chapter 5 gives the details of the large-scale path loss model development and the model parameters. UWB signal large-scale path loss models are established for both the LOS and NLOS environments. The path loss exponents are presented as well.

Chapter 6 presents the time domain analysis of the channel impulse response and finds out the multipath parameters including the Mean Excess Delay τ_m , the RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} . Then we established the multipath delay models to characterize the UWB multipath channel.

Chapter 7 presents the analysis of multipath delay profile. A statistical model is established to predict the power of the i_{th} individual multipath component.

Chapter 8 provides the conclusions and findings of this thesis. Some future research work is also proposed.

Chapter 2

Background

This chapter gives an introduction of the Ultra-Wideband (UWB) technology and the channel characterization fundamentals. First a global overview of the UWB technology is presented. The important definitions and equations for channel characterization are then introduced in detail. A brief description of different channel sounding techniques is also given at the end of this chapter.

2.1 Ultra-Wideband Technology Overview

UWB technology is currently developing very fast and attracts more and more global attention. The term “Ultra-Wideband” refers to the spectral characteristics of this technology. It also has been called impulse radio, carrierless radio or carrier-free radio at the early days.

UWB technology uses the radio spectrum differently than the vast majority of radio communication technologies. First, the bandwidth of UWB systems is much greater than the bandwidth of any current wideband wireless system. The FCC defines an USB transmitter as an intentional radiator for which the -10dB bandwidth is equal to or greater than 20% of the center frequency or the -10dB bandwidth is at least 500MHz [13].

Second, unlike conventional wireless systems that use radio frequency (RF) carriers, many UWB applications such as radar-ranging systems and imaging systems usually do not use RF carriers. Instead, they use short impulses that have extremely fast rise and fall times in the sub-nanosecond range and in frequency domain these signals occupy large frequency range from near-DC to several GHz.

UWB technology can be used in a wide range of applications from ranging, imaging to communications. Especially, UWB technology is now being considered as a wireless air interface for high-speed data transmission, e.g., The IEEE 802.15.3a high data rate alternative PHY for Wireless Personal Area Networks (WPANs) [3].

2.1.1 History of Ultra-Wideband Technology

The idea of UWB is actually not new. In fact, the history of UWB concept can be dated back to the work done by Helmholtz in the 19th century. Early research work on UWB technology started in the 1960s to characterize the impulse responses of a certain class of microwave networks. The US military then used the UWB technology for radar and secured communications systems.

In the 1980's, the foundations of modern UWB systems were laid down in work done at the Sperry Research Center by Ross. Ross's work started in 1970's and was on the use of UWB technology as an analytical tool to explore the properties of microwave networks and to determine the intrinsic properties of materials [14]. After that his work also started to include the use of these techniques for the analysis and synthesis of antenna elements.

The development of techniques for sub-nanosecond base-band pulse generation has also been applied for short pulse radar and military communication systems.

Then started in the 1990s, more and more interests for UWB have switched to the area of short-range communications in dense multipath environments such as indoor wireless communication in both residential and commercial buildings. There are two main reasons behind this. First, UWB technology can support very high data rate which is very attracting now. Second, the very wide bandwidth of UWB signal makes it more immune to the deleterious multipath fading which usually is a big problem for indoor communication systems.

2.1.2 Regulatory Issues

Since UWB signals usually occupy a very wide frequency band so it's inevitable that the UWB system has to share the already allocated frequency spectrum with other communications systems. One big concern here is that the extent of its interference with existing RF systems is not well known and studied. So some operators of existing RF systems are worried about the impact of UWB interference on their services. This brings the important regulatory issues if the UWB technology wants to go commercial and practical.

In the United States, the FCC released its first authorization for UWB in February 2002 [1], which governed the use of UWB technology as a broadband wireless technology. The current FCC regulation for UWB is the Code Of Federal Regulations (CFR), Title

47, Part 15, “Radio Frequency Devices” Subpart F for Ultra-Wideband Operation, published in June 2005 [13]. Specifically FCC opened 7500 MHz frequency band, 3.1 GHz to 10.6 GHz for the indoor communication applications. The FCC limits the Effective Isotropic Radiated Power (EIRP) to -41.3 dBm per MHz from 3.1GHz to 10.6 GHz for both indoor and handheld devices. Additional restrictions are placed on the radiated power outside of this band to minimize interference with Global Positioning Systems (GPS), aviation systems, and other sensitive communication systems. The EIRP limits are measured using a spectrum analyzer with a 1 MHz Resolution Bandwidth (RBW) and average detection or Root-Mean-Square (RMS) detection. In addition, the regulations limit the average radiated power over the 1164MHz to 1210MHz and the 1559MHz to 1610MHz bands to -85.3 dBm per kHz when measured using a 1kHz RBW.

According to the above FCC regulations, it is likely that UWB technology will be initially used for short-range personal area networks in the tens of meters range. Thus, UWB will be competing with existing WLAN technologies like IEEE 802.11b, 802.11g, 802.11a, Bluetooth, and ETSI Hiperlan/2. However, UWB is actually capable of longer distance transmission up to a few kilometers. But FCC will require more studies and trials be carried out before they consider the possibility of using UWB for extended range transmissions.

Note that the FCC does not specify the RF physical layer interface in the regulations but rather specify the spectrum requirements to balance the needs of the many licensed and

unlicensed users. The RF physical layer interface is then defined through the standards community such as in the IEEE 802.15.3a standard for UWB communications, and industry groups like the Multi-Band OFDM Alliance (MBOA).

2.1.3 Advantages of Ultra-wideband Technology

UWB technology has many unique properties compared to the existing wireless communication technologies. Below are some major advantages of UWB technology.

- Support Extremely High Data Rate

Due to the extremely wide bandwidth, UWB technology can provide very high data rate up to several hundred Mbps in the range of tens of meters. The Institute of Electrical and Electronics Engineers (IEEE) sets the UWB specification, known as the 802.15.3a, at 110 Mbps over a distance of 10 meters [4]. Another specification using Multi-Band Orthogonal Frequency Division Multiplexing (OFDM) technology with UWB together can bring the data rate up to 480 Mbps. The initial products using these specifications are anticipated to be on the market in 2006. Future versions of UWB communication systems are expected to have data rate up to 1 Gbps.

- Immune to Multipath Fading

UWB technology is also more immune to the multipath fading issue compared to the existing wireless technologies. The extremely wide bandwidth of UWB signal means extremely short time duration of waveform in the time domain – short pulses. We know that multipath fading is a big problem for the wireless communication systems working at

indoor environments. For narrowband signals the multipath cancellation occurs when a strong reflected signal – e.g., off of a wall, ceiling, window, objects, etc. – arrives partially or totally out of phase with the direct path signal, causing a reduced amplitude response in the receiver. While transmitting UWB signals which are basically extremely short pulses, most of the reflected multipath signals can be resolved from each other and from the direct arrival signal so that the cancellation effect will be much less compare to the narrowband signals. As a consequence, the likelihood for distortion caused by the reflected signals is low. This makes UWB technology particularly well suited for high-speed, mobile wireless applications in the multipath-rich indoor environments.

- Reuse of Existing Spectrum Resources

Another advantage of UWB is because UWB signal pulses have low power and spread into a wide frequency range, they have extremely low power spectrum density that is often below the noise floor. Thus, in theory, UWB signals can coexist with the RF carriers of conventional wireless systems such as cellular band, PCS band or 802.11 networks and the interference is neglectable. The UWB services can then be deployed using existing frequency spectrum, without acquiring new and expensive spectrum in the high frequency bands. Essentially, the UWB technology enables the reuse of valuable spectrum resources. As the radio spectrum becomes more and more crowded due to the today's increasing demand for wireless communications and the RF spectrum is a scarce resource worldwide, this advantage makes UWB very attracting to the industry.

- Low Cost and Easy to Design

Besides the above advantages, the UWB technology can be implemented at a lower cost than existing wireless communication technologies. With recent advances in VLSI, UWB systems can be implemented nearly “all-digital” with minimal RF or microwave electronics required. Because UWB transceivers use short pulses for transmission, so the UWB systems are highly frequency adaptive, enabling them to be positioned anywhere within the RF spectrum.

- Low Power Consumption

UWB systems consume relatively lower power compare to today’s popular wireless technologies. Due to the low power density limits from the regulatory requirements, UWB systems work in the milliwatts power range. This makes it possible for UWB systems to use existing VLSI technology and is then suitable for battery powered portable devices, such as cell phones or PDAs that users carry all the time.

- Accurate Positioning

UWB technology also offers accurate and inexpensive geographic position location capabilities that can be combined with communications capability. For radar applications, the extremely narrow UWB pulses can provide very fine range resolution and precision distance and positioning measurements capabilities.

2.2 Indoor Wireless Channel Characterization

As we all know, the radio channel places fundamental limitations on the performance of wireless communication systems. For indoor wireless communications the research work to characterize the signal propagation channels and to establish statistical channel models has become even more important, since the indoor signal propagation environments are very complicated and are generally full of multipath signals. The transmission between the transmitter and the receiver can vary from simple Line-of-Sight (LOS) to one that is severely obstructed by walls or other objects. The modeling of such indoor wireless channels has been one of the most difficult parts of radio system design, and is typically done in a statistical fashion, based on measurements made specifically for an intended communication environment. These channel models will be very helpful to evaluate the performance of wireless communication systems. In the following sections we will look at some important definitions and equations about channel characterization and channel modeling.

2.2.1 Large-scale Propagation models

Propagation models that predict the average received signal strength at a given transmitter – receiver (T-R) separation distance are useful in estimating the radio coverage area of a transmitter and are called **Large-scale Propagation models**, since they characterize the relatively slow large-scale variations for the average received power. Two useful models are introduced below.

- Free Space Propagation Model

As a beginning we introduce the free space propagation model here. The free space environment assumes that the transmitter and receiver have a clear and unobstructed line-of sight path between them and there are no reflecting objects around. It also assumes that, within this environment, the atmosphere behaves as a perfectly uniform and non-absorbing medium.

The free space model predicts that the received power decays according to an inverse-square law with the T-R separation distance d , as described by the Friis free space equation [15],

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2} \quad (2.1)$$

where,

P_t is the transmitted signal power

$P_r(d)$ is the received signal power at T-R distance d

G_t is the transmitter antenna gain

G_r is the receiver antenna gain

d is the T-R separation distance in meters

λ is the signal wavelength in meters.

We define large-scale Path Loss (PL) as the difference in dB between the transmitted power and the received power. The large-scale here refers our interest of predicting the

average received power, not the small-scale fast variations of signal strength. The large-scale path loss (in dB) at distance d for the free space model is given by [15],

$$PL(d) = 10 \log\left(\frac{P_t}{P_r}\right) = 10 \log\left[\frac{(4\pi)^2 d^2}{G_t G_r \lambda^2}\right] \quad (2.2)$$

From equation 2.2 we see that the free space path loss PL increase with T-R separation distance d at a rate of 20dB/decade.

Note that the Friis free space model is only valid when the receiver is located in the far-field of the transmitter antenna. The far-field of a transmitting antenna is defined as the region beyond the far-field distance d_f , which is given by [15],

$$d_f = \frac{2D^2}{\lambda} \quad (2.3)$$

where D is the largest physical linear dimension of the antenna. Additionally, to be in the far-field region, d_f must also satisfy $d_f \gg D$ and $d_f \gg \lambda$.

- Log-distance Path Loss Model

Both theoretical and measurement-based propagation models indicate that average received signal power decreases logarithmically with distance. Such models have been used extensively in the literature. The average large-scale path loss for any T-R separation distance d is expressed as a function of distance by using a path loss exponent, n [15].

$$PL(d) \propto \left(\frac{d}{d_0}\right)^n \quad (2.4)$$

or if expressed in dB form [15],

$$PL(dB) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) \quad (2.5)$$

where n is the path loss exponent which indicates the rate at which the path loss increases with T-R separation distance, d_0 is a reference distance which is close to the transmitter but far enough to be in the far-field region of the transmitting antenna, and d is the T-R separation distance.

If we plotted the above path loss model on a log-log scale, the path loss is a straight line with a slope equal to $10n$ dB per decade. The value of n depends on the specific propagation environment and the type of signal transmitted. For example, the free space model is the case when n is equal to 2, and for non-line-of-sight (NLOS) scenarios, the typical n will be greater than 2.

- Log-normal Shadowing

The above log-distance path loss model in equation (2.5) does not consider the fact that the surround environment might be totally different at two different receiver locations that have the same T-R separation distance d . This means that the actual path loss at a certain location might be a lot different than the **average** value predicted by equation (2.5). Extensive measurements have shown that at any value of d , the path loss $PL(d)$ at a particular location is a random value and is distributed log-normally (normal in dB) about the mean value. This phenomenon is referred as **log-normal shadowing**. Thus equation (2.5) can be revised as [15]

$$PL(dB) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (2.6)$$

where X_σ is a zero-mean Gaussian distributed random variable (in dB) with standard deviation σ also in dB.

2.2.2 Small-scale Multipath Propagation

Small-scale fading is used to describe the rapid fluctuation of the received power over a small change of T-R distance. This small-scale variation of signal power is often caused by interference between two or more versions of the transmitted signal which arrive at the receiver at slightly different times. These different signals are called multipath signals.

Multipath in the radio channel occurs because of the reflections from the ground and surrounding structures. So that the radio waves at the receiver arrive from different directions with different propagation delays. This phenomenon creates a constantly changing environment for the wireless communication.

2.2.3 Impulse Responses Model of a Multipath Channel

For a general multipath channel, we can assume it's linear so that the channel can be completely characterized by its impulse response. Usually, a radio channel is time and location dependent and so the channel impulse response can be written as below [15]:

$$h(t; \tau; d) = \sum_{i=0}^{N-1} a_i(t; \tau; d) e^{j\theta_i(t; \tau; d)} \delta(\tau - \tau_i) \quad (2.7)$$

where,

$a_i(t; \tau; d)$ and $\theta_i(t; \tau; d)$ are the amplitude and phase of the i_{th} multipath signal

τ is the multipath delay spread

τ_i is the excess delay of the i_{th} multipath signal

d is the T-R separation distance in meters

N is the number of resolvable multipath components above a certain threshold

For an indoor environment, the T-R separation distance is short and we can assume no big movement in the surrounding objects, then we can assume the channel is essentially time-invariant and is so called the quasi-static channel. This will be proved in the following chapters with the measurement data and the time variables will be further eliminated by averaging a number of snapshots of the channel response at a certain location. In this case, we can re-write the channel impulse response at a certain receiving location as:

$$h(\tau) = \sum_{i=0}^{N-1} a_i e^{j\theta_i} \delta(\tau - \tau_i) \quad (2.8)$$

2.2.4 Multipath Intensity Profile and Power Delay Profile

For a multipath channel there are two types of multipath delay profiles: Multipath Intensity Profile (MIP) and Power Delay Profile (PDP).

We define the time invariant **Multipath Intensity Profile (MIP)** for a given receiving location as the square of the absolute value of the channel impulse response $h(\tau)$ [5], so we have:

$$\text{MIP} = |h(\tau)|^2 \quad (2.9)$$

From the channel impulse response equation (4.4) we then define the MIP for each individual measurement data set as:

$$MIP_{Individual} = |h(\tau)|^2 = \sum_{i=0}^{N-1} |a_i|^2 \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} p_i \delta(\tau - \tau_i) \quad (2.10)$$

where $p_i = |a_i|^2$ is the relative power of the i_{th} multipath component. In my measurements, in order to get a quasi-static MIP, 10 snapshots of channel frequency responses were measured at each receiver location and we will have 10 individual MIPs. The average of these 10 individual MIPs is the time averaged channel MIP which we need for our channel characterization and it is denoted as $\overline{MIP}$:

$$\overline{MIP} = \overline{|h(\tau)|^2} = \sum_{i=0}^{N-1} \overline{|a_i|^2} \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} \overline{p_i} \delta(\tau - \tau_i) \quad (2.11)$$

where the upper bar represents time average.

The **Power Delay Profile (PDP)** is defined as the local spatial average of the individual MIPs [5] while each individual MIP is related to a grid point in the 5 x 5 grids. So the PDP is related to the small-scale variation of the received signal. Usually we measure a certain number of MIPs according to a spatial grid and then take a spatial average of all the $|a_i|^2$ for all the different grid positions. In my measurements the grid is 5x5 with grid increment 10cm, so the whole measurement area is 40cm x 40cm. So we have:

$$PDP = |\tilde{h}(\tau)|^2 = \sum_{i=0}^{N-1} |\tilde{a}_i|^2 \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} \tilde{p}_i \delta(\tau - \tau_i) \quad (2.12)$$

while the “ $\sim$ ” represents spatial average.

2.2.5 Path Gain and the Normalization of Multipath Intensity Profile

We define **Path Gain G** as [8]

$$G = \sum_{i=0}^{N-1} |a_i|^2 \quad (2.13)$$

So that the channel impulse response can be rewritten as

$$h(\tau) = \sqrt{G} \sum_{i=0}^{N-1} \hat{a}_i e^{j\theta_i} \delta(\tau - \tau_i) \quad \text{where} \quad \hat{a}_i = a_i / \sqrt{G} \quad (2.14)$$

In this case the channel is characterized by the path gain G and the N -fold set $\{\tau_i\}$. The path gain is actually related to large-scale path loss. And the set $\{\tau_i\}$ is related to small-scale multipath effects.

Accordingly we define the normalized MIP as

$$\hat{MIP} = \sum_{i=0}^{N-1} \hat{p}_i \delta(\tau - \tau_i) \quad (2.15)$$

where

$$\hat{p}_i = \frac{|a_i|^2}{\sum_{i=0}^{N-1} |a_i|^2} \quad \text{and} \quad \sum_{i=0}^{N-1} \hat{p}_i = 1 \quad (2.16)$$

Since the sum $\sum_{i=0}^{N-1} \hat{p}_i = 1$, this means that the normalized MIP is non-negative and integrates to 1, so it will have the properties of a probability distribution function over the excess delay τ_i , then the Mean Excess Delay τ_m and the RMS Delay Spread τ_{rms} will just be the first and second moment of the normalized MIP.

2.2.6 Multipath Channel Parameters

In order to compare different multipath channels and to develop some general design guidelines for wireless communication systems, some parameters which grossly quantify the multipath channel are used. The Mean Excess Delay τ_m , RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} are key multipath channel parameters that can be determined from a MIP or PDP.

The **Mean Excess Delay** τ_m is defined as the first moment of the normalized MIP, as following:

$$\tau_m = \frac{\sum_{i=0}^{N-1} \tau_i \cdot |h(\tau_i)|^2}{\sum_{i=0}^{N-1} |h(\tau_i)|^2} = \frac{\sum_{i=0}^{N-1} \tau_i \cdot |a_i|^2}{\sum_{i=0}^{N-1} |a_i|^2} = \sum_{i=0}^{N-1} p_i \tau_i \quad (2.17)$$

The **RMS Delay Spread** τ_{rms} is the square root of the second central moment of the normalized MIP, as following:

$$\tau_{rms} = \sqrt{\frac{\sum_{i=0}^{N-1} (\tau_i - \tau_m)^2 \cdot |h(\tau_i)|^2}{\sum_{i=0}^{N-1} |h(\tau_i)|^2}} = \sqrt{\frac{\sum_{i=0}^{N-1} (\tau_i - \tau_m)^2 \cdot |a_i|^2}{\sum_{i=0}^{N-1} |a_i|^2}} = \sqrt{\sum_{i=0}^{N-1} p_i (\tau_i - \tau_m)^2} \quad (2.18)$$

These delays are measured relative to the first detectable signal arriving at the receiver at $\tau_0 = 0$. Equations (2.17) and (2.18) do not rely on the absolute power levels of the multipath components, but only on the relative amplitudes of the multipath components. That's why we can use the normalized MIP or PDP to calculate the two delay parameters.

The RMS Delay Spread τ_{rms} is often used to estimate the maximum data rate that the UWB channel can transmit. So it is a very important multipath parameter for the channel.

The **Maximum Excess Delay** τ_{max} is defined to be the time delay during which multipath power falls to below a certain threshold (X dB). This threshold can either be chosen as a certain value (in dB) below the level of the strongest arriving multipath component, or as a certain value (in dB) above the noise level. In other words, the maximum excess delay is defined as $\tau_x - \tau_0$, where τ_0 is the absolute delay of the first arriving signal and τ_x is the maximum delay at which a multipath component is below the X dB threshold.

Figure 2.1 on the next page show examples of two types of maximum excess delay computation. As in the Figure, the strongest arriving component is at $\tau_0 = 0$ and the level is at -12 dB. One threshold is set to be -32 dB which is 20 dB below the maximum signal level, and the corresponding maximum excess delay is 33.6 ns. Another threshold is set to be -40 dB which is 3 dB above the noise level, and the corresponding maximum excess delay is 61.4 ns in this case.

Figure 2.1 An Example of Maximum Excess Delay Computation

2.3 Channel Sounding Techniques

Because of the importance of the multipath structure in determining the channel characteristics, a number of wideband channel sounding techniques have been developed to measure the multipath delay profile of the wireless channels. These techniques include direct pulse measurements, spread spectrum sliding correlator channel measurements and swept frequency channel sounding measurements.

2.3.1 Direct RF Pulse Channel Sounding

Direct RF pulse channel sounding is a simple and effective method to measure the multipath delay profile of any wireless channel. As shown in Figure 2.2, the measurement system includes a transmitter which transmits a repetitive RF wideband probing pulse of width of τ_{bb} s and carrier frequency f_c , and a receiver with a wide band-pass filter (BW $= 2/\tau_{bb}$ Hz). The received signal is then amplified, detected with an envelope detector, and then displayed and stored on a high-speed digital oscilloscope. This gives an immediate measurement of the square of the channel impulse response $h(t)$ convolved with the probing pulse. The time resolution ($\Delta\tau$) between multipath components is equal to the probing pulse width τ_{bb} [15].

Figure 2.2 Direct RF Pulse Channel Sounding System Block Diagram

The problem with this channel sounding system is that it is subject to interference and noise, due to the wide band-pass filter required for multipath time resolution. Also, the

direct pulse channel sounding system relies on the ability to trigger the oscilloscope on the first arriving signal. So if the first arriving signal is very low, the system may not trigger properly. Another disadvantage of this technique is that the phases of the individual multipath components are not measured, due to the use of an envelope detector.

2.3.2 Spread Spectrum Sliding Correlator Channel Measurements

The Spread Spectrum Sliding Correlator method is another channel sounding technique.

It also includes a transmitter and a receiver as shown in Figure 2.3.

Figure 2.3 Spread Spectrum Sliding Correlator Channel Sounding System

In the transmitter, a carrier signal of frequency f_c is spread over a large bandwidth by mixing it with a binary pseudo-noise (PN) sequence having a chip duration T_c and a chip rate R_c equal to $1/T_c$ Hz [15].

This spread spectrum signal is then received, filtered, and despread using a PN sequence generator that is identical to that used at the transmitter and the chip rate is slightly lower. Mixing the chip sequences in this fashion implements a sliding correlator. When the PN code of the faster chip clock catches up with the PN code of the slower chip clock, the two chip sequences will be virtually identically aligned, giving maximal correlation. When the two sequences are not maximally correlated, mixing the incoming spread spectrum signal with the unsynchronized receiver chip sequence will spread the received signal into a bandwidth at least as large as the bandwidth of the receiver's reference PN sequence. In this way, the narrowband filter that follows the correlator can reject almost all of the incoming signal power. This is how the spread spectrum receiver can reject passband interference, unlike the direct pulse channel sounding system.

When the received signal is correlated with the receiver sequence, the signal is despread back to the original bandwidth, envelope detected, and displayed on an oscilloscope. Since different incoming multipath signals will have different time delays, they will maximally correlate with the receiver PN sequence at different times. Therefore, after envelope detection, the channel impulse response convolved with the pulse shape of a single chip is displayed on the oscilloscope.

The time resolution $\Delta\tau$ of multipath components using a spread spectrum system with sliding correlation is

$$\Delta\tau = 2T_c = 2/R_c \quad (2.19)$$

In other words, the system can resolve two multipath components as long as they are equal to or greater than $2T_c$ seconds apart.

One advantage of this method is that, while the probing signal is still wideband, it is possible to detect the transmitted signal using a narrowband receiver preceded by a wideband mixer, thus reducing the noise level of the receiver and improving the dynamic range of the system as compared to the direct pulse channel sounding method. Also, the required transmitter power can be considerably lower than the comparable direct pulse system due to the inherent processing gain of spread spectrum systems.

Of course this method has its disadvantages too. One is that the measurements are not made in real time, but they are compiled as the PN codes slide past one another. So the measurement time might be too long for a fast changing channel. Another disadvantage of this system is that a noncoherent detector is used in the receiver, so that the phases of the individual multipath components can not be measured. Even if a coherent detector is used, the phases of individual multipath components with different time delays would still be measured at substantially different times, during which the channel might change if the channel is time-variant.

2.3.3 Frequency Domain Channel Sounding Technique

The last channel measurement method we introduce here is the frequency domain channel sounding technique. Because of the relationship between time domain and frequency domain, it is possible to measure the channel impulse response in the frequency domain [15].

Figure 2.4 Frequency Domain Channel Measurement System

The system block diagram is shown in Figure 2.4. 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 channel impulse response by using Inverse Discrete Fourier Transform (IDFT) processing. The number and spacing of these frequency steps will impact the time

resolution and time duration of the channel impulse response so they have to be carefully chosen before the measurements.

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 the vector network analyzer is required. So it is useful mainly for very close measurements such as the 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, we should keep the total sweep time interval as small as possible.

Chapter 3

UWB Channel Measurements

This chapter presents the details of the frequency domain channel sounding measurements performed on the indoor UWB channel for this thesis work. The channel details are introduced at the beginning. Then the measurement setup and test equipment specifications are given in details. After that we present the measurement plan for the channel characterization.

3.1 Channel Details and Measurement Technique

3.1.1 UWB Channel Details

This thesis work is concerned with the experimental characterization of an indoor UWB channel from 2GHz to 5GHz in a typical lab/office environment. Both the Line-of-Sight (LOS) and Non-Line-of-Sight (NLOS) scenarios are measured and analyzed.

- Location

The location for the channel measurements is in room 4050 of MINTO building at Carleton University. It is a lab/office environment with big glass windows on two sides and metal racks beside the wall. There are a number of equipments on the rack and inside the room. There is also a bookshelf sitting at the middle of the room serving as obstacle

objects to provide NLOS measurement locations. Overall, this room can be used as a typical lab/office environment for indoor UWB channel measurements.

- Channel Frequency Parameters

The frequency band of the measured UWB channel is chosen to be from 2GHz to 5GHz, centered at 3.5GHz. The UWB bandwidth in this case is 3GHz which corresponds to a multipath resolution of 0.33ns on time domain.

The reason that we did not choose a higher upper frequency is due to the equipment limitation and the consideration to reduce the measurement uncertainty.

3.1.2 Determination of Channel Sounding Technique

As we introduced in Chapter 2, there are three different kinds of channel sounding techniques for we to choose. For this particular UWB channel, we decided to use the frequency domain channel sounding method.

First of all, the advantage of frequency domain channel measurement is that it can measure both the amplitude and phase information of each individual multipath component. While both the direct pulse method and the spread spectrum method can not provide this. Also this setup does not need pulse generator and digital oscilloscope. Since this UWB channel has a 3GHz bandwidth and centered at 3.5 GHz, if we use direct pulse sounding technique, a sub-nanosecond pulse generator and an extremely high-speed digital oscilloscope would be required and the proper triggering of the receiver would be

difficult to accomplish. On the other hand, a VNA that works up to 6GHz is easy to find and there is no need for synchronization.

We know that there are certain disadvantages for the frequency domain measurement technique. For example, this kind of measurement system requires careful calibration and a cable between receiver and the vector network analyzer is required. However, in this particular channel measurement, because it is an indoor environment and all the receiving locations are close to the transmitter (T-R distance is less than 20 meters), the above advantages are not significant. Another limitation with frequency domain measurements is that there is a certain time required for scanning the whole frequency band so the measurement is not real-time. In our case, also because it is an indoor environment and no fast moving objects in present so the channel is quasi-stationary. It's not a rapid changing time-variant channel. Also we try to time interval as small as possible to reduce the sweep time.

3.2 Channel Measurement System

3.2.1 Measurement Setup

The UWB channel measurements carried out for this thesis work are frequency domain channel sounding measurements.

The block diagram of the frequency domain UWB channel measurement setup is shown in Figure 3.1 and the pictures are presented in Appendix A.

Figure 3.1 Channel Measurement Setup Block Diagram

As shown in Figure 3.1, The HP 8753C Vector Network Analyzer (VNA) generates a frequency swept signal from port 1 and the signal is then transmitted by a monopole transmitting (Tx) antenna close to the VNA. The Tx antenna is connected to the VNA by a short cable. The signal is propagated through the indoor UWB multipath channel and then received by a receiving (Rx) antenna of the same type. From the Rx antenna the received signal is first passed through a 60 feet long cable and then feed into a HP 8349B Microwave Amplifier with a gain of about 30dB. The signal is then returned to VNA port 2. By comparing the output signal from port 1 and the input signal from port 2 we then have the transfer function $S_{21}(\omega)$ as we needed.

3.2.2 The Vector Network Analyzer

The main equipment in this measurement setup is a HP 8753C Vector Network Analyzer (VNA) used to measure the channel transfer function of the channel. The main specifications of this Network Analyzer are listed below:

- Frequency range from 300kHz to 6GHz
- Frequency points: up to 1601 points
- Sweep time: 0.8s for 1601 points
- Maximum output power: 10dBm
- System Dynamic Range: 105dB
- Noise floor: -115dBm (10Hz BW)

3.2.3 The Microwave Amplifier

A HP8349B microwave amplifier is used to magnify the received signal after the cable loss of the long cable. This will increase the dynamic range of the measurement system.

The main specifications of this amplifier are listed below:

- Frequency range from 2.0 to 20.0 GHz
- Gain: 30dB
- Maximum output power: 19 dBm from 2.0 to 18.6 GHz, 17 dBm from 18.6 to 20.0 GHz

3.2.4 Measurement Antennas

The transmitting and receiving antennas used for this measurement setup are vertically polarized omni-directional monopole antennas as shown in Figure 3.2 on the next page.

Figure 3.2 Quarter wavelength vertical antenna with drooping radials

They are quarter wavelength monopole antennas and the center frequencies are chosen as below:

- Tx antenna: center frequency at 3GHz and thus the antenna length is 2.5cm.
- Rx antenna: center frequency at 4GHz and thus antenna length is 1.9cm.

Therefore the combined frequency response of the Tx and Rx antennas will have a relatively flatter response over the range of 2GHz to 5GHz. The antenna has drooping radials which would be about 45 degrees from horizontal. These drooping radials simulate an artificial ground and lead to antenna impedance of about 50 ohms.

3.2.5 Measurement Automation

A laptop computer is used for equipment control and data collection purposes. An Agilent VEE Pro program is written and running on the laptop for automated testing and data collection. The communication of the Laptop and the VNA is achieved through the GPIB interface.

Figure 3.3 VEE Program GUI

The Graphic User Interface (GUI) of this VEE Pro program is shown in Figure 3.3. Agilent VEE Pro is a powerful graphical programming environment for automated testing. The program automatically takes 10 snapshots of the channel transfer function and save the complex data into the Laptop hard drive.

3.3 Channel Measurement Plan

3.3.1 Frequency Domain Channel Sounding Parameters

The parameters for the channel sounding measurements should be carefully selected in order to assure adequate multipath resolution and at the same time reducing the total time required for frequency sweep.

- The measurement frequency range is chosen to be 2GHz to 5GHz. So the UWB bandwidth is $W = 3\text{GHz}$ and it corresponds to a time resolution of 0.33ns for the impulse response.
- The VNA sweeps the 2GHz – 5GHz frequency range for 801 points and records the 801 received complex tones. So the frequency step is 3.75MHz which corresponds to time domain duration of 267ns. In other words, the measurement system is capable of catching multipath components that arrive with a delay up to 267ns. This duration of impulse response is found to be long enough for an indoor environment measurement.
- The sweep time for each scan is 400ms, which corresponds to a maximum measurable Doppler shift of about 2.5Hz. The sweep time is found to be short enough for our measurements since there are no fast-moving objects for indoor environments.
- The VNA's output power level is chosen to be 10 dBm to assure that the VNA is working in the linear range of its detectors and well above the noise floor.

The complex frequency response data from the VNA is then stored on a laptop via GPIB interface. For each sweep, the VNA gives the 801 points of magnitude and phase of

each frequency point and then an Inverse Discrete Fourier Transform (IDFT) can be performed to get the time domain impulse response for further analysis.

3.3.2 Measurement System Calibration

Before we take the channel measurements, the frequency domain channel sounding system needs to be carefully calibrated. The calibration is performed with the transmitting (Tx) and receiving (Rx) antenna apart with 1m separation distance. This 1m T-R separation distance d_0 is chosen to be the reference distance for the large-scale path loss model. This choice of the reference distance ensures that the Rx antenna is close to the transmitter and far enough to fall into the far field range of the Tx antenna at the channel center frequency.

By doing this calibration at the reference distance, all the following components are removed from our measured frequency response $S_{21}(\omega)$:

- Antenna gain for both the Tx and Rx antennas
- Cable loss and other losses due to the connectors or adapters.
- LNA amplifier gain and frequency response

3.3.3 Measurement Locations

Using the above measurement setup, channel sounding measurements were performed in MINTO 4050, a typical Lab/office environment in a campus building at the Carleton University. Hundreds of measurement data were taken at different T-R distances and covered both LOS and NLOS scenarios, as shown in Figure 3.4.

Figure 3.4 MINTO 4050 floor plan and measurement locations

The transmitting monopole antenna from the VNA was always located at a fixed position, and the receiving monopole antenna was moved throughout the lab and in the hallway for all the receiving locations. The heights of the Tx and Rx antennas are both fixed to 1.5m above the floor.

The T-R separation distance d between the transmitting monopole antenna and the receiving monopole antenna at each measurement location was actually measured during the measurement process. This measured T-R distance d is used for the data analysis process.

Most of the channel measurements were performed inside the room for both LOS and NLOS scenarios. Some NLOS measurements were also performed in the hallway outside the room. As a total more than 40 different receiver locations were measured and the biggest T-R distance is 12 meters.

Measurements were made for 3 different scenarios. The first one is within Line of Sight (LOS), the second one is within Non Line of Sight (NLOS) in the lab room. The third scenario is also for NLOS but the receiving antenna is located in the hallway. (See Figure 3.4).

For the LOS scenario 25 different locations were measured with T-R distance from 2m – 8m. For the NLOS scenario in the room 11 different locations were measured with T-R distance from 2m – 7m. For the NLOS scenario in the hallway 9 different locations were measured and with T-R distance from 6.5m – 12m. The transmitting antenna location is chosen to be close to the middle of the lab room.

A summary of all the locations is presented in Table 3.1.

Table 3.1 Measurement locations

Scenario	Number of locations	Min T-R distance	Max T-R distance
LOS	25	1	8
NLOS (In the Lab)	11	2	7
NLOS (Hallway)	9	6.5	12

All measurements were made while the transmit antenna and receiving antenna remained stationary. Also no fast moving objects are present during measurements. In this case the UWB channel is assumed to be quasi-stationary. In order to prove this and to get an average impulse response, 10 snapshots of the channel frequency responses were measured continuously at each location.

In order to analyze the small-scale characteristics of the UWB channel, for some typical receiving locations under both LOS and NLOS scenarios, additional measurements were performed on a 5 x 5 horizontal grid with 40cm on each side. The grid spacing is chosen to be 10cm, which is a number between the 6cm wavelength of 5GHz and the 15cm wavelength of 2GHz. Also the total 40cm length on each side will cover about 5λ for the center frequency 3.5GHz. So we can get the local spatial average of the channel impulse response and then to analyze the small-scale fading effects of the UWB channel.

Chapter 4

Measurement Data Post Processing

This chapter discusses the post processing of the measurement data. As discussed in the last chapter, the raw data we obtained from the channel measurement is the S_{21} data series recorded by the VNA, which is the transfer function $H(\omega)$ of this 2GHz to 5GHz UWB channel.

In order to characterize the channel and establish the channel model we need to get the channel impulse response $h(t)$ from the $H(\omega)$ raw data. This can be accomplished by using the Inverse Discrete Fourier Transform (IDFT) method. Once we have the channel impulse response $h(t)$, we can then calculate the important parameters of this channel, such as pass loss, multipath intensity profile, RMS delay spread, mean excess delay and maximum excess delay, etc.

4.1 Calculate the channel impulse response

After the measurement data were taken, we need to apply a multi-steps post processing of the data for analysis purpose. The first step would be to convert the raw data measured in the frequency domain into the channel impulse response in the time domain.

The raw data is the S_{21} data points acquired from the VNA reading. It is the frequency domain transfer function $H(\omega)$ of this indoor UWB channel. It contains a series of 801 complex data points related to individual frequencies from 2GHz to 5GHz so that each data point has an amplitude and phase.

To show the $H(\omega)$ more clearly, we here plot one set of measurement $H(\omega)$ data as an example below. This data set is the one of the 10 snapshots measured at the receiver location with T-R distance $d=3\text{m}$ of the LOS scenario.

Figure 4.1 shows the amplitude plot of this $H(\omega)$ in dB form and against frequency. Frequency range is from 2GHz to 5GHz.

Figure 4.1 Amplitude of Channel Transfer Function

Figure 4.2 shows the phase plot of this $H(\omega)$ against frequency. Frequency range is from 2GHz to 5GHz.

Figure 4.2 Phase of Channel Transfer Function

We then perform an Inverse Discrete Fourier Transform (IDFT) on the above frequency domain $H(\omega)$ data, and obtain the channel impulse response $h(t)$ also in a complex data form. The time duration of this impulse response $h(t)$ will be 267 ns as we discussed in the last chapter. The amplitude of $h(t)$ calculated from the above example data set is plotted in Figure 4.3. The same impulse response expressed in dB is shown in Figure 4.4.

Figure 4.3 Channel Impulse Response Amplitude in Linear Scale

Figure 4.4 Channel Impulse Response Amplitude in Logarithmic Scale

4.2 Calculate the Multipath Intensity Profile and Power Delay Profile

The next step would be to calculate the **Multipath Intensity Profile (MIP)** and **Power Delay Profile (PDP)** for each individual measurement data set.

Let's start with the channel impulse response $h(t)$ we obtained in the last step and as an example the amplitude of $h(t)$ was plotted in Figure 4.3. First of all we find the first arrival signal which is at $t_1 = 6.8$ ns in this example. The $t_1 = 6.8$ ns is calculated as the signal propagation time for 2m distance which is the difference between the T-R separation $d = 3$ m and the reference $d_0 = 1$ m.

Then we define the **Excess Delay τ** to replace the time variable t as:

$$\tau = t - t_1 \quad (4.1)$$

Thus the time point for this first arrival signal is always at $\tau = 0$ ns. The whole impulse response curve $h(t)$ is then shifted for t_1 ns on the time axis corresponding to this change and becomes $h(\tau)$.

$$h(\tau) = \sum_{i=0}^{N-1} a_i e^{j\theta_i} \delta(\tau - \tau_i) \quad (4.2)$$

As an example the amplitude of this impulse response $h(\tau)$ is plotted in Figure 4.5 against excess delay τ . We can see that the whole curve is shifted to the left hand side for 6.8 ns compare to the $h(t)$ shown in Figure 4.3.

Figure 4.5 Channel Impulse Response Amplitude in Linear Scale

(Time axis shifted to align with the first arrival signal)

Then we take the square of the magnitude of the individual impulse response $h(\tau)$ and we now have the MIP for each individual snapshot as

$$MIP_{Individual} = |h(\tau)|^2 = \sum_{i=0}^{N-1} |a_i|^2 \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} p_i \delta(\tau - \tau_i) \quad (4.3)$$

To obtain the time averaged MIP for each measurement location, we calculate each individual MIP for all the 10 snapshots and then take an average of the 10 values on each time point τ_i . The time averaged MIP for a certain receiver location is expressed as below:

$$\overline{MIP} = \overline{|h(\tau)|^2} = \sum_{i=0}^{N-1} \overline{|a_i|^2} \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} \overline{p_i} \delta(\tau - \tau_i) \quad (4.4)$$

while the upper bar denotes time average of 10 snapshots.

To simplify the expression, in the following chapters we will just use the term MIP to refer to the time averaged MIP of 10 snapshots and do not use the upper bar anymore.

As an example the time averaged MIP curve for the above LOS, d=3m example is shown in Figure 4.6.

Figure 4.6 Time averaged Multipath Intensity Profile (MIP)

The way to calculate the PDP for each measurement location is quite similar to the time averaged MIP calculation. The only difference is that instead of taking time average of the 10 snapshots we take a spatial average of all the 25 individual MIPs of the 5x5 grid of measurement points. And we have:

$$PDP = |h(\tau)|^2 = \sum_{i=0}^{N-1} |\tilde{a}_i|^2 \delta(\tau - \tau_i) = \sum_{i=0}^{N-1} \tilde{p}_i \delta(\tau - \tau_i) \quad (4.5)$$

while the $\tilde{}$ denotes spatial average of all the 25 individual locations and the $|\tilde{a}_i|^2$ is the spatial average of the i^{th} multipath component power.

4.3 Find out the Maximum Excess Delay

Now the next step is to find out the Maximum Excess Delay $\tau_{\max}$ for each measurement location. In order to do this, we apply a certain threshold to the calculated MIP. Any multipath component which is below the threshold are eliminated from the MIP as false multipath signal caused by noise spikes. In other words, all the $|\tilde{a}_i|^2$ below the noise threshold was set to zero, and the maximum τ_i for the non-zero $|\tilde{a}_i|^2$ is the maximum excess delay $\tau_{\max}$.

First we need to choose a suitable threshold. As we introduced in Section 2.2.6, there are two types of threshold we can choose. One is a certain dB value below the highest multipath signal level which in this case would be a relative value and varies with different measurement locations. Another one is a certain dB value above the noise level so that the threshold would be an absolute value and be the same for all measurement

locations. The second choice is suitable for our channel measurements because we are interested in all the resolvable multipath components not just the few high level signals.

The threshold we chose here is 3 dB above the maximum noise level. The noise level can be calculated by averaging the noise before the first arriving signal on all the MIP plots for all measurement locations. By doing this we have -40 dB as the threshold. Then we force all the $|a_i|^2$ values below this threshold to be zero so the MIP plot as shown in Figure 4.5 now becomes a reduced curve as shown in Figure 4.7. The Maximum Excess Delay $\tau_{\max}$ is the maximum non-zero τ_i so in this MIP curve the $\tau_{\max}$ is 61.4 ns.

Figure 4.7 Time averaged Multipath Intensity Profile (MIP) with threshold 3 dB above noise level

4.4 Calculate the Large-scale Path Loss

The large-scale path loss for a measurement location is the attenuation in dB between the total received signal power P_r and the transmitted signal power P_t .

It is obvious that the Path Gain G we introduced in section 4.1.4 is actually the ratio between the total received UWB signal power at T-R distance d with the reference signal power at calibration distance $d_0 = 1\text{m}$. The Path Gain G is calculated from MIP coefficients $|a_i|^2$ using equation (4.7), so we have the large-scale path loss at the measurement location with T-R distance d is:

$$PL(d) = PL(d_0) - 10\log(G) = PL(d_0) - 10\log\left(\sum_{i=0}^{N-1} |a_i|^2\right) \quad (4.6)$$

Since at the reference location $d_0 = 1\text{m}$ the direct path signal is dominant, so the value of $PL(d_0)$ can be calculated based on a free space assumption from the transmitting antenna to d_0 using equation (2.2) and center frequency $f = 3.5\text{GHz}$ and $G_t = G_r = 0\text{dBi}$, then we have:

$$PL(d_0) = 10\log\left(\frac{(4\pi)^2 (d_0)^2}{G_t G_r \lambda^2}\right) = 10\log\left(\frac{(4\pi)^2 (1)^2}{1 * 1 * (3 * 10^8 / 3.5 * 10^9)^2}\right) = 43.32 \text{ (dB)} \quad (4.7)$$

For the above example of LOS, $d = 3\text{m}$, we calculated the Path Gain G from the MIP data as $G = 0.1407$. So that the path loss is:

$$PL(d = 3\text{m}) = PL(d_0) - 10\log(G) = 43.32 - 10\log(0.1407) = 51.84 \text{ (dB)}$$

So we define this is the measured large-scale path loss at this particular measurement location. Repeat the same process for all other locations we will have the set of measured large-scale path loss data for all the locations.

4.5 Calculate the Mean Excess Delay and RMS Delay Spread

In order to calculate the **Mean Excess Delay** τ_m and the **RMS Delay Spread** τ_{rms} we need to normalize the MIP or PDP first to achieve unit area below the profile curve as we discussed in section 2.2.4 and 2.2.5.

For the above example of LOS scenario, $d = 3\text{m}$, we have the Path Gain $G = 0.1407$. By using the equation (4.8) to (4.12) we calculated the two parameters as below:

Mean Excess Delay $\tau_m = 8.6 \text{ ns}$,

RMS Delay Spread $\tau_{rms} = 13.0 \text{ ns}$

Chapter 5

Large-scale Path Loss Model

This chapter gives the details of the large-scale path loss model development and the model parameters. First the general indoor large-scale path loss model is introduced. Then the large-scale path loss models of this UWB channel are established for both the LOS and NLOS environments. The model parameters such as path loss exponent n , the reference path loss $PL(d_0)$ and the shadowing factor $X\sigma$ are then computed and characterized based on the measurement data.

5.1 Indoor Large-scale Path Loss Model

In this thesis work our main interests are concentrated in characterizing radio propagation inside buildings and the channel is so called indoor radio channel.

The indoor radio channel differs from the other traditional radio channel in at least two aspects. First of all, T-R distances are small, usually in the order of meters or tens of meters. Second, there are strong multipath signals existing and usually with very small differential delays between direct and reflected multipath components. So the small-scale multipath fading usually would be a problem. Also the layout of the building, the construction materials and the arrangement of furniture will strongly influence the signal propagation. So the indoor radio propagation channel needs to be carefully modeled. In

general, indoor radio channels may be classified either as line-of-sight (LOS) or non-line-of-sight (NLOS).

One widely used large-scale indoor propagation model is introduced here. It shows that the indoor path loss obey the distance power law (in dB) as:

$$PL(d) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (5.1)$$

where the value n is the path loss exponent which depends on the signal type and the surroundings and building type, d_0 is the reference distance which is close to the transmitter and far enough for the Rx antenna to be located in the far field range of the Tx antenna. For indoor propagation, we usually choose d_0 as 1 meter. $PL(d_0)$ is the path loss at this reference point. X_σ is the shadowing factor. It's often modeled as a zero-mean Gaussian distributed random variable with standard deviation σ dB.

Note that the indoor propagation environment usually has a strong multipath, so the received signal power may fluctuate when the receiver is moved by only a fraction of a wavelength. In order to mitigate this so called small-scale fading effect, we take a spatial average of received power when dealing with large-scale path loss issues. This spatial average power is computed by averaging signal measurements over many points located in a local area in the order of $5\lambda - 10\lambda$.

5.2 Comparison of two types of Path Loss Data

First let's compare the values of two different types of path loss (PL) we calculated at the measurement locations. One set is based on temporal average and the other is based on spatial averages.

The temporal averaged path loss is calculated from the time averaged Multipath Intensity Profile (MIP). And the MIP is acquired by averaging the 10 individual MIPs calculated from the 10 time snapshots of measured channel responses at the same location.

The spatial averaged path loss, on the contrary, is calculated from the spatial averaged PDP data. And this spatial averaged PDP is acquired by averaging the 25 individual MIPs at the 25 measurement points on the 5x5 grids.

The two different types of path loss values are listed and compared in the Table 5.1 below:

Table 5.1 Path Loss Values calculated from MIP and PDP

	T-R distance d	PL calculated from MIP	PL calculated from PDP	Delta
	(m)	(dB)	(dB)	(dB)
LOS	6	55.01	55.55	0.54
	7	57.54	58.18	0.64
	7	57.35	57.04	-0.31
	7	58.40	57.55	-0.85
	8	57.36	58.09	0.72
NLOS	3	58.88	56.60	-2.28
	4	65.78	66.04	0.26

From the table we can see that the differences between the PL calculated from MIP and the PL from PDP are very small, less than 1 dB. The reason behind this close agreement is because the wide bandwidth (3GHz) makes almost all the multipath components resolvable and so the receiver can collect all these multipath components which contribute to the total received power. This means that the UWB signal suffer far less in terms of small scale fading by comparison to narrow-band signals. In the following chapters in this thesis the temporal averaged MIP data will be used for all the analysis purposes.

5.3 The UWB Channel Path Loss Model Parameters

It is found from the measurement data that the general indoor channel path loss model (5.1) also fits well with our 2GHz – 5GHz indoor UWB channel. We have

$$PL(d) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) + X_\sigma = PL(d_0) + 10n \log(d) + X_\sigma \quad (5.2)$$

where the shadowing factor X_σ is modeled as a zero-mean Gaussian distributed random variable with standard deviation σ dB.

Now we are going to find out the model parameters such as the pass loss exponent n , the reference path loss $PL(d_0)$ and the standard deviation σ from our measurement data. These parameters statistically describe the path loss model for an arbitrary location having a specific T-R separation, and this model may then be used in computer simulation to provide received power levels for random locations in UWB communications system design and analysis.

5.3.1 Path Loss Data

First we calculate all the path loss values for both the LOS and NLOS measurement locations. These are the temporal averaged path loss calculated from 10 individual MIPs at each location. The results are shown in table 5.2 for LOS and table 5.3 for NLOS as the “Measured PL” columns.

Table 5.2 LOS Path Loss Results

	d	Measured PL	Calculated PL	Shadowing Factor
	(m)	(dB)	(dB)	(dB)
LOS_Direction1	2	48.42	48.78	0.36
	3	51.84	51.77	-0.07
	4	54.04	53.90	-0.14
	5	55.82	55.55	-0.28
	6	55.01	56.89	1.89
	7	57.54	58.03	0.49
	7	57.35	58.03	0.68
	7	58.40	58.03	-0.37
	8	57.36	59.02	1.65
LOS_Direction2	1	44.07	43.65	-0.41
	2	49.51	48.78	-0.73
	3	49.93	51.77	1.84
LOS_Direction3	1	43.53	43.65	0.13
	2	49.05	48.78	-0.27
	3	51.84	51.77	-0.07
	4	53.57	53.90	0.33
	5	56.86	55.55	-1.31
	6	59.66	56.89	-2.76
	7	60.46	58.03	-2.43
LOS_Direction4	1	43.50	43.65	0.16
	2	48.46	48.78	0.32
	3	52.08	51.77	-0.30
	4	53.79	53.90	0.11
	5	55.49	55.55	0.05
	6	55.76	56.89	1.14

Table 5.3 NLOS Path Loss Results

	d	Measured PL	Calculated PL	Shadowing Factor
	(m)	(dB)	(dB)	(dB)
NLOS_1	3	58.88	59.90	1.02
	4	65.78	63.86	-1.92
	5	65.35	66.93	1.58
	6	68.59	69.44	0.85
	7	71.20	71.56	0.36
NLOS_2	2	54.26	54.33	0.07
	3	61.05	59.90	-1.15
	4	66.36	63.86	-2.50
	5.1	67.66	67.20	-0.46
	6.1	68.27	69.67	1.40
	7	68.99	71.56	2.57
NLOS_3	8	72.21	73.40	1.19
	9	72.78	75.02	2.24
	10	77.03	76.47	-0.56
	11.5	79.43	78.39	-1.04
	12	80.06	78.98	-1.08
NLOS_4	6.5	68.28	70.54	2.26
	7.5	73.85	72.51	-1.34
	8.5	75.74	74.23	-1.51
	9.7	78.03	76.05	-1.98

Note: The numbers in the “Calculated PL” columns in the above two tables are calculated from equation (5.5) and (5.6) of this chapter, which are the statistical models we presented in this thesis to predict the mean path loss value at a certain T-R distance d of this UWB channel.

5.3.2 Find out the Reference Path Loss and the Path Loss Exponent

The path loss exponent n and the reference path loss $PL(d_0)$ are computed from measured PL data by using linear regression such that the difference between the measured and estimated path losses is minimized in a mean square error sense over the whole range of measurement locations.

We plot the path loss numbers against the T-R distance d as shown in Figure 5.1 below for both the LOS and NLOS scenarios. The least squares fitting curves are also shown in the Figure.

Figure 5.1 Path Loss Curves for LOS and NLOS

From Figure 5.1 we have:

- For the LOS scenario the path loss exponent n is equal to $17.013/10 = 1.7013$, and the reference path loss is found to be $PL(d_0) = 43.654$ dB. This is very close to the free space path loss at 1m which we calculated as 43.32 dB for center frequency 3.5 GHz.

If we take two digits for the values and we have the LOS path loss model as:

$$PL(d) = 43.65 + 10 * 1.70 * \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (5.3)$$

where $d_0 = 1\text{m}$.

- For the NLOS scenario the exponent n is equal to $31.677/10 = 3.1677$ and the reference path loss is found to be $PL(d_0) = 44.791$ dB. Similar, if we take two digits for the values and we have the NLOS path loss model as:

$$PL(d) = 44.79 + 10 * 3.17 * \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (5.4)$$

where $d_0 = 1\text{m}$.

The interesting finding here is that for the LOS case the exponent $n = 1.70$ which is smaller than the free space exponent $n=2$, the reason behind this is because of the collection of all the multipath components so that a higher power is received than the single direct signal in the free space case.

5.3.3 Characterization of the Shadowing Factor X_{σ}

The shadowing factor X_{σ} is defined as the deviation between the calculated (or expected) path loss value and the actual measured path loss value.

From the results in section 5.3.1 we can calculate the expected PL from the two equations below:

$$\text{LOS:} \quad \text{PL}(d) = 43.65 + 10 \cdot 1.70 \cdot \log(d/d_0) \quad (5.5)$$

$$\text{NLOS:} \quad \text{PL}(d) = 44.79 + 10 \cdot 3.17 \cdot \log(d/d_0) \quad (5.6)$$

The deviation numbers for each location is also shown in the “Shadowing Factor” columns in table 5.2 and table 5.3 for LOS and NLOS respectively.

Let’s look at the deviation numbers in the two tables. They show that our models fits very well because the deviation is less than 3 dB from the actual measured value. Also we notice that for some locations close to the wall, the actual path loss is smaller than the calculated path loss from our model. That can be understood as the wall will cause relatively strong reflections which contribute to the total received power.

To characterize the shadowing factor X_{σ} , we first plot the cumulative distribution function (CDF) of the deviations between the measured path loss data and the expected PL data. The deviation data is shown in table 5.2 and 5.3 as shadowing factor. The two plots for LOS and NLOS cases are shown in Figure 5.2 and 5.3 respectively and are compared with the ideal normal CDF curve.

Figure 5.2 CDF of the Shadowing Factor - LOS

Figure 5.3 CDF of the Shadowing Factor - NLOS

The CDFs show that the shadow fading is log-normally distributed (normally distributed in dB) in an approximate sense with R-square values 0.9591 and 0.9553 for LOS and NLOS respectively.

From the data we calculated that the shadowing factor X_σ has zero mean and has the standard deviation σ as following:

- LOS scenario: Mean value = 0dB and standard deviation $\sigma = 1.09$ dB.
- NLOS scenario: Mean value = 0dB and standard deviation $\sigma = 1.57$ dB.

5.4 Summaries and Comparison to Other Research Work

In summary, the large-scale path loss model we derived from measurement data to predict the average path loss at an arbitrary T-R distance d for this UWB indoor channel is

$$PL(d) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (5.7)$$

And the model parameters are listed in Table 5.4

Table 5.4 UWB Channel Large-scale Path Loss Parameters

	PL(d_0)	n	σ
LOS	43.65	1.70	1.09
NLOS	44.79	3.17	1.57

Similar modeling work has been done in recent years by other research people and different models were acquired. A research work to characterize the UWB channel in

residential environments was carried out by S. S. Ghassemzadeh and a statistical path loss model for in-home UWB channels was established [5]. The modeling efforts are based on over 3000,000 frequency response measurements performed in 23 homes in the northern and central New Jersey area.. Measurements were made with both the LOS and NLOS scenarios. The UWB channel is centered at 5GHz and with 1.5GHz UWB bandwidth.

The large-scale path loss model established in [5] as:

$$PL(d) = PL_0 + 10\gamma \log\left(\frac{d}{1m}\right) + S(d) \quad (5.8)$$

and the model parameters are listed in Table 5.5 below.

Table 5.5 UWB Path Loss Model Parameters in [5]

	PL₀	γ	σ of S(d)
LOS	47.0	1.70	1.6
NLOS	51.0	3.50	2.7

Compare equation (5.7) and (5.8), we see that the two models are in identical form. Now we look at the model parameters. The path loss exponents with LOS scenario for the two models are the same, both are 1.7. However, for the NLOS scenario, the path loss exponent is 3.17 as presented by this thesis work and is 3.5 as reported in [5]. Also the shadowing factors are found to be 1.6 with LOS scenario and 2.7 with NLOS scenario, which are bigger than what we have in this thesis work. This is due to the big variance of all the homes with different structure, age, size and clutters.

Another experimental research conducted by J. A. Dabin to analysis the UWB multipath propagation for indoor wireless channels [7]. The UWB channel been characterized here is in the 2GHz to 6GHz band and measurements were performed in a classroom and laboratory sitting in the Faculty Hall building at NJIT. The large-scale path loss model presented in this work is

$$PL(d) = PL(d_0) + 10\alpha \log\left(\frac{d}{d_0}\right) + X_\sigma \quad (5.9)$$

And the model parameters are listed in Table 5.6

Table 5.6 UWB Path Loss Model Parameters in [7]

	PL(d₀)	α	σ
LOS		1.55	1.98

So the model in [6] has the same format as the model in my thesis work. The path loss exponent here is also smaller than 2. This showed the same trend as my results and the shadowing factor is higher but in the same order as the model parameter in my thesis.

Chapter 6

Multipath Delay Spread Analysis

This chapter presents the time domain analysis of the multipath delay spread parameters for this 2GHz to 5GHz UWB channel. These parameters include the Mean Excess Delay τ_m , the RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} . The results of these multipath delay parameters are calculated and analyzed for both LOS and NLOS scenarios.

6.1 Multipath delay

Besides Path loss issues, it's also important to study the multipath delay characteristics of the channel. The reason is that for indoor propagation environment the received signal power is a sum of many signals of different amplitudes, phases and directions. We call them *Multipath Signals*. The experimental results suggest that the multipath signals are strong and will have significant effects on the total received power. So we need to find out the multipath delay characteristics of the UWB channel.

6.2 Calculated Multipath Data

Following the procedure described in section 4.3 to 4.5, we can calculate the Mean Excess Delay τ_m , the RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} for all the LOS and NLOS measurement locations. The results are listed in table 6.1 and 6.2.

Please refer to Figure 6.1 for the measurement locations for all the different LOS and NLOS directions.

Table 6.1 Multipath Parameters for the LOS Channel

	d	$\tau_{\max}$	τ_m	τ_{rms}
	(m)	(ns)	(ns)	(ns)
LOS_Direction1	2	88.0	12.1	10.2
	3	61.4	8.6	13.0
	4	85.2	15.6	21.7
	5	61.2	15.1	12.7
	6	55.2	10.3	12.4
	7	66.0	22.1	12.0
	7	52.4	12.8	12.0
	7	49.6	9.6	12.2
	8	40.0	9.1	10.4
LOS_Direction2	1	72.8	3.1	7.6
	2	60.8	4.9	9.5
	3	75.8	5.1	9.5
LOS_Direction3	1	74.6	2.8	7.8
	2	68.8	4.3	10.2
	3	69.0	7.8	11.7
	4	71.8	6.7	12.2
	5	57.8	10.5	13.8
	6	59.2	11.2	14.1
	7	48.6	15.3	16.5
LOS_Direction4	1	76.8	3.1	8.6
	2	64.6	5.2	11.5
	3	75.4	8.2	15.2
	4	69.0	9.0	14.6
	5	69.0	9.6	13.3
	6	61.2	8.1	11.4
Average Value		65.37	9.21	12.15
Standard Deviation		11.35	4.61	2.98

Table 6.2 Multipath Parameters for the NLOS Channel

	d	$\tau_{\max}$	τ_m	τ_{rms}
	(m)	(ns)	(ns)	(ns)
NLOS_1 (Inside the Lab)	3	59.4	12.6	10.4
	4	47.6	10.9	11.8
	5	63.4	14.4	13.7
	6	62.4	9.0	12.2
	7	56.4	14.4	14.4
NLOS_2 (Inside the Lab)	2	68.6	5.5	7.8
	3	65.6	16.1	13.0
	4	69.0	22.3	17.9
	5.1	62.0	13.2	14.4
	6.1	78.0	16.9	13.7
	7	41.4	12.2	13.5
NLOS_3 (In the Hallway)	8	63.8	4.7	7.8
	9	68.2	5.6	7.2
	10	63.4	15.2	13.6
	11.5	79.4	12.8	14.7
	12	84.6	20.2	17.6
(In the Hallway)	6.5	51.4	3.8	8.5
	7.5	61.8	10.4	13.1
	8.5	59.0	12.6	13.4
	9.7	77.0	18.0	16.8
Average Value		63.9	12.6	12.8
Standard Deviation		8.93	5.02	3.11

Figure 6.1 MINTO 4050 floor plan and measurement locations

6.3 The Maximum Excess Delay $\tau_{\max}$

From the data in table 6.1 and 6.2 we can see that the maximum excess delay $\tau_{\max}$ are in the 60ns to 70ns range. No maximum excess delay above 90ns is observed.

- LOS scenario: The mean value of $\tau_{\max}$ is 65.4ns and standard deviation of $\tau_{\max}$ is 11.35ns.

- NLOS scenario: The mean value of $\tau_{\max}$ is 63.9ns and standard deviation of $\tau_{\max}$ is 8.93ns.

We see that the maximum excess delay for the two scenarios have similar numbers, one is 65.4ns for LOS scenario and 63.9ns for NLOS. This is because of the rich multipath signals exist for both scenarios due to the reflections from the big glass window and from the metal rack sitting besides the wall. Also it is because that we use an absolute threshold (-40dB) as discussed in section 4.3 to compute the maximum excess delay, not using the relative threshold to the highest multipath signal.

The Maximum Excess Delay $\tau_{\max}$ is a very important channel parameter for the design of UWB communication systems. As an example, the OFDM system design has to make sure the Cyclic Prefix of the OFDM signal is larger than the channel Maximum Excess Delay.

6.4 The RMS Delay Spread τ_{rms}

From the results in table 6.1 and 6.2 we can see the following:

- For the LOS scenario, the mean value of τ_{rms} is 12.15ns, and the standard deviation of τ_{rms} is 2.98ns among all locations.
- For the NLOS scenario, the mean value of τ_{rms} is 12.8ns, and the standard deviation of τ_{rms} is 3.11ns among all locations.

The two scenarios yield similar RMS delay spread numbers. Both are between 12ns and 13ns. This is also because of the same rich multipath components contained in the MIPs in both LOS and NLOS scenarios. The relatively stronger first arriving signal in the LOS scenario MIPs makes its RMS delay slightly smaller than the NLOS scenario.

6.4.1 The Distribution of τ_{rms}

Now let's take a statistical look at the τ_{rms} data and plot the CDF of τ_{rms} for both the LOS and NLOS cases in Figure 6.2 and 6.3.

Figure 6.2 CDF of RMS Delay Spread – LOS

Figure 6.3 CDF of RMS Delay Spread – NLOS

From the Figure 6.2 and Figure 6.3 we found out that for both LOS and NLOS scenarios, the values of RMS delay spread τ_{rms} are close to normally distributed over all measurement locations. The R-square values are 0.9852 and 0.9146 for LOS and NLOS respectively. This hypothesis can only be proved in an approximate sense since we only have limited number of measurement locations.

6.4.2 Relation of τ_{rms} and T-R Separation Distance

We then plot the RMS delay spread τ_{rms} against the T-R distance d . We have Figure 6.4 for the LOS scenario and Figure 6.5 for the NLOS scenario.

Figure 6.4 RMS Delay versus T-R distance - LOS

Figure 6.5 RMS Delay versus T-R distance - NLOS

The curves show the relationship between the RMS delay spread τ_{rms} and the T-R separation distance d . It shows that the RMS delay spread τ_{rms} increases as a function of $d^{0.23}$ and $d^{0.35}$ for the LOS and NLOS respectively.

The increase of τ_{rms} with the bigger T-R separation distance can be explained from the nature of the multipath components. When the T-R distance increases, the energy at the higher excess delay multipath components also increase, this will result in bigger RMS delay numbers.

6.4.3 Relation of τ_{rms} and Path Loss PL

We could also look at the relationship between the RMS delay τ_{rms} and the Path Loss.

The plots of Path Loss against RMS delay for both LOS and NLOS scenarios are shown in Figure 6.6.

Figure 6.6 Mean Path Loss versus RMS Delay Spread

The Figure 6.6 shows that the RMS delay spread τ_{rms} increases with path loss. This also can be explained as that the multipath component with longer excess delays usually occur more reflections and so they tend to have lower power levels. And lower power levels mean larger path loss. So a measurement location with larger path loss tends to have

more components with long excess delays which results in a bigger number of RMS delay spread τ_{rms}

6.5 Summary

Measurement data shows that the maximum excess delay τ_{max} for this 2GHz to 5GHz UWB channel is in the range of 60ns to 70ns and no maximum excess delay above 90ns is observed.

Again we compare with the modeling results presented in [5] for the residential environment measurements. For the maximum excess delay τ_{max} , as reported in [5], no significant maximum excess delay above 70ns was observed for a 30dB threshold. So the two experimental works showed similar results as regards to the maximum excess delay τ_{max} .

The results of the RMS delay spread τ_{rms} and its standard deviations are summarized in table 6.3 below and compare with results presented in [5].

Table 6.3 UWB Channel τ_{rms} Values

		Mean value of τ_{rms} (ns)	Standard Deviation of τ_{rms} (ns)
My work	LOS	12.15	2.98
	NLOS	12.8	3.11
Research work in [5]	LOS	4.7	2.3
	NLOS	8.2	3.3

The results show that in this particular lab environment there are more multipath components than the measurement results in [5] which were measured in a home residence environment. The reason for this can be explained as that the MINTO 4050 lab has lots of equipments and metal racks inside the room which serve as good reflectors for the UWB signal and thus have more multipath signals with relatively large excess delays.

Also my measurement results show that the mean values of RMS delay for the LOS scenario and NLOS scenario are close to each other, with 12.15 ns for LOS and 12.8 ns for NLOS, while in research work [5] it showed a significant difference between the two scenarios. This is mainly due to different threshold types used in these two research works. In research work [5], a relative threshold is used which is 30dB below the highest multipath component. Thus, for LOS scenario, since the direct arrive signal is stronger than other multipath components, most of the relatively smaller multipath components with larger excess delays were removed from the RMS delay calculation.

On the contrary, in my thesis work a fixed threshold of 3 dB above noise level was used for both LOS and NLOS scenarios. In this case more multipath components with large excess delay spread were counted for the RMS delay calculation. This is the main reason that I got relatively large RMS delay value for LOS scenario which is averaged at 12.15 ns. As an example to prove this, for the LOS2 direction and $d = 3\text{m}$ location, the RMS delay is calculated to be 9.5ns using the threshold of 3dB above noise level. If we define a new threshold as 30 dB below the highest multipath signal level, the RMS delay is recalculated to be 6.5 ns.

Chapter 7

Multipath Intensity Profile Model

In this chapter we will discuss about the decaying models of Multipath Intensity Profile (MIP) for this indoor UWB channel.

As discussed in chapter 4, we obtained the time averaged MIP curve for each measurement location, and we talked about the large-scale path loss model in chapter 5. Also we calculated and analyzed the important multipath delay parameters such as the Mean Excess Delay τ_m , the RMS Delay Spread τ_{rms} and the Maximum Excess Delay τ_{max} in chapter 6. Now we will look at the inside of a Multipath Intensity Profile (MIP) curve and try to find out the decaying trend of the individual multipath power $|a_i|^2$.

7.1 Multipath Decaying Models

7.1.1 Decaying Trend

We will start with NLOS scenario and then move to the LOS scenario. First let's recall some basic definitions we introduced before. The channel impulse response at a certain receiving location can be written as (2.8)

$$h(\tau) = \sum_{i=0}^{N-1} a_i e^{j\theta_i} \delta(\tau - \tau_i)$$

And the corresponding individual MIP for one single measurement at a certain receiving location can be expressed as (2.10)

$$MIP = |h(\tau)|^2 = \sum_{i=0}^{N-1} |a_i|^2 \delta(\tau - \tau_i)$$

In this thesis, we calculate the temporal average of 10 individual MIPs from the 10 individual snapshots of measurement data at each receiving location and we denote this temporal average result just as MIP. An example of MIP is shown below in Figure 7.1

Figure 7.1 An Example of Multipath Intensity Profile for NLOS Scenario

In order to see the decaying trend of individual multipath power $|a_i|^2$ clearer, we change the threshold here to be 10 dB above the maximum noise level and we also change the plot type to scatter type.

Note this 10dB above noise threshold is different from the threshold we used in the previous chapters to compute the maximum excess delay. We use a higher threshold here is just because that we are now interested in the trend of the amplitudes of the relatively stronger multipath components and are not interested at the low level multipath components which are close to the noise level.

Figure 7.2 Scatter plot of an Example MIP for NLOS Scenario

After we apply this 10dB above noise threshold, we also normalize the excess delay τ_i with the averaged RMS delay spread value that we calculated to be 12.8 for the NLOS scenario in chapter 6. All the $|a_i|^2$ below the noise threshold was set to zero, now the MIP curve will become a new curve as shown in Figure 7.2.

From Figure 7.2 we can see that the individual multipath power $|a_i|^2$ has a general downward trend as the excess delay τ_i increases. Also we see that the multipath components are not obviously clustered. Normally the exponential models are the best fit to the multipath decaying curve as indicated in [8].

7.1.2 The NLOS Model

Here we introduce the following decaying model to characterize the MIP delay profile for NLOS scenario as proposed in [8]:

$$|a_i|^2 = k * s(\tau_i) * \exp[-\hat{\alpha}(\tau_i / \overline{\tau_{rms}})]; \quad \tau_i = i / W, i \geq 0 \quad (7.1)$$

where:

k is a constant. It's a scale factor chosen such that the $|a_i|^2$ sums to the total received signal power at the particular location. So that the sum of $|a_i|^2$ will match with the average large-scale path loss predict by the path loss model for the particular location.

τ_i is the excess delay in ns for the i_{th} multipath component.

$s(\tau_i)$ is a noise-like variation with a mean value = 1 over all the excess delays.

$\hat{\alpha}$ is a dimensionless exponential decay constant.

$\overline{\tau_{rms}}$ is the global average of τ_{rms} over all NLOS location as shown in chapter 6 is 12.8ns.

W is the UWB channel bandwidth, in our case W = 3GHz.

To be easier to calculate the model parameters from our measured MIP data, we rewrite the model (7.1) in dB form as following:

$$P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i \quad (7.2)$$

where,

$P_i = 10 * \log(|a_i|^2)$ represents the dB values of $|a_i|^2$

$K = 10 * \log(k)$ is the dB values of the scale factor k

$\alpha = [10 / \ln(10)] * \hat{\alpha} = 4.343 * \hat{\alpha}$ is a dimensionless exponential decay constant

$S_i = 10 * \log[s(\tau_i)]$ represents the deviation between actual measured P_i and the median value of P_i calculated from the above prediction equation. S_i is found to be a zero-mean, normally distributed random process with standard deviation σ_s [8].

7.1.3 The LOS Model

For the LOS scenario the model will be similar to the NLOS scenario except that for LOS there is a large signal at $i=0$. This is the direct arriving signal and it sums up with the rest multipath components together to contribute to the total received power. So we need to consider both direct arrival and other multipath components and we have the following model for the LOS scenario as proposed in [8]:

$$\begin{cases} \text{For } i = 0, & P_i = A \\ \text{For } i > 0, & P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i \end{cases} \quad (7.3)$$

Where A is the direct arrive signal power. We see that for $i > 0$, the model is exactly the same as NLOS scenario.

7.2 The Model Parameters

7.2.1 Calculation of Exponential decay constant α and the deviation factor S_i

The two important parameters for the above models, both for NLOS and LOS scenarios, are the exponential decay constant α and the deviation factor S_i .

Again we look at the NLOS case first, because for the LOS model the only difference is the “A” factor which does not affect the calculation of the exponential decay constant α and the deviation factor S_i .

For the example shown in Figure 7.2, from the best-fit curve we can find out that the $\alpha = 1.82$ and $K = -16.398$ dB. So the median values of P_i can be calculated from

$$P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) = -16.398 - 1.82 * (\tau_i / \overline{\tau_{rms}}) \quad (7.3)$$

And so S_i can be calculated as the deviation between actual measured P_i from the MIP data and the median value of P_i calculated from the above equation (7.3).

To characterize the deviation S_i , first we plot the cumulative distribution function (CDF) of the deviation between the measured P_i and calculated P_i using equation 7.3. The CDF plot is shown in Figure 7.3 and is plotted versus the normal CDF curve.

Figure 7.3 CDF of the Deviation S_i - NLOS

The CDF plot show that the deviation S_i has zero mean and is normally distributed in dB. For the example showed above the standard deviation of the S_i is $\sigma_s = 1.63$ dB.

7.2.2 The Model for Exponential decay constant α

We now could repeat the same process for all the NLOS and LOS locations and get the exponential decay constant α data listed in table 7.1 and table 7.2 for NLOS and LOS respectively:

Table 7.1 Exponential decay constant α for all the NLOS locations

	d	Measured α	Calculated α	Delta
	(m)			
NLOS_1	3	2.696	2.379	0.317
	4	1.935	2.110	-0.175
	5	1.865	1.900	-0.035
	6	1.573	1.729	-0.156
	7	1.083	1.585	-0.502
NLOS_2	2	3.185	2.760	0.425
	3	2.657	2.379	0.278
	4	1.821	2.110	-0.289
	5.1	1.327	1.882	-0.555
	6.1	0.985	1.714	-0.729
	7	1.699	1.585	0.114
NLOS_3	8	1.746	1.459	0.287
	9	1.982	1.349	0.633
	10	1.598	1.250	0.348
	11.5	1.282	1.119	0.163
	12	1.135	1.079	0.056
NLOS_4	6.5	1.350	1.654	-0.304
	7.5	1.603	1.520	0.083
	8.5	1.453	1.402	0.051
	9.7	1.261	1.279	-0.018

Table 7.2 Exponential decay constant α for all the LOS locations

	d	Measured α	Calculated α	Delta
	(m)			
LOS_Direction1	2	2.107	2.345	-0.238
	3	2.705	2.055	0.650
	4	1.469	1.850	-0.381
	5	1.762	1.690	0.072
	6	1.577	1.560	0.017
	7	1.124	1.450	-0.326
	7	1.166	1.450	-0.284
	7	1.336	1.450	-0.114
	8	2.185	1.354	0.831
LOS_Direction2	2	2.578	2.345	0.233
	3	2.376	2.055	0.321
LOS_Direction3	2	2.492	2.345	0.147
	3	2.307	2.055	0.252
	4	1.966	1.850	0.116
	5	1.484	1.690	-0.206
	6	1.549	1.560	-0.011
	7	1.445	1.450	-0.005
LOS_Direction4	2	2.089	2.345	-0.256
	3	1.587	2.055	-0.468
	4	1.497	1.850	-0.353
	5	1.682	1.690	-0.008

From the above NLOS and LOS data we can plot the exponential decay constant α against the T-R distance d as shown in Figure 7.4 and Figure 7.5

Figure 7.4 Exponential decay constant α vs. T-R distance d

Figure 7.5 Exponential decay constant α vs. T-R distance d

From the two curves in Figure 7.3 and 7.4 we see that the α is a decreasing function of the T-R distance d and we can use the following model to characterize the exponential decay constant α [8]:

$$\alpha(d) = \alpha_0 - 10\gamma \log\left(\frac{d}{d_0}\right) + \varepsilon \quad (7.4)$$

where α_0 and γ are the intercept and slope of the best fit curve, respectively. The α_0 is then the exponent α at reference distance $d_0 = 1\text{m}$. And ε is a zero mean Gaussian random process over d with standard deviation σ_ε . These parameters can be found through least squares fitting to the two curves as shown also in Figure 7.4 and Figure 7.5.

And we then have:

- For NLOS scenario (Figure 7.4) the slope γ is equal to $0.9383/10 \cdot \ln(10) = 0.216$, and the α_0 is found to be 3.41 and we have $d_0 = 1\text{m}$. So the model for NLOS is:

$$\alpha(d) = 3.41 - 10 \cdot 0.216 \cdot \log(d) + \varepsilon \quad (7.5)$$

- For the LOS scenario (Figure 7.5) the slope γ is equal to $0.7143/10 \cdot \ln(10) = 0.164$, and the α_0 is found to be 2.84 and we have $d_0 = 1\text{m}$. So the model for LOS is:

$$\alpha(d) = 2.84 - 10 \cdot 0.164 \cdot \log(d) + \varepsilon \quad (7.6)$$

7.2.3 Characterization of the Deviation Factor ϵ

We can calculate the expected value of exponential decay constant α for different T-R distance d , by using the two equations below:

NLOS: $\alpha(d) = 3.41 - 10 * 0.216 * \log(d)$

LOS: $\alpha(d) = 2.84 - 10 * 0.164 * \log(d)$

So the deviation factor ϵ is then the delta between the calculated (or expected) value and the actual measured value. The deviation numbers for each location is also shown in the table 7.1 and table 7.2 for NLOS and LOS respectively.

We look at the deviation factor ϵ values in the table 7.1 and 7.2. To characterize deviation factor ϵ , we first plot the cumulative distribution function (CDF) of the deviations between the measured data and the calculated data. The two plots for NLOS and LOS cases are shown in Figure 7.6 and Figure 7.7 respectively and are compared with the ideal normal CDF curve.

The CDFs show that the deviation factors are approximately normally distributed. The deviation factor ϵ has zero mean and with standard deviation σ_ϵ as following:

- NLOS scenario: Mean value = 0 and standard deviation $\sigma_\epsilon = 0.35$
- LOS scenario: Mean value = 0 and standard deviation $\sigma_\epsilon = 0.33$

Figure 7.6 CDF of the Deviation Factor - NLOS

Figure 7.7 CDF of the Deviation Factor - LOS

7.3 Summary

In this chapter we discussed about the decaying model of Multipath Intensity Profile for this 2GHz – 5GHz indoor UWB channel.

We found out from the measurement data that the MIP decaying model for the NLOS case of this UWB channel can be expressed as following:

$$P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i \quad (7.2)$$

And accordingly for the LOS scenario the model is:

$$\begin{cases} \text{For } i = 0, & P_i = A \\ \text{For } i > 0, & P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i \end{cases} \quad (7.3)$$

Also we characterized the exponential decay constant α of the above model as:

$$\alpha(d) = \alpha_0 - 10\gamma \log\left(\frac{d}{d_0}\right) + \varepsilon \quad (7.4)$$

And more specifically we then calculated the model parameters for this UWB channel as following:

- For NLOS scenario the slope $\gamma = 0.216$, and $\alpha_0 = 3.41$ so the model becomes:

$$\alpha(d) = 3.41 - 10 * 0.216 * \log(d) + \varepsilon \quad (7.5)$$

- For LOS scenario the slope $\gamma = 0.164$, and the $\alpha_0 = 2.84$ so the model becomes:

$$\alpha(d) = 2.84 - 10 * 0.164 * \log(d) + \varepsilon \quad (7.6)$$

Chapter 8

Conclusions and Future Work

8.1 Conclusions

As presented in former chapters, we have carried out frequency domain channel measurements and performed statistical modeling efforts to characterize an indoor 2GHz to 5GHz UWB channel. The modeling results are summarized in this section.

8.1.1 Large-scale Path Loss

It turned out that the large-scale path loss of this UWB channel could be modeled as:

$$PL(d) = PL(d_0) + 10 * n * \log\left(\frac{d}{d_0}\right) + X_\sigma$$

And the shadowing factor X_σ is found to follow a log-normal distribution with a standard deviation σ dB. The model parameters were calculated for both LOS and NLOS scenarios as below.

Table 8.1 UWB Channel Large-scale Path Loss Parameters

	PL(d₀)	n	σ
LOS	43.65	1.70	1.09
NLOS	44.79	3.17	1.57

8.1.2 Multipath Delay Spread Parameters

The time domain analysis was also performed based on the calculated MIP data. The maximum excess delay $\tau_{\max}$ is found to be in the range of 60ns to 70ns and no significant multipath signal above 90ns is observed. The RMS delay spread is found to be normally distributed over all receiver locations and it increases with T-R distance and path loss.

The values for the RMS delay spread τ_{rms} and its standard deviations are summarized in table 8.2 below:

Table 8.2 UWB Channel τ_{rms} Values

	Mean value of τ_{rms} (ns)	Standard Deviation of τ_{rms} (ns)
LOS	12.15	2.98
NLOS	12.8	3.11

8.1.3 Multipath Intensity Profile Model

We also looked at the inside of a Multipath Intensity Profile (MIP) and discussed about the multipath decaying models for both LOS and NLOS scenarios.

We found out from the measurement data that the multipath decaying model for the NLOS case is as following:

$$P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i$$

And accordingly for the LOS scenario is:

$$\left\{ \begin{array}{l} \text{For } i = 0, \quad P_i = A \\ \text{For } i > 0, \quad P_i = K - \alpha * (\tau_i / \overline{\tau_{rms}}) + S_i \end{array} \right.$$

Also we characterized the exponential decay constant α of the above model as:

$$\alpha(d) = \alpha_0 - 10\gamma \log\left(\frac{d}{d_0}\right) + \varepsilon$$

8.2 Future Work

The work done for this thesis is a good start to characterize indoor UWB channels. In order to gather more information about the UWB signal propagation environment, some possible research work are suggested here:

- Research on the attenuation of UWB signal when going through different type of construction materials.
- Research on the large-scale path loss model when the Tx and Rx antennas are in different rooms and in different levels of the building.
- Simulation on UWB communication system performance using the modeling results presented in this thesis.

Appendix A

Picture of the Measurement Set-up

Photo 1 Measurement Set-up

Photo 2 HP 8753C Vector Network Analyzer

Photo 3 Calibration Kit

Appendix B

Matlab Codes for IDFT processing

```

rootname1 = 'mag';

rootname2 = 'phase';

rootname3 = 'h_abs1';

rootname4 = 'h_angle1';

extension1 = '.xls';

extension2 = '.txt';


for data = 0:9


    open_mag_name = [rootname1, int2str(data), extension1];
    open_ang_name = [rootname2, int2str(data), extension1];
    save_mag_name = [rootname3, '-',int2str(data), extension2];
    save_ang_name = [rootname4, '-',int2str(data), extension2];


    mag1=(load(open_mag_name)).';
    ang1=(load(open_ang_name)).';
    mag2=[10.^((mag1)/20)];
    ang2=[ang1/180*pi];
    mag=[mag2(1:1:801)];
    ang=[ang2(1:1:801)];

```

```
realpart=mag.*cos(ang);  
imgpart=mag.*sin(ang);  
  
path_spectrum=complex(realpart,imgpart);  
  
path_signal=(ifft(path_spectrum));  
  
path_abs=(abs(path_signal)).';  
eval(['save ', save_mag_name , ' path_abs /ascii'])  
  
path_angle=(angle(path_signal)).';  
eval(['save ', save_ang_name , ' path_angle /ascii'])  
  
end
```

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