ABSTRACT

MA, LI. Investigation of Transmission, Propagation, and Detection of UWB Pulses Using Physical Modeling. (Under the direction of Dr. Alexandra Duel-Hallen and Dr. Hans Hallen).

Recent experimental and physical modeling studies demonstrate that, as opposed to systems with smaller bandwidth, the Ultra-Wideband (UWB) channel exhibits frequency-dependent distortion of individual multipath components. This per-path distortion is particularly significant in outdoor UWB applications, where line-of-sight (LOS) or non-distorted reflected signals might not be available at the receiver (for example, in a canyon-like street). In these cases, the dominant propagation mechanisms involve shadowing (diffraction) and reflection by small objects (e.g. signs or lamp-posts). In this dissertation, a physical model is developed to investigate the position-dependent distortion of the UWB pulse. The results indicate that both the shadowed pulse and the reflected pulse (by small objects with dimensions bounded by the wavelengths present in the signal) are distorted. Design of optimal and suboptimal templates for the correlation receiver are investigated. The UWB pulses that accommodate robust template choice given by the transmit pulse shape for all propagation conditions and satisfy the FCC spectral mask for outdoor channels are identified. Finally, we analyze the frequency-dependent propagation gain of the UWB channels in various outdoor conditions. This knowledge quantifies the potential benefits of adapting the transmitted signal to the dominant propagation mechanism.
INVESTIGATION OF TRANSMISSION, PROPAGATION, AND DETECTION OF UWB PULSES USING PHYSICAL MODELING

by

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To my family
BIOGRAPHY

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Chapter 1

INTRODUCTION

1.1 Background

UWB has emerged as a strong candidate for high speed wireless communications since the release of the FCC spectral masks [1] (Fig. 1.1). These masks allow the use of 0~0.96 GHz and 3.1~10.6 GHz bands on an unlicensed basis subject to certain restrictions on the signal power spectrum density (PSD). Specifically, UWB technology is defined as any wireless transmission scheme that possesses a fractional bandwidth $\frac{W}{f_c} \geq 20\%$, where $W$ is the transmission bandwidth and $f_c$ is the center frequency, or an absolute -10 dB bandwidth which exceeds 500 MHz [1]. These systems have many desirable properties, e.g. immunity to multipath propagation, precise geolocation capabilities, and most importantly, the associated low power spectral density that allows coexistence with other wireless systems [2][3][4].

Since the early 90’s, the vast majority of UWB research has focused on impulse radio (I-UWB) techniques that employ transmission of subnanosecond pulses that occupy very wide bandwidth. Unique advantages of impulse radio include potentially low implementation complexity and good capability to penetrate through obstacles [5][6]. Time-Hopping Pulse Position Modulation (TH-PPM) [4][7] and the Direct-Sequence UWB (DS-UWB) [8][9] are examples of impulse radio methods. Recently, multi-band Orthogonal Frequency Division Multiplexing (MB-OFDM) [10][11] that does not utilize UWB pulses has emerged as strong candidate for the Wireless Personal Area Network (WPAN) standard [12]. Currently, two PHY proposals, DS-UWB [13] and MB-OFDM [14], are under consideration by IEEE
802.15 WPAN task group 3a (TG3a), and a decision is still to be made between these two proposals. In addition, IEEE 802.15 Low Rate Alternative PHY Task Group 4a (TG4a) was chartered to develop a low data rate solution with multi-month to multi-year battery life and very low complexity and the potential applications including sensor networks, remote control and home automation [15]. Low complexity I-UWB systems are attractive for the latter proposal.

Figure 1.1 FCC spectral masks for UWB communications.
Recent research has focused on many challenges of UWB transmission, including I-UWB pulse design [4][7][16-27], multiple access techniques [2][4][7-9] [13][14][17][28-30], channel estimation and synchronization [4][7][21][28][31-43], and receiver design [4][7][20] [28][31][38][44-56]. Our work focuses on improving the UWB systems by taking into account their physical channel characteristics. While several models have been proposed for narrowband systems [57-59], they were inadequate for the UWB studies due to their restriction on measurement bandwidth and impulse response resolution. Several measurement campaigns have been performed recently and used to create statistical and deterministic UWB channel models [60-66]. For example, a modified Saleh-Valenzuela (S-V) model [57] was adopted by IEEE 802.15 TG3a [6]. This model is valid only for the high frequency band of the FCC mask (3.1−10.6 GHz), and is designed for indoor residential and office environments with the distance between the transmitter and the receiver restricted to < 10 m [6][67]. Motivated by its targeted applications, e.g. sensor networks, the channel model adopted by IEEE 802.15.4a is more comprehensive [68]. The generic description of the model is provided for 2-10GHz, with the parameterization for different indoor and outdoor environments. However, due to lack of channel measurements, the channel model for the lower band of FCC spectral mask is not detailed in [68], and the models in [60] and [69] are adopted as possible generic channel models for 100-1000MHz. In addition, the structure and parameterization for body-area network (BAN) are also described in [68].

In our study, we focus on the frequency-dependent per-path distortion and the associated propagation loss for various propagation mechanisms. The received UWB signal is subject to multipath propagation [6][70]. Due to the exceptionally wide bandwidth of
impulse radio signals, the frequency independence assumption implied for the impulse response of the individual multipath components does not always hold. While the models used in [2][4][44] do not address this issue (with the exception of the antenna gains), several experimental and physical modeling studies demonstrate that, as opposed to systems with smaller bandwidth, the UWB channel can exhibit frequency-dependent (per-path) distortion [45-46,48,71-73]. For example, the frequency dependency of the path loss associated with diffraction was included in the recent UWB model described in [68]. Note that the effect of per-path distortion is not severe for indoor applications such as those targeted by IEEE 802.15.3a where the LOS and/or reflection paths (by relatively large reflectors, e.g. walls, ceilings and furniture) are often present and dominate the received signal, and the propagation mechanisms associated with these paths do not cause frequency dependent distortion [71]. In our study, we focus on the outdoor UWB channel, where per-path distortion can occur due to shadowing (diffraction), reflections from small objects, rain, trees, etc. For example, while diffracted signals are typically weaker than Line-of-Sight (LOS) or reflected signals (for sufficiently large reflector sizes), they can be dominant in certain outdoor environments, e.g. a canyon-like street or when the receiver antenna is located behind a hill from the transmitter, where the signal arrives at the receiver only after it has been diffracted around an intervening object (and/or after reflection by a small reflector). The power loss in the shadow is often lower than the loss caused by building penetration, thus resulting in more reliable reception in most outdoor environments [74]. The strength of the diffracted signal depends strongly on the wavelength of the signal with larger loss for higher frequencies. Physics-based studies on UWB pulse distortion have been reported in
in which the diffracted pulse is derived directly from expressions of the Uniform Theory of Diffraction (UTD) and Geometry Theory of Diffraction (GTD). The proposed physical model is based on a Fresnel diffraction augmentation of the method of images [76-78], and provides a more accurate description of how the strength and shape of the received pulse changes with position in given local environment, even in the proximity of the reflectors.

In previous research, the effect of per-path distortion on the choice of template for the correlation receiver of I-UWB systems was investigated in [7][45][46]. This problem requires further investigation and is included in this thesis. It is important to address several other issues related to the effect of the propagation mechanism in the UWB channel. In our study, we utilize the developed physical model to jointly design the transmitter and the receiver of UWB systems by taking into account per-path distortion and the associated path loss.

1.2 I-UWB System Fundamentals

Two types of UWB techniques have been proposed: one is based on sending very short pulses to convey information bits, i.e. I-UWB, and the other uses multiple subbands of 500MHz and within each band, OFDM modulation is used, i.e. MB-OFDM. Each technique has its own technical merits and disadvantages. Due to the reasons stated previously, we focus on I-UWB in our study. Some fundamental aspects of I-UWB will be introduced in the remainder of the chapter.
1.2.1 Modulation Schemes

In I-UWB systems, short pulses on the order of nanosecond or less in width are used to form a communication link. But one single UWB pulse does not carry any information itself and we need to add information onto a series of UWB pulses through modulation. Listed below are several modulation schemes for I-UWB. The transmit antenna of UWB systems could distort the pulse fed to the antenna, which is one feature of UWB systems. In our system model, we exclude the antenna effects and assume the distortion introduced by the antenna systems is included in the transmit pulse $p_t(t)$. The introduced pulse modulation schemes are illustrated in Fig. 1.2.

A. On-off keying (OOK)

OOK is a very simple form of pulse modulation, in which the transmission of a pulse represents a data bit ‘1’ and the absence of a pulse represents a data bit ‘0’. An OOK modulated signal can be represented by

$$s(t) = \sum_{m=1}^{M} b_m p_t(t-mT),$$  \hspace{1cm} (1.1)

where $b_m \in \{0, 1\}$ is the $m$-th data bit, $M$ is the total number of bits to be transmitted and $T$ is the pulse repetition interval (PRI). One main advantage of OOK is its simplicity and low implementation cost. In OOK transmitter, a simple RF switch can be turned on and off to represent data bits and in that way, transmit power can be saved while transmitting data ‘0’. The detection of OOK-modulated signal is usually done with a non-coherent energy detector receiver [79]. Despite the simplicity of OOK transmitter, there are some disadvantages associated with OOK modulation. In systems with OOK, an unwanted signal or noise can be
detected as a data bit ‘1’ falsely, so it is very sensitive to noise and interference. The synchronization of UWB system with OOK is very difficult due to the possible ‘silent’ period associated with data bit ‘0’ [20]. Fortunately, energy detectors are less sensitive to synchronization errors [21].

B. Pulse amplitude modulation (PAM)

PAM encodes the data bits onto different pulse amplitudes, e.g. a pulse with higher amplitude represents a data bit ‘1’ and a pulse with lower amplitude represents a data bit ‘0’. A PAM modulated signal can be represented by

\[
s(t) = \sum_{m=1}^{M} A_m \ p_t(t-mT),
\]

(1.2)

Where \(A_m\) is the specific pulse amplitude (level of power) for the \(m\)-th data bit. Actually, OOK can be redeemed as a special case of PAM where \(A_m = b_m, b_m \in \{0, 1\}\). Another special case of PAM is bi-phase modulation (BPM), in which the polarity of the pulse carries the data information. In bi-phase modulation, a pulse with positive polarity, i.e. \(A_m = 1\), represents a data bit ‘1’ and a pulse with negative polarity, i.e. \(A_m = -1\), represents a data bit ‘0’. PAM modulated signal can be detected with an energy detector receiver or a conventional matched filter (MF) receiver. PAM modulated signal is also very sensitive to interference and noise. Another disadvantage of PAM is the spectral lines present in its power spectral density (PSD) due to the periodicity of the transmitted pulses [7]. These discrete lines can cause harmful interference to other systems within the same frequency band and it can also correlate with strong narrow-band signals and cause interference to UWB signals. Therefore, it is very important to ‘flatten’ these spectral lines by choosing the
right modulation or special spectrum-whitening techniques [4][7][22]. PAM modulated signals can be detected by energy detector or correlation receiver. The correlation template of the receiver should be chosen very carefully due to the per-path distortion problem mentioned previously.

C. Pulse position modulation (PPM)

PPM is generally adopted in the literature. In PPM, information bits are encoded onto the position of the transmitted pulses by shifting the pulse in a predefined window. A PPM modulated signal can be represented by

\[ s(t) = \sum_{m=1}^{M} p(t - mT - b_m \delta) , \]

(1.3)

where \( b_m \in \{0, 1\} \) is the \( m-th \) data bit and \( \delta \) is the modulation index that provides a time shift to differentiate data bit ‘0’ and ‘1’. \( M \) is the maximum number of transmitted bits, \( \delta \) is an important design parameter. PPM with a value of \( \delta \) greater than the width of \( p(t) \) belongs to orthogonal modulation. Compared to OOK and PAM, PPM modulated signals are less sensitive to the channel noise. This is because the pulse that represents different data bits has the same amplitude, so the probability of detecting a false data bit is reduced. PPM transmitted signal are usually demodulated with a correlation receiver and again, the received UWB pulse can be distorted, which makes it challenging to choose an appropriate correlation template. One drawback of PPM is its sensitivity to timing synchronization. Because data bits are recovered exclusively based on their exact position in time, timing uncertainties, e.g. jitter and drift, can cause synchronization errors that cause increased error probability.
D. Pulse shape modulation (PSM)

In PSM modulation, data bits are encoded onto the shape of the transmitted pulse, e.g. \( p_1(t) = p_1(t) \) when ‘1’ is to be transmitted and \( p_0(t) = p_2(t) \) when ‘0’ is to be transmitted. Orthogonal pulse modulation (OPM) [80] is a subset of PSM. In OPM, the pulse shapes representing different data symbols are orthogonal to each other, which could help reduce the inter-symbol interference and in multiple-user case, this orthogonality can be used to avoid

![Comparison of different pulse modulation schemes.](image)

Figure 1.2 Comparison of different pulse modulation schemes.
multiple access interference (MAI) [23]. Modified Hermite function based pulse (MHP) [17][18][26] and prolate spheroidal wave function based pulse (PSWF) [19][24] are two pulse families that have been used to achieve OPM. In Fig. 2.2, 2\textsuperscript{nd} order MHP and 3\textsuperscript{rd} order MHP are employed to realize PSM.

1.2.2 Multiple Access Techniques

In I-UWB system, multiple users transmit information simultaneously and independently over a shared bandwidth. Therefore, the received signal at any desirable receiver is likely to be the superposition of all users’ signal. A multiple access technique has to be implemented to surpass the MAI, i.e. the signal from all other users other than the desired transmitter. Various multiple access techniques have been proposed for I-UWB [2][7][8][9][17][28]. Among all these techniques, time hopping (TH) is the most common technique found in the I-UWB literature.

A. Time Hopping (TH) multiple access

In I-UWB, the pulse duty cycle is very small, i.e. the pulse duration is extremely short with respect to the PRI. This is actually the reason for the high processing gain (PG) of I-UWB systems [29]. This low duty cycle can be utilized to accommodate multiple users. To eliminate the catastrophic collision in multiple access, each transmitter is assigned a distinct pulse shift pattern, i.e. \( \{c_m^{(k)}\} \) for the \( k-th \) user, which is called time hopping code [2][7]. The periodic pseudorandom code \( \{c_m^{(k)}\} \) has period \( N_p \), and each code element is an integer within the range \( 0 \leq c_m^{(k)} \leq N_h \). Then the additional time shift to the \( m-th \) transmitted monocycles (from the \( k-th \) user) is \( c_m^{(k)} \cdot T_c \) seconds, \( T_c \) is the chip duration. \( N_h \cdot T_c \) must be
smaller than $T$, which is the PRI. As an example, the transmitted signal for the $k$-th user with PPM and TH multiple access is given by

$$s^{(k)}(t) = \sum_{m=1}^{M} p(t - mT - c_m^{(k)} T_c - b_{k,m} \delta) ,$$  \hspace{1cm} (1.4)$$

where $b_{k,m} \in \{0,1\}$ is the $m$-th date bit for the $k$-th user, $p(t)$ is the transmit pulse and $\delta$ is the modulation index for PPM. Although TH multiple access is often used in conjunction with PPM [2][4][7], it can be combined with various modulation schemes, e.g. PPM, PAM, OOK and BPM [30]. In addition to provide multiple access, this pseudorandom *time hopping code* also helps flatten the discrete spectral lines in the PSD of the uniformly spaced pulse train by randomizing the pulse positions [7].

**B. Direct-Sequence (DS) multiple access**

In DS-UWB, the multiple access is achieved by code division multiple access (CDMA). Unlike conventional Direct-sequence spread spectrum (DS-SS) systems, the use of the spreading codes in DS-UWB systems is solely for accommodating multiple users because the short pulses already realize spreading spectrum over a very large bandwidth on the order of GHz [8][28]. In DS-UWB, each data symbol is represented by a series of chips and each of these chips is a UWB pulse. It is straightforward to combine DS multiple access with bi-phase modulation [8][9] and a DS-UWB system using bi-phase modulation has been proposed to IEEE 802.15 WPAN task group 3a (TG3a) [13]. In [8], the transmit signal for the $k$-th user is

$$s^{(k)}(t) = \sum_{m=1}^{M} \sum_{n=0}^{N-1} b_{k,m} c^{(k)}(n) p(t - mT - nT_c) ,$$  \hspace{1cm} (1.5)$$
where $c^{(k)}(n) \in \{+1, -1\}$, $N$ is number of chips (pulses) per data bit and $T_c$ is the chip duration. Note that $T_c$ must be greater than or equal to the pulse duration and $N \cdot T_c = T$. $b_{k,m} \in \{+1, -1\}$ is the $m$-th data bit for the $k$-th user and $p(t)$ is the transmit pulse. In [9], it is found that the DS-UWB system proposed in [13] is found to be more robust to imperfect channel estimation than the MB-OFDM system proposed in [14].

C. Orthogonal Pulse multiple access

In [17], the orthogonality between MHP of different orders is utilized to accommodate multiple users.

1.2.3 Channel Characteristics and Modeling

UWB has attracted great research interest since early 90’s. The potential benefits associated with the extraordinarily wide bandwidth includes accurate position location, better ability to combat fading, high multiple access ability and possible easier material penetration [6]. In traditional spread spectrum (SS) systems, e.g. W-CDMA and IEEE 802.11, the system bandwidth is no more than 20MHz, while UWB signaling occupies a bandwidth up to 7.5GHz (the 3.1~10.6GHz band under FCC spectral masks [1]). This extremely wide bandwidth results in some unique characteristics of UWB channels, i.e. very fine multi-path resolution [6][71], frequency dependent path loss and possible per-path distortion of the received pulse [45][48][67]. The fine multi-path resolution on the order of nanosecond or less mitigates the fading problem associated with narrowband or traditional SS systems, in which each resolvable multiple path is actually the superposition of a large number of unresolvable multi-path components (MPC) [6][70]. While the UWB signals can resolve multiple paths with delay difference on the order of nanosecond or less and within such a short period, only
few MPC overlaps, thus fading is no longer a major problem with UWB systems. However, due to the increased number of resolvable multi-paths, it is more difficult to collect multi-path energy for reliable detection, which is a major challenge for UWB receivers [31][71].

The frequency dependency of the path loss associated with diffraction paths has been reported in [72][73], and this feature has been included into the channel model by IEEE 802.15 TG4a [68]. The frequency dependency of the path loss implies potential distortion to the transmit UWB pulse. Due to the extraordinarily wide bandwidth the transmit pulse occupies, different frequency components of the pulse could undergo considerably different propagation loss and the spectral structure of the transmit pulse changes, which in turn means distortion in pulse waveform [48][70]. The distortion associated with the received pulse poses difficulty in choosing an appropriate correlation template for the correlation receiver or Rake receiver, which is the most common receiver cited in UWB literature [4][7][28][31][49]. Pulse distortion is not so severe for indoor applications such as those targeted by IEEE 802.15.3a [71] since a LOS path or reflection path (by relatively large reflector, e.g. walls, ceilings and big furniture) is often present and the associated path loss is frequency independent, while in certain outdoor environments, e.g. a canyon-like street, diffraction path can be dominant, thus making the pulse distortion problem more severe.

In order to design UWB systems and fairly compare different proposals, a common channel model must be created. Several models have been proposed for narrowband systems [57][58][59]. However, these models were inadequate for the UWB system studies due to their restrictions on measurement bandwidth [60]. For instance, in narrowband models, the amplitude of each path is usually modeled as Rayleigh distribution, while we have previously
discussed that this is not the case in UWB systems. For establishing a UWB channel, UWB measurement campaigns have been performed since a few years ago [60]. Two different measurement techniques are often used, i.e. frequency domain sounding technique and time domain (pulse) sounding technique [61]. In frequency domain sounding, a vector network analyzer is used to measure the channel frequency response over a given band at certain sampling rate and the channel impulse response is obtained using Inverse Fourier Transform (IFT). In [60][65][66], measurements have been performed based on frequency domain sounding technique. In pulse sounding, the response of the channel to a UWB pulse is directly measured in the time domain, and the impulse response (IR) of the channel can be recovered from these measurements by deconvolving the transmitted pulse from the channel output [60]. Time domain sounding technique is employed by [62][63][64]. In such case, the resolution of the obtained IR is limited by the bandwidth of the transmitted pulse.

Channel measurement results are used to determine the parameters in statistics channel models or to evaluate how well a statistics channel model fits the measurement data. The tap-delay line Rayleigh fading model [81], the Saleh-Valenzuela (S-V) model [57] and the Δ-K model described in [59] are three channel models considered by IEEE 802.15 TG3a and the model finally adopted was based on a modified S-V model that seemed to best fit the channel measurements [6]. This adopted model is valid for the high frequency band (3.1 – 10.6 GHz) only, and is designed only for indoor residential and office environments, and the distance between transmitter and receiver is restricted to < 10 m [6][67]. Motivated by its targeted applications, e.g. sensor networks, the channel model adopted by IEEE 802.15.4a is more comprehensive [68]. In [68], the generic channel model for 2-10GHz is given and the
parameterization of this model for different indoor and outdoor environments is provided. However, due to lack of channel measurements, the channel model for 100M-1000MHz, which corresponds to the lower band of FCC spectral masks, is still missing in [68], and the models suggested by [60] and [69] are adopted by [68] as possible generic channel models for 100-1000MHz. In addition, the structure and parameterization for body-area network (BAN) are also described in [68]. In some sensor network applications, communications between devices located on the human body might be needed and [68] suggests that modifications to indoor/outdoor channel models are necessary in order to accurately model BAN scenario.

1.2.4 Channel Estimation and Synchronization

Channel estimation is a core issue for receiver design in any wireless communication system. One of the most attractive features of UWB is its ability to resolve multiple paths. A correlation or Rake receiver can be employed in UWB systems to exploit the multipath diversity [4][7][28][31]. However, such a receiver requires knowledge of multipath delays and amplitudes. In UWB systems, the channel parameters, i.e. path delays and amplitudes, are usually derived jointly from the received waveform using a maximum likelihood (ML) approach [31][33][35]. Note that in these references, the multi-path channel model is assumed and the per-path distortion is ignored. In [31], an isolated pulse is transmitted and the received waveform is recorded for channel estimation, while in [33][35], the information-bearing received waveforms are employed in channel estimation. In these references, a channel with $L$ branches is used to approximate the actual channel based on the received waveform. Apparently, the degree of matching depends on the predetermined value of $L$ and
the minimum value of $L$ required for a good matching would suggest the number of fingers a Rake receiver must implement in order to efficiently exploit the multipath diversity [33][38]. Both data-aided (DA) and nondata-aided (NDA) schemes are proposed in [33], and the results show that the performance of DA is superior to that of NDA. The computational complexity associated with DA is less, while NDA is useful in broadcast networks where training sequence would impede the transmitter when the new users enter the network [33].

In I-UWB systems, the accurate information on the arrival times of the incoming pulses must be provided, and a timing error of a small fraction of pulse width could cause dramatic performance loss [33][34]. (For MB-OFDM UWB systems, timing error can cause intersymbol interference (ISI) that destroys the orthogonality of the OFDM subcarriers and degrades the system performance [36].) As in conventional systems, the timing recovery can be viewed as a two-part process [37]. The first part consists of estimating the beginning times of individual frames, which is referred as frame timing. The second part consists of identifying the first frame of each symbol, which is referred as symbol timing. The frame timing is often accomplished by looking for the peak correlation between the received waveform and the local template, and different search strategies are proposed aiming at minimizing the mean acquisition time [39][40][41]. The symbol timing can operate in both DA [42] and NDA mode [43], and DA improves acquisition time as compared with NDA. Accurate timing can be achieved with frame-rate sampling, while channel estimation, which is necessary for coherent receivers, requires sub-pulse sampling rate [33][37].
1.2.5 Receiver Design

One of the most attractive features of UWB is its ability to resolve multiple paths. As the most common receiver cited in UWB literature, a Rake receiver can be employed in UWB systems to exploit the multipath diversity [4][7][28][31][49]. However, the implementation of a Rake can be extremely complex due to the large number of fingers needed to capture a significant part of the signal energy [31]. In addition, the distortion associated with each received pulse will pose difficulty for choosing an appropriate correlation template for the correlator [45][46][48] and the ‘mismatch’ between the received pulse and the correlation template could lead to low output signal-to-noise ratio (SNR) from the correlators. Thus, knowledge on how the channel will distort the transmit pulse would help when implementing such a receiver consisting of a series of correlators.

Some research has been done on the template design. Most of the proposed schemes fall into the following two categories: (1) Given the received signal r(t) and an initial template waveform, apply Minimum Mean Square Estimate (MMSE) directly to get the optimal template [38][55]. This process can be iterative and is often combined with the MMSE estimate of $c_j$ and $\tau_j$, which are amplitude and delay for the $j$-th path. The total number of path, $L$, is usually pre-selected. (2) Use some set of elementary waveforms to construct the template at the receiver. In [46], it is shown that the received pulse over any channel must lie in the space spanned by the transmitted waveform, all derivatives of the transmitted waveform, the Hilbert transform of the transmitted waveform and all derivatives of its Hilbert transform. By finding the least square solution to the received waveform restricted to this space (or some subspace of it), narrow band interference (NBI), which is
orthogonal to the space, can be eliminated. In [56], an idea similar to [46] is presented. However, the possible per-path distortion associated with the received pulse was not considered in [56] and it focused on constructing arbitrary transmit pulse with a given set of elementary waveforms. In addition, a different set of orthogonal elementary waveforms was employed.

Instead of trying to find the optimal template, some transmitter-aided schemes aiming at helping the receiver capture the received energy with lower complexity are developed, e.g. differential detection (DD) [50][51] and transmitted reference (TR) schemes [52][53][54]. In DD systems, the symbols are differentially encoded, and each received data pulse is correlated with the previous one, which serves as the template. In TR schemes, an unmodulated reference pulse is sent prior to each data pulse, and the channel response to the former is exploited as a template for the latter. The performance a TR system is degraded due to the use of template ‘polluted’ by interference and noise, and this problem can be mitigated by averaging the channel responses from several reference pulses [50][53]. No channel estimation will be needed in system with DD and TR schemes.

In addition, OOK-modulated signal is usually detected with a non-coherent energy detector receiver [79]. In such a system, an unwanted signal or noise can be detected as a data bit ‘1’ falsely, so it is very sensitive to noise and interference. Compared to correlation receiver, energy detectors are less sensitive to synchronization errors [21].
1.3 Outline of the Thesis

This thesis focuses on four topics, i.e. physical modeling of UWB channels, pulse distortion, correlation template design and frequency-dependent propagation gain, and the thesis is organized as follows.

In chapter 2, we describe our physical model and present several modeled channel frequency responses. Discussion on frequency responses associated with different propagation mechanisms is also provided in this chapter.

In chapter 3, the channel frequency responses introduced in chapter 2 are utilized to study the pulse distortion under different channel conditions. We discuss the effect of distortion on several families of pulses that are widely adopted in I-UWB literature.

In chapter 4, we propose some suboptimal correlation templates and identify pulses that fit the FCC spectral mask and perform well under all investigated channel conditions with simple and robust correlation templates.

In chapter 5, we compute the frequency-dependent propagation gain for varying channel conditions using pulses that occupy disjoint bands of the spectral mask, and discuss the potential benefits of adapting the transmitted signal to the dominant propagation mechanism.

Finally, in chapter 6, we propose some open problems and future work.
Chapter 2

PHYSICAL MODELING OF UWB CHANNELS

2.1 Introduction

The frequency dependency of the path loss implies potential distortion to the transmit UWB pulse. Due to the extraordinarily wide bandwidth the transmit pulse occupies, different frequency components of the pulse could undergo considerably different propagation loss and the spectral structure of the transmit pulse will be changed, which in turn means distortion in pulse waveform [48][70]. The distortion associated with each received pulse poses difficulty in choosing an appropriate correlation template for the correlator [45][46][48] and the ‘mismatch’ between the received pulse and the correlation template could lead to low output signal-to-noise ratio (SNR) from the correlators. Thus, knowledge on how the channel distorts the transmit pulse would help when implementing such a receiver consisting of a series of correlators.

A realistic physical model was initially developed to test the performance of the adaptive transmission aided by the long-range fading channel prediction algorithm (LRP) algorithm for multipath fading channels [76][82][83]. The model was motivated by the observation that algorithms that predict the wireless channel for up to a few wavelengths ahead cannot be adequately tested with the Jakes model [84] or other existing fading channel models. Moreover, ray-tracing or finite difference time domain (FDTD) methods do not provide insights into the relationship between the reflector configurations and the performance of the long-range prediction, making identification of typical and challenging
test cases difficult. A novel physical model was developed to: (i) create non-stationary datasets to test the adaptive long range prediction algorithm, which enables practical realization of adaptive transmission techniques; (ii) classify the reflector geometries that will have the typical or the most severe parameter variations, so that the reflector configurations for test datasets can be appropriately chosen, (iii) illuminate the origins of the temporal and statistical properties of measured data. It has been demonstrated that the LRP algorithm performs similarly on channels given by the physical model or actual measured data, with significant deviation from performance indicated by the Jakes model. Moreover, the insights of the model accurately described the performance of the algorithm in several scattering environments. This model is used as foundation for developing the physical UWB channel model in our UWB study.

Physics-based studies on UWB pulse distortion have been reported in [45][48][75], in which the diffracted pulse is derived directly from expressions of the Uniform Theory of Diffraction (UTD) and Geometry Theory of Diffraction (GTD). Our proposed physical UWB channel model is based on a Fresnel diffraction augmentation of the method of images [76-78], and provides a more accurate description of how the strength and shape of the received pulse changes with position in given local environment, even in the proximity of the reflectors.

2.2 The UWB Physical Channel Model

The physical model is based on the Fresnel diffraction theory combined with the method of images [76][77][78]. The model calculation requires a specified geometry including the locations, sizes, and orientations of the reflectors, and the transmitter/receiver...
positions. Variations of these parameters in time are allowed, so that realistic scenarios can be considered. Motion of the transmitter, the receiver, or a reflector, distances from the mobile to the reflectors, as well as the sizes and the shapes of the reflectors (flat or curved) affect the gains and the delays of the dominant propagation paths. These variations were utilized in previous research to test the tracking ability of the long-range prediction algorithm [76][82][85]. For example, the phase change of each reflected component induced by motion is frequency dependent. We have used this property as a challenging test case for prediction in frequency hopping and other wideband systems [86][87][88].

The physical model produces the frequency response of the simulated wireless channel for diverse propagation conditions. The received signal is obtained by computing the Inverse Fourier Transform (IFT) of the product of the Fourier Transform (FT) of the transmitted signal with the modeled channel frequency response. Fig. 2.1 shows a simple example of input geometry that contains only one reflector. An aperture for the transmitter is used to model regions in line of sight and those shadowed from the transmitter. Three propagation mechanisms, the line-of sight (LOS) (path 2), diffraction (path 1) and reflection (path 3) are shown in the figure.

To perform a general model calculation, the method of images is used to determine the positions of effective sources. Fresnel diffraction through an aperture that defines the reflector size yields the contribution at the receiver for each reflector. The complex sum of the electric field contributions from the transmitter and all reflectors is the channel frequency response. The Fresnel reflection from reflectors gives an accurate representation of the channel as a function of both the wavelength and position, even close to a small reflector or
the edge of a large reflector. It is more accurate than typical forms of the generalized diffraction model [45][48][75][89], and models based upon Fraunhoffer diffraction [77], which is only accurate far from the reflectors or building corner. The frequency response of the channel is computed for the frequency range specified by the wireless application. Previously, we have concentrated on relatively narrowband systems where the amplitudes of the frequency responses for individual paths were almost flat for all propagation mechanisms, and the change in the carrier frequency primarily affected the measured delay response of the channel [76][86][87]. However, these amplitudes can vary significantly when the model is extended to the UWB frequency range.

![Figure 2.1 A simple Geometry for the UWB physical model.](image)

**2.3 Modeled Channel Frequency Responses and Discussions**

While the LOS channel does not distort the transmitted pulse [70], the amplitudes of the frequency responses of the diffracted and reflected paths (when the size of the reflector is
on the order of the wavelength or less) shown in Fig. 2.2 and Fig. 2.3 indicate that significant distortion is present over the frequency range of the UWB channel [90]. We note that these responses are calculated from the ratio of the electric field at the receiver antenna to the electric field 1 m from the transmitter [60][64][70] and do not include the response of either antenna. As illustrated in Fig. 2.2, high frequency components are strongly attenuated in diffracted channels. We observed that the slope of the amplitude of the channel transfer function resembles that of $H(f)=1/\sqrt{f}$ for deep shadowing, while the integrator with response $1/f$ provides satisfying intuitive explanation for the distortion associated with diffracted channels (Fig. 2.3). The latter two responses were reported for diffracted channels in [48][75], with the $1/\sqrt{f}$ representing the most common diffraction case in outdoor applications - diffraction by a wall corner of a building. As the receiver moves from the diffracted region to the LOS region, the shape of the frequency response becomes gradually flatter as shown in Fig. 2.2.

For the reflection by small reflector (Fig. 2.4), the low frequencies experience great loss, and the slope of the response is modeled well by that of $H(f)=\sqrt{f}$, while the derivative $H(f)=f$ is a simple approximation to this channel (Fig. 2.5). As the reflector size increases, the channel response in Fig. 2.4 approaches the ideal flat channel response. When the reflector size is 20m or greater, the received and the transmit pulses are indistinguishable. For all path responses, the phase response is approximately linear, with its derivative representing the path delay of the corresponding multipath component in the received signal.
Figure 2.2 Amplitudes of frequency responses of distorted UWB channels: the diffraction path (path 1) at different positions in Fig. 2.1.
Figure 2.3 Amplitudes of frequency responses of the diffraction path (path 1) at (10, 20), the integrator $H(f)=1/f$, and $H(f)=1/\sqrt{f}$. 
Figure 2.4 Amplitudes of frequency responses of distorted UWB channels: the reflection path (path 3) for different reflector sizes, coordinate (10, 20) in Fig. 2.1.
Figure 2.5 Amplitudes of frequency responses of the reflection path (path 3) at (10, 20)
for reflector size=0.3m, the differentiator $H(f)=f$, and $H(f)=\sqrt{f}$.
2.4 Modeled Channel Impulse Response

In order to get the channel impulse, we need to take the Inverse Discrete Fourier Transform (IDFT) of the modeled frequency response. When we do an N-point IDFT, we are in effect multiplying an infinite run of sampled data by a window of width $N\Delta f$, where $\Delta f$ is the sampling interval in frequency domain, and the amplitude of the window is unity. In other words, the data are windowed by a square window function. By the convolution theorem and the properties of the Fourier Transform, the IFT of the product of the data with a square window function is equal to the convolution of the data’s IFT with the window’s IFT.

It is well known that the sinc function and the rectangular function are a Fourier Transform pair and one significant characteristic of the sinc function is it has oscillatory side lobes which fall off rather slowly. This tells us that if we apply IDFT directly to the sampled frequency response, there will be a significant leakage between adjacent points of the resulting channel impulse response. One way to mitigate this problem is the use of data windowing. Applying data windowing gives us an opportunity to trade the ‘less leakage’ with ‘resolution’. Reduced resolution and leakage are two primary effects brought by data windowing. The resolution is influenced primarily by the width of the main lobe of the IFT of the window, while the degree of leakage depends on the relative amplitude of the main lobe and the side lobes. The main-lobe width and the relative side-lobe amplitude is a trade-off and different windows make different trade-offs between these two characteristics. The rectangular window inherited with IDFT can be regarded as a window which gives the best resolution in time domain and the worst (most serious) leakage at the same time.
Rectangular window, Gauss window, Hamming window, Blackman window and Kaiser window are windows with high to moderate resolutions, while Nuttall window and Blackman-Harris window are low-resolution windows [96]. Among these windows, the Kaiser window allows control of the trade-off between the main-lobe width and the highest side-lobe level by adjusting the value of the window parameter, $\alpha$. The Kaiser window is defined by the formula [96]:

$$w_k = \begin{cases} 
\frac{I_0(\pi \alpha \sqrt{1 - (2k/N - 1)^2})}{I_0(\pi \alpha)} & \text{if } 0 \leq k \leq N \\
0 & \text{otherwise}
\end{cases}$$

(2.1)

where $I_0$ is the 0th order modified Bessel function, $\alpha$ is an arbitrary real number that determines the shape of the window, and the integer $N$ gives the length of the window ($N+1$ points). With $\alpha = 0$, the window Rectangular window, and a larger value of beta widens the main lobe and decreases the amplitude of the side lobes (leakage). Thus, the Kaiser window is a good general-purpose choice.

Fig. 2.7 shows the amplitude of the frequency response of the UWB channel seen at position (40, 0.1) in Fig. 2.6. Note that this UWB channel consists of two paths, i.e. a LOS path (path 1 in Fig. 2.6) and a reflection path (path 2 in Fig. 2.6, reflector size is 3m). We have discussed previously that neither LOS nor reflection (by reflector with large dimension compared to the signal wavelength) introduces frequency-dependency. That explains why the amplitude of the frequency response is flat for the given frequency band, 0.1GHz~100GHz.

Fig. 2.8 shows the channel impulse response obtained from the channel frequency response shown in Fig. 2.7 without using any data windowing and Fig. 2.9 shows the channel
impulse response derived from the same set of modeled frequency response with a Kaiser window of $\alpha=5.1$. Comparing Fig. 2.8 and Fig. 2.9, one can see that the use of the Kaiser window reduces the leakage (smear) between adjacent points considerably.

Fig. 2.9 indicates that two paths are present in the channel. The path with delay of 166.7 ns corresponds to the LOS path, which has a path length of $\sqrt{(50^2 + 0.1^2)} \approx 50.0001$ m. The other path with delay of 167 ns corresponds to the reflection path, which has a path length of $50+0.1=50.1$ m. The reflection path is about 20 dB weaker than the LOS path, which is due to the small reflectivity, 0.1. The reflectivity for all reflectors is 1 in our study, unless otherwise stated.

Figure 2.6 An input geometry (for Fig. 2.7, Fig. 2.8 and Fig. 2.9). Reflector size is 3 m, reflectivity=0.1 (the default value of reflectivity is 1).
Figure 2.7 Amplitude of the frequency response of the UWB channel: the LOS path (path 1) and the reflection path (path 2), coordinate (40, 0.1) in Fig. 2.6.
Figure 2.8 The amplitude of a derived channel impulse response, no window is used.
Figure 2.9 The amplitude of a derived channel impulse response, using Kaiser window with $\alpha=5.1$. 
Chapter 3

UWB PULSE DISTORTION

3.1 Introduction

We have indicated previously that the received UWB pulses can be severely distorted and the performance of the correlation receiver can be degraded due to the distortion. Thus, it is desirable to know how the UWB channel will distort the received pulse. In this chapter, we will utilize the UWB physical model described in chapter 2 to investigate the effect of distortion on several families of pulses that are widely adopted in I-UWB literature.

In I-UWB communications, there is usually no carrier frequency and the short duration of the pulses directly generates the extremely wide bandwidth on the order of gigahertz. The power spectrum of a UWB signaling is determined by the spectral shape of the transmit pulse $p(t)$, so the UWB pulse design problem is equivalent to designing the basic pulse $p(t)$ to meet all the system requirements [25]. The PSD of any UWB signaling must comply with the FCC spectrum masks [1] in order to protect the existing systems from UWB interference. Another important issue to consider while designing (choosing) a UWB pulse is the per-path distortion associated with the received pulse. According to our study on pulse distortion, the received pulse could be much wider than the transmit pulse, which can degrade the performance of the correlation receiver and potentially cause inter-pulse interference if not accommodated properly. In this chapter, we will first provide the system model and define SNR capture, $\text{SNR}_c$, which will be used in chapter 4 to measure how well the correlation template matches with the (distorted) received pulses. Then we will introduce
several pulse families, i.e. Gaussian monocycles, MHP, modulated MHP and PSWF for their popularity, and use our physical model to investigate their associated pulse distortion. In addition to these pulse families that have been widely employed in the UWB research, other pulses have also been proposed recently [25][27].

3.2 UWB System Model and Receiver Signal Processing

The multipath UWB channel impulse response can be modeled as [46]:

\[ h(t) = \sum_{k=1}^{L} h_k(t-\tau_k), \]

where for the \( k^{th} \) path, \( h_k(t) \) is the impulse response that incorporates channel gain (but not the delay) and \( \tau_k \) is the delay, respectively, of the \( k^{th} \) path (or multipath component), and \( L \) is the total number of paths. All channel parameters vary slowly, but the time dependency is suppressed for simplicity of notation. Assume transmit pulse waveform \( p_t(t) \). While we do not directly model the antenna effects [46][78], we assume that the transmit pulse shape is designed to account for the transmit antenna frequency-dependent filter since the output of the transmit antenna must satisfy the FCC spectral mask. In practice, the receiver template should include an additional filter that accounts for the receive antenna distortion. However, this filter does not depend on the propagation mechanism that is the focus of this proposal.

The received signal can be expressed as

\[ r(t) = \sum_{k=1}^{L} p_r^{(k)}(t-\tau_k) + n(t), \]

(3.2)
where $p_r^{(k)}(t) = h_k(t) * p(t)$ represents the received pulse waveform associated with the $k$th path, $n(t)$ is zero-mean, Additive White Gaussian Noise (AWGN) random process with double-sided power spectrum density $N_0/2$, and ‘∗’ denotes convolution.

One of the benefits of the extraordinarily wide bandwidth of the UWB channel is better ability to combat multipath fading relative to narrowband or traditional Spread Spectrum (SS) systems [6][70]. In typical I-UWB channels, fading mitigation is achieved due to the fine multi-path resolution on the order of nanosecond by employing the RAKE receiver [4][31][44][45][46]. Each of these multipath components corresponds to a path (or a superposition of several paths) affected by certain propagation mechanism, and can correspond to one of the paths illustrated in Fig. 2.1. The transmitted signal is detected by collecting the energy associated with dominant (strongest) multipath components using the RAKE receiver [7][31]. Channel response of each of these paths could affect both the receiver correlation template (i.e., the impulse response of the receive filter) and the received signal-to-noise ratio [45][46][48]. For simplicity, we focus on single-path case, i.e. $L=1$, in this preliminary investigation [90]. Then the received signal reduces to

$$r(t) = p_r(t - \tau) + n(t), \quad (3.3)$$

where $p_r(t)$ is the received pulse and $\tau$ is the delay. The correlation receiver is employed at the detector [7][16]:

$$\int_{-\infty}^{\infty} r(t) v(t - \hat{\tau}) dt, \quad (3.4)$$
where \( v(t) \) is the normalized (unit energy) correlation template delayed by \( \hat{\tau} \). Denote the energy of the received pulse \( E_r = \int_{-\infty}^{\infty} p_r^2(t) dt \). We employ the ideal assumption that the peak of the cross-correlation between the received pulse and the correlation template

\[
\rho = \frac{\int_{-\infty}^{\infty} p_r(t-\tau) v(t-\hat{\tau}) dt}{\sqrt{E_r}},
\]

is achieved at \( \hat{\tau} \). Of course, in practice, finite integration window is employed when calculating the cross-correlation. The output signal-to-noise ratio is defined as:

\[
\text{SNR}_{\text{out}} = \frac{\rho^2 E_r}{N_0}.
\]

The parameter \( \rho \) in (3.6) plays a key role in the performance of the correlation detector (3.4). When per-path distortion is absent, the optimum template is the normalized transmit pulse \( p_t(t) \), while the optimum choice of the template for any received pulse is \( v(t) = \frac{p_r(t)}{\sqrt{E_r}} \), resulting in the maximized value of \( \rho = 1 \). However, if the template is not matched to the channel response, then \( \rho < 1 \), resulting in reduced energy capture [46]. For example, to keep the dB loss (SNR\textsubscript{out} reduction in (3.6)) caused by pulse distortion within 1dB, the value of \( \rho \) must be above 0.89. From (3.6), this loss is defined in terms of the SNR capture:

\[
\text{SNR}_c = \rho^2 \text{ (dB)}.
\]
3.3 UWB Pulse Waveforms

3.3.1 Gaussian Monocycles

Gaussian monocycles, particularly 2\textsuperscript{nd} order Gaussian monocycle, are the most generally adopted UWB pulses [7][16] in I-UWB literature. Part of the reason that Gaussian monocycles are so popular is the ease of mathematical modeling, i.e. the n-order Gaussian monocycle is simply the n\textsuperscript{th} derivative of the basic Gaussian monocycle. The n-order Gaussian monocycle is defined as the n\textsuperscript{th} derivative of the basic Gaussian monocycle:

\[ w_n(t) = \frac{d^n}{dt^n} \left(e^{-2\pi(t/t_p)^2}\right), \]  

(3.8)

where \( t_p \) is a parameter which controls the pulse width. Besides the ease of mathematical modeling, these pulses provide two degrees of freedom that aid in coping with a given spectral mask, i.e. the pulse order, \( n \), and \( t_p \). Note that \( t_p \) determines the bandwidth of the pulse, and \( n \) corresponds to the shift in the central frequency of the spectrum [16]. The pulse waveforms and the PSD of Gaussian monocycles of several orders are given in Fig. 3.1 and Fig. 3.2. As shown in Fig. 3.2, the shape of the PSD for Gaussian monocycles of different orders is almost invariable and the corresponding 10dB bandwidth also maintains the same for the given value of \( t_p \), i.e. 0.25ns.
Figure 3.1 Gaussian monocycles (with normalized amplitude) of different orders, $t_p=0.5$ ns.
Figure 3.2 Normalized PSD of the Gaussian monocycles shown in Fig. 3.1, $t_p=0.5\text{ns}$.
3.3.2 MHP/MMHP

MHP have also been proposed for UWB channels [17][26]. The main advantage of MHP is the orthogonality between pulses of different orders that can be utilized in M-ary orthogonal modulation or to accommodate multiple users. The n-order MHP is defined by

\[ \text{MHP}_n(t) = (-1)^n e^{0.25(t/tp)^2} \frac{d^n}{dt^n} (e^{0.5(t/tp)^2}). \]  (3.9)

The pulse waveforms of MHP of different orders are given in Fig. 3.3 and the corresponding PSD is shown in Fig. 3.4. As indicated in Fig. 3.4, the PSD of MHP (for \( n > 1 \)) contains multiple lobes with the band of spectral occupancy from very low frequency to a few GHz [18]. Therefore, it is not straightforward for MHP to cope with the FCC spectrum mask [1]. To remedy this problem, Modulated Modified Hermite Pulse (MMHP) [18] can be employed. For generating MMHP, the corresponding MHP is multiplied by a cosine wave of frequency \( f_c \):

\[ \text{MHP}_n(t) = (-1)^n e^{0.25(t/tp)^2} \frac{d^n}{dt^n} (e^{0.5(t/tp)^2}). \cos(2\pi f_c t). \]  (3.10)

The value of \( f_c \) is usually chosen to be 6.85GHz, which is the central frequency of the 3.1~10.6GHz band of the FCC mask [18]. As an example, the pulse waveform and PSD of 1st order MMHP are given in Fig. 3.5 and Fig. 3.6. As shown in Fig. 3.6, the PSD of the MMHP centers around the carrier frequency, i.e. 6.85GHz. The pulse waveform shown in Fig. 3.5 shows that the MMHP contains multiple zero-crossing points, which implies systems employing this pulse could be relatively more sensitive to the ‘timing jitter’ and channel estimation error than the original MHP with the same value of \( t_p \).
Figure 3.3 MHP (with normalized amplitude) of different orders, $t_p=0.1$ns.
Figure 3.4 Normalized PSD of the MHPs shown in Fig. 3.3, $t_p=0.1\text{ns}$. 
Figure 3.5 2\textsuperscript{nd} order MMHP (with normalized amplitude), $t_p=0.1\text{ns}, f_c=6.85\text{GHz}$. 
Figure 3.6 Normalized PSD of the MMHP shown in Fig. 3.5, $t_p=0.1\text{ns}, f_c=6.85\text{GHz}$. 
3.3.3 Prolate Spheroidal Wave Function based pulse (PSWF)

PSWF was investigated in [19][24]. It is straightforward to cope with a given frequency spectrum mask using PSWF. The basic idea is for a given spectral mask $H(f)$, we can obtain its impulse response, $h(t)$, by IFT. Then we construct a matrix $H$ whose elements are samplings of $h(t)$ (see Fig. 3.7). All the eigenvectors of matrix $H$ are candidate waveforms, because the discrete convolution of $H$ and its eigenvector equals to multiplying the eigenvector by a constant, i.e. the corresponding eigenvalue. Each eigenvalue is related to the percentage of its corresponding eigenvector’s power concentrated inside the desired frequency mask. The greater the eigenvalue, the better the power spectrum fits. Therefore, only the eigenvectors corresponding to large eigenvalues should be selected for implementation [19]. $H$ can be constructed as a Hermitian matrix [19], such that the eigenvectors corresponding to different eigenvalues are orthogonal to each other [91]. This orthogonality can be used to achieve $M$-ary orthogonal modulation or to accommodate multiple users. The associated shortcoming with PSWF is they have to be generated numerically and there is no closed-form expression. Note that when calculating the PSWF for a given spectral mask, one should also specify the targeted pulse duration, $T_m$.

Fig. 3.8 shows a PSWF pulse obtained using an ideal band-pass spectral mask between 3.1GHz and 10.6GHz with $N = 128$, $T_m=$0.5ns. The PSD of this pulse is shown in Fig. 3.9 together with the targeted spectral mask. Fig. 3.10 shows a PSWF pulse obtained using an ideal band-pass frequency mask between 0GHz and 0.96GHz with $N = 128$, $T_m=$5ns. The PSD of this pulse is shown in Fig. 3.11 together with the corresponding spectral mask. As we can see that the power spectrum of this pulse is well confined within the given
frequency mask. The reason we take 0~0.96GHz and 3.1~10.6GHz band-pass spectral masks as examples is these two bands are actually the two bands over which the highest radiation level (-41.3dBm/MHz) are allowed by the FCC spectral masks [1].

\[
\begin{pmatrix}
\psi \left[ \frac{-N}{2} \right] \\
\psi \left[ \frac{-N}{2} + 1 \right] \\
\vdots \\
\psi [0] \\
\vdots \\
\psi \left[ \frac{N}{2} \right]
\end{pmatrix}
\begin{pmatrix}
h[0] & h[-1] & \cdots & h[-N] \\
h[1] & h[0] & \cdots & h[-N+1] \\
\vdots & \vdots & \ddots & \vdots \\
h \left[ \frac{N}{2} \right] & h \left[ \frac{N}{2} - 1 \right] & \cdots & h \left[ \frac{-N}{2} \right] \\
\vdots & \vdots & \ddots & \vdots \\
h[N] & h[N-1] & \cdots & h[0]
\end{pmatrix}
\begin{pmatrix}
\psi \left[ \frac{-N}{2} \right] \\
\psi \left[ \frac{-N}{2} + 1 \right] \\
\vdots \\
\psi [0] \\
\vdots \\
\psi \left[ \frac{N}{2} \right]
\end{pmatrix}
\]

**Figure 3.7 Illustration on how to generate PSWF for a given spectral mask.**
Figure 3.8 Normalized pulse waveform of the PSWF generated for the ideal rectangular spectral mask between 3.1GHz and 10.6GHz, N=128, T_m=0.5ns.
Figure 3.9 PSD of the pulse shown in Fig. 3.8 and the targeted spectral mask.
Figure 3.10 Normalized pulse waveform of the PSWF generated for the ideal rectangular spectral mask between 0GHz and 0.96GHz, N=128, $T_m=5\text{ns}$.
Figure 3.11 PSD of the pulse shown in Fig. 3.10 and the targeted spectral mask.
3.4 Results and Discussions

We employ the paths 1 and 3 in Fig. 1 with frequency responses shown in Fig. 2.2 and Fig. 2.4 to investigate the effect of diffraction and reflection by small reflector on those families of pulses introduced in section 3.2. (The same qualitative conclusions were obtained for other paths with these propagation mechanisms). For example, Fig. 3.12 and Fig. 3.13 illustrate the received pulse waveforms (with normalized amplitude) when the Gaussian monocycle (3.1) and MHP (3.2), both of order 2, are transmitted (note that the distortion results do not depend on the pulse width $t_p$). We denote the received pulse waveform associated with the diffraction path and the reflection path as $p_{\text{diff}}(t)$ and $p_{\text{refl}}(t)$, respectively.

We have found that while all pulses are distorted by these propagation mechanisms, the distortion is particularly dramatic for the diffracted MHP pulses of all orders. Note that the PSD of the MHP pulses is significant for very low frequencies (Fig. 3.4) where the response of the diffracted path varies the most (see Fig. 2.2), while the PSD of the Gaussian monocycle occupies relatively “flat” region of the diffracted channel frequency response (recall that its amplitude resembles $H(f)=1/\sqrt{f}$). In addition, observe in Fig. 3.13 that the diffracted 2nd order MHP that has much wider tail than the transmit pulse. The latter effect is due to the similarity of the diffracted channel to the integrator. The integral of the 2nd order MHP is not zero (its PSD is significant at dc), and thus diffraction causes the dispersion of the tail of the received pulse. In fact, all even-order MHP pulses and the PSWF shown in Fig. 3.10 contain dc component, and experience similar dispersion. This pulse widening can cause interference in the receiver and reduce the data rate. Moreover, these pulses cannot be used in practice directly due to poor antenna radiating efficiency for very low frequencies [92]. How-
Figure 3.12 Normalized diffracted and reflected (reflector size 0.3m) pulses. The transmitted pulse, \( p_t(t) \), is 2\textsuperscript{nd} order Gaussian monocycle, \( t_p = 0.25 \text{ ns} \).
Figure 3.13 Normalized diffracted and reflected (reflector size 0.3m) pulses. The transmitted pulse, $P_t(t)$, is $2^{nd}$ order MHP, $t_p=0.1$ ns.
ever, their properties are of interest since they give rise to modulated waveforms with potentially low complexity receiver implementation that involves filtering of the original pulse situated at dc [93].

For comparison purpose, the received pulse waveforms associated with the diffraction path and the reflection path (reflector size is 0.3m) when 1\textsuperscript{st} order MHP is transmitted are given in Fig. 3.14. No ‘pulse widening’ is observed in Fig. 3.14, and the diffracted 1\textsuperscript{st} MHP has comparable pulse width as the transmitted pulse, \( p(t) \). That is because although the PSD of odd-order MHP is significant for low frequencies, it diminishes as the frequency goes to zero (see Fig. 3.4). That implies that the corresponding pulses do not contain dc component and have zero-integral.

We also note that the received pulses of the reflected signals for all investigated pulses are also visibly distorted, but retain general shape characteristics of the transmitted pulses. In addition to the distortion, we observe the shift of the peak of the pulse that is related to the approximately linear phase of the frequency response. The pulse distortion associated with the received UWB pulses will be measured quantitatively in chapter 4 when we calculating the cross-correlation between the received pulse with the transmit pulse and other correlation template.
Figure 3.14 Normalized diffracted and reflected (reflector size 0.3m) pulses. The transmitted pulse, $P_t(t)$, is 1st order MHP, $t_p=0.1$ ns.
Chapter 4

PULSE AND TEMPLATE DESIGN

4.1 Introduction

The pulse distortion associated with the received signal will pose difficulty for choosing an appropriate correlation template for the correlator [45][46][48] and the ‘mismatch’ between the received pulse and the correlation template could lead to low output signal-to-noise ratio (SNR) from the correlators. In chapter 3, we discussed the characteristics of several families of pulses that are widely adopted in I-UWB literature and investigated the effect of distortion on these families of pulses. In this chapter, those observations obtained in chapter 3 can be utilized when designing templates for the correlation receiver. In addition, UWB pulses that accommodate robust template choice for all propagation conditions and satisfy the FCC spectral mask for outdoor channels are also identified.

We will propose some suboptimal templates based on the observations in chapter 3. The value of SNR_c is an indication of how well the suboptimal templates perform. Finally, some UWB pulses that accommodate robust template choice for all propagation conditions and satisfy the FCC spectral mask for outdoor channels are identified. These results greatly simplify the receiver design for channels with distortion, since accurate per-path channel response estimation is not required.
4.2 Pulses that Fit the FCC Masks

In any UWB communication systems, the PSD of the transmitted signal must satisfy the FCC spectral masks for UWB communications [1]. The PSD of I-UWB signaling is largely determined by the PSD of the employed transmit pulse, \( p(t) \). Thus, it is important to identify the pulses that fit in the FCC spectral masks. We have introduced several families of pulses that are widely adopted in I-UWB literature. Within these pulse families, there are multiple choices of UWB pulses that satisfy the FCC spectral masks.

As examples, the PSD and pulse waveforms of some pulses that satisfy both the indoor and outdoor FCC spectral masks are given. 0~0.96GHz and 3.1~10.6GHz are two bands of FCC spectral masks over which the highest radiation level (-41.3dBm/MHz) are allowed (see Fig. 2.1). In Fig. 4.1 and 4.2, the PSD and pulse waveforms of some pulses that satisfy the 0~0.96GHz band of the FCC masks are given. In Fig. 4.3 and 4.4, the PSD and pulse waveforms of some pulses that satisfy the 3.1~10.6GHz band of the FCC masks are given. Note that the outdoor and indoor mask are identical over the 0~0.96GHz and 3.1~10.6GHz bands. The primary difference between the outdoor and indoor mask is the higher degree of attenuation required for the out-of-band region for outdoor operation, as shown in Fig. 2.1. This difference is for providing extra protection for some outdoor services, e.g. (Global positioning system) GPS.

The comparison in the time domain reveals that the pulses that fit the 0~0.96GHz band (Fig. 4.2) are much wider than those that fit the 3.1~10.6GHz band (Fig. 4.4) (note that the spectral components of these pulses outside of the corresponding frequency bands are negligible), and all pulses that fit the same frequency band have comparable pulse widths.
Moreover, the pulse waveforms of the three pulses in the higher band are very similar, although they belong to different pulse families, and their corresponding PSD are also very close to each other (Fig. 4.3). Fig. 4.1 shows that the PSWF from 0~0.96GHz band are more significant for very low frequencies than the other two ‘low band’ pulses, i.e. 2\textsuperscript{nd} order Gaussian monocycle with $t_p=2.5\text{ns}$ and 1\textsuperscript{st} order MHP with $t_p=0.4\text{ns}$. 
Figure 4.1 PSD of pulses that fit the 0-0.96GHz band of FCC spectral masks.
Figure 4.2 Pulse waveforms of the pulses shown in Fig. 4.1.
Figure 4.3 PSD of pulses that fit the 3.1–10.6GHz band of FCC spectral masks.
Figure 4.4 Pulse waveforms of the pulses shown in Fig. 4.3.
4.3 Suboptimal Correlation Templates

As indicated in section 4.2, the optimum choice of the correlation template for any received pulse is the (normalized) received pulse itself, which maximizes the output signal-to-noise ratio, SNR\textsubscript{out} (3.6). Since it is computationally expensive to estimate the received pulse shape based on the received signal [38][46][55][56], it is desirable to design a receiver which does not require accurate per-path channel response estimation. In chapter 2, the plots of amplitudes of channel frequency responses indicate the similarity between diffraction (reflection by small objects) mechanism and the integrator (differentiator) and the plots of pulse distortion in chapter 3 also verify the conclusion. Based on these, we propose some suboptimal correlation templates, which could greatly simplify the receiver design for channels with distortion. The three proposed suboptimal correlation templates are the transmit pulse, \( p_t(t) \), the integral \( \int p_t(t) \text{d}t \) and the derivative \( \frac{dp_t(t)}{dt} \) of the transmit pulse.

These templates are easy to generate, e.g. if \( p_t(t) \) is \( n^{th} \) order Gaussian monocycle, then \( \int p_t(t) \text{d}t \) and \( \frac{dp_t(t)}{dt} \) are simply \( (n-1)^{th} \) and \( (n+1)^{th} \) order Gaussian monocycle, respectively.

Consider 2\(^{nd}\) order Gaussian monocycle with \( t_p=2.5\text{ns} \) in 0~0.96GHz band as example. We plot the cross correlation, \( C(\alpha) = \int_{-\infty}^{\infty} p_r(t) v(t-\alpha) \text{d}t / \sqrt{E_r} \), between the diffracted pulse (path 1 in Fig. 2.1) and the proposed suboptimal templates in Fig. 4.5. It is observed in Fig. 4.5 that both \( p_t(t) \) and \( \int p_t(t) \text{d}t \) produce very high peak cross correlation (>0.9). On the other hand, the cross correlation between the reflected pulse (path 3 in Fig. 2.1, reflector
size=0.3m) and the proposed suboptimal templates (Fig. 4.6) indicates that in reflection case, \( p(t) \) and \( \frac{dp(t)}{dt} \) are better choices than \( \int p(t) dt \). In section 4.4, we will evaluate the performance of the three suboptimal templates in terms of SNR capture (3.7).

Figure 4.5 The cross correlation between diffracted 2nd order Gaussian monocycle, \( t_p=2.5\text{ns} \), and three suboptimal correlation templates.
Figure 4.6 The cross correlation between the reflected 2\textsuperscript{nd} order Gaussian monocycle, $t_p=2.5\text{ns}$, and three suboptimal correlation templates.
4.4 Results and Discussions

In Fig. 4.7, Fig. 4.8 and Fig. 4.9, we plot the SNR capture (3.7) that corresponds to the peak correlation (4.4, 4.5), between the distorted received pulses and the proposed suboptimal correlation templates for the diffracted path 1 (with response $p_{\text{diff}}(t)$) and reflected path 3 ($p_{\text{refl}}(t)$) in Fig. 2.1. In all cases, the transmit pulse is utilized as one of the templates. The integral $\int p(t)dt$ and the derivative $\frac{dp(t)}{dt}$ of the transmit pulse were employed as templates for the diffracted and the reflected cases, respectively.

For the Gaussian monocycles (Fig. 4.7), all template choices for $n>1$ result in less than 1 dB loss due to template mismatch. The transmit pulse shape, $p(t)$, results in the highest peak correlation for the diffracted channel, and is only slightly inferior to the derivative for the reflected channel. Note that the pulse shift that produces peak correlation for these mismatched templates differs from the actual pulse delay $\tau$. The shift $\tau - \hat{\tau}$ is within 20% of the pulse width for all pulses, propagation mechanisms and suitable simple templates considered in this thesis.

In Fig. 4.8, the peak cross-correlations for the MHP signals are shown. While the results are similar to those in Fig. 4.7 for the reflected path, our suboptimal correlation templates (the transmit pulse, $p(t)$ and the integrator, $\int p(t)dt$) do not yield satisfying results for the diffracted MHP pulses, especially for the MHP of even orders due to the ‘pulse widening’ problem. We have discussed above that MHP pulses undergo more significant distortion due to diffraction, and have observed that the integration-like diffraction mechanism causes the dispersion of the tail of the diffracted even-order MHP, which is
referred to as ‘pulse widening’. The latter effect results in especially low peak correlation at
the receiver since the time-limited correlation template is only able to “catch” a portion of the
received energy. Like even-order MHP, the PSWF pulse in the 0–0.96GHz band (Fig 3.8 and
Fig. 4.2) also contains dc component, and the corresponding SNR$_c$ when the transmit pulse is
used as template is on the order of -1.5 dB when the integration window is large.

Fig. 4.9 illustrates the comparison of the SNR$_c$ between the MHP and the MMHP
when the transmit pulse shape is used as template. The high-frequency carrier of the MMHP
introduces odd symmetry and spectral shaping that is beneficial for the diffracted channel as
discussed above (see Fig. 3.5 and Fig. 3.6). Therefore, MMHP results in much higher peak
cross-correlations than MHP. Similar large correlation was observed for the PSWF pulse in
the higher band (see Fig. 3.8 and Fig. 4.4). The corresponding SNR$_c$ when the transmit pulse
is used as template is always less than 1dB regardless of propagation mechanisms.
Figure 4.7 SNR capture, Gaussian monocycles, $t_p=0.25$ns.
Figure 4.8 SNR capture, MHP, $\tau_p=0.1\text{ns}$. 
Figure 4.9 SNR capture, MHP ($t_p=0.1\text{ns}$) and MMHP ($t_p=0.1\text{ns}, f_c=6.85\text{GHz}$).
From these results, we conclude that when MMHP, PSWF in the 3.1-10.6GHz band and Gaussian monocycles are used as transmit pulses, the transmit pulse shape represents near-optimal template for both diffracted and reflective single-path channels, resulting in the SNR capture (3.7) within 1dB of the ideal template. This conclusion greatly simplifies the receiver design for channels with distortion, since accurate channel estimation and adaptive receiver design are not required. Thus, these transmit pulses are ‘robust’ for outdoor UWB channels where LOS, diffraction, and reflection are dominant propagation mechanisms. This conclusion is consistent with results in [46], where the transmit pulse with spectrum in the higher band (similar to those in Fig. 4.3) was investigated for indoor channels with distortion due to penetration through various materials. For the single path case, using the transmit pulse as the template produces near-optimal peak correlation, while more accurate channel estimation provided lower path loss for the multipath case. However, this higher order estimation also causes higher susceptibility to noise and narrowband interference [46]. From our results and [46] we conclude that when the transmit pulse shape is chosen appropriately, the template given by the transmit pulse, $p_t(t)$, is a simple and desirable choice for diverse propagation conditions. On the other hand, our results indicate that for certain pulses (e.g. MHP or PSWF in Fig. 4.1), more accurate template design is required to obtain high energy capture.
5.1 Introduction

It is well known that shadowing, obstructions, etc. can significantly attenuate received signal power. This propagation loss is referred to as large-scale fading and affects all frequency components of conventional narrowband and spread spectrum systems similarly. However, in the UWB channel, the received power is frequency-dependent due to the exceptionally wide bandwidth of impulse radio signals. The modeled channel frequency responses in Chapter 2 indicate that both diffraction and reflection (by small objects) cause the strength of the received power to be frequency-dependent. For example, high frequency components are strongly attenuated in diffracted channels (Fig. 2.2). The frequency dependency of the path loss associated with diffraction was included in the recent UWB model described in [68].

Most of the research on impulse radio to date has focused on the utilization of the 3.1–10.6 GHz band of the FCC mask. However, this spectrum allocation can potentially result in substantial performance loss due to the frequency dependency associated with the received power. We have discussed previously that diffraction can be dominant propagation mechanism for some outdoor environments. The loss due to diffraction is much lower in the 0–0.96GHz band than in the 3.1–10.6GHz band. In this chapter, we will study the propagation gain (PG) associated with pulses that reside in the two different bands, i.e. 0–0.96GHz band and 3.1–10.6GHz band for different channel conditions, and show that
there exists a large gap in the PG for pulses in the two different bands of the FCC masks. Motivated by this large gap, we propose an adaptive transmission scheme, i.e. to utilize different bands of the spectrum (e.g. pulse from 0~0.96GHz band or 3.1~10.6GHz band) as a function of the dominant propagation mechanism.

5.2 Frequency-dependent Propagation Gain

In order to show the dependency of the received signal power on the frequency, define the PG as the ratio of the received and the transmitted signal energies:

\[
\text{PG} (p_t) = \frac{\int_{-\infty}^{\infty} p_r^2(t) d(t)}{\int_{-\infty}^{\infty} p_t^2(t) d(t)}. \quad (5.1)
\]

Note that this definition does not include the SNR capture (3.7) (received energy loss) due to the template mismatch, while the free space path loss due to the distance between the transmitter and the receiver is included, and the energy of the transmit pulse is measured at 1m (LOS) from the Tx antenna [60][64][70].

In Fig. 5.1, the PG (5.1) is compared for the UWB pulses shown in Fig. 4.1-4.4 that reside in two different bands of the FCC mask as the receiver moves along the horizontal line A (see Fig. 2.1) from the shadowed area to the LOS area, and the reflector is not present. While all pulses become weaker as the receiver advances deeper into the shadowed region, the loss for the pulses in the 3.1~10.6GHz band is at least 13dB greater than for the pulses in the lower frequency band. While the PSWF is about 5dB stronger in diffraction than the other two ‘low frequency’ pulses since its PSD occupies very low frequencies (see Fig. 4.1)
where the diffracted channel response is the least attenuated, this pulse is not feasible in practice due to its dc component.

On the other hand, reflection by a small reflector attenuates low frequencies more severely. In Fig. 5.2, the channel strength loss due to reflection by small reflector is illustrated for the Gaussian monocycles that occupy two different bands of the spectral mask (Fig. 4.1 and Fig. 4.3). We observe that the propagation gain of the reflected signal decreases linearly with the logarithm of the size of the reflector for sufficiently small reflector sizes. Above certain threshold (that decreases with the wavelength of the transmitted signal), the received signal does not experience loss due to reflection. For small reflectors, the pulse in the lower band experiences about 13 dB greater loss than the pulse in the higher band.

The insight into the 13dB gap between pulses in the two bands for both diffraction and reflection can be obtained by the examining the channel responses at the ‘mode frequencies’ $f_{m,l}$ and $f_{m,u}$ (corresponding to the peak of the pulse PSD) for the lower and upper bands, respectively. For the 2nd order Gaussian monocycle with $t_p=2.5$ns, $f_{m,l} \approx 0.34$GHz, while the 8th order Gaussian monocycle with $t_p=0.25$ns has $f_{m,u} \approx 6.5$GHz. Using the approximations $1/\sqrt{f}$ and $\sqrt{f}$ of the diffracted and reflected channel amplitude responses, respectively, the ratio of the gains at the lower and upper mode frequencies is

$$\left( \sqrt{f_{m,u}} / \sqrt{f_{m,l}} \right)^2 \approx 12.8\text{dB} \quad (5.2)$$

gap for the diffracted path, and the inverse of this expression, or

$$\left( \sqrt{f_{m,l}} / \sqrt{f_{m,u}} \right)^2 \approx -12.8\text{dB} \quad (5.3)$$

gap for the reflection by small reflector.
Figure 5.1 Propagation gain of pulses in Fig. 4.1-4.4 as the receiver moves along line A (see Fig. 2.1). The y-coordinate of the receiver position is 20 (m), fixed, reflector is not present.
Figure 5.2 Propagation gain of the reflected signal (path 3 in Fig. 2.1), coordinate (10, 20) for different reflector sizes.
Using Fig. 5.1 and 5.2, we can compare the strengths of the reflected and the diffracted signals at test point (10, 20) in Fig. 2.1. For example, for the 8\textsuperscript{th} order Gaussian monocycle, the propagation gain of the diffraction path at this location is -62.96 dB. Thus, when reflector size is below 0.013 m, the reflected path at this location is weaker than the diffraction path. However, for the 2\textsuperscript{nd} order Gaussian monocycle, the corresponding PG in is -49.81 dB (i.e. ~20dB additional loss relative to the free space path loss at this location), and diffracted path is stronger than the reflected path for reflector sizes below 0.25 m. While these quantitative results are applicable only for given geometry, the same qualitative conclusions hold for other scenarios. These examples indicate that in the absence of LOS and large reflectors, either diffraction or reflection by a small reflector can be the dominant propagation mechanisms. For example, propagation into a canyon-like street can be dominated by diffraction. When a small reflector such as a sign or lamp-post is within LOS of both the transmitter and the canyon-like street, it can reflect energy into the street, and will likely be the dominant signal provider for part of that street. A similar situation is found for a tunnel. In the absence of reflectors and when the transmitter is not in LOS, the tunnel will receive the signal via diffraction. However, even a small reflector near the entrance to the tunnel can provide stronger signal, with a very different UWB channel character.

In Fig. 5.3, PG of the reflection path 3 (Fig. 2.1) is illustrated for the Gaussian monocycles that occupy two different bands of the spectral mask (Fig. 4.1 and Fig. 4.3) as the receiver moves along the horizontal line A (Fig. 2.1). The size of the reflector is 0.3m. PG of the corresponding diffraction path 1 (Fig. 2.1) is also plotted for comparison purpose. Fig. 5.3 shows when the receiver is at (10, 20), i.e. the point corresponding to specular
reflection in the geometry given by Fig. 2.1, the reflected 2\textsuperscript{nd} order Gaussian monocycle by the reflector of size 0.3m has comparable strength as the diffracted 2\textsuperscript{nd} order Gaussian monocycle. For all receiver positions to the left of (10, 20), the reflected signal is stronger and for all positions to the right of (12, 20), the diffracted signal is stronger. While for the investigated 8-order Gaussian monocycle, the reflected signal is 28dB stronger than the diffracted signal when the receiver is at (10, 20). For all position to the right of (16, 20), the reflected 8\textsuperscript{th} order Gaussian monocycle is considerably stronger than the diffracted 8\textsuperscript{th} order Gaussian monocycle. Comparing the strength of the reflected 2-order Gaussian monocycle and the reflected 8-order Gaussian monocycle, we can see that except for the point (10, 20) and its vicinity, the reflected 2-order Gaussian is always 12~13dB stronger than the reflected 8-order Gaussian. This is due to the fact that ‘direct-reflection’ is not present at those positions and the dominant propagation mechanism is diffraction. On the other hand, when the receiver is at (10, 20), the frequency response of the corresponding reflected path can be approximated as $\sqrt{f}$, which explains why the reflected 8-order Gaussian monocycle from high frequency band is stronger.

In Fig. 5.4, PG of the reflection path 3 (Fig. 2.1) is for a larger reflector size: 3m. Again, the receiver moves along the horizontal line A (Fig. 2.1) and the two Gaussian monocycles that occupy two different bands of the spectral mask (Fig. 4.1 and Fig. 4.3) are employed. It is observed in Fig. 5.4 that for the reflector of size 3m, the reflected 2-order Gaussian monocycle is as strong as the reflected 8-order Gaussian monocycle. At position (10, 20) and it is vicinity, the reflected 2-order Gaussian monocycle is more than 20 dB stronger than the diffracted 2-order Gaussian monocycle, while the reflected 8-order
Gaussian monocycle is 35 dB stronger than the diffracted 8-order Gaussian monocycle. The observation indicates that a large reflector located in LOS area would help greatly when the receiver is in the shadowed area and a specular reflection can be seen at the receiver position.

From Fig. 5.3 and Fig. 5.4 we can see that when the receiver is shadowed from the transmitter and no specular reflection is available, transmitting low-band pulse, e.g. 2-order Gaussian monocycle with $t_p=2.5\text{ns}$, would likely result in stronger received signal. Combined with the observations in Fig. 5.1 and Fig. 5.2, we can conclude that dominant propagation mechanism plays an important role in determining the strength of the received signal in UWB communications and it is important to take into account in signal design.
Figure 5.3 Propagation gain of the reflection path 3 (Fig. 2.1) as the receiver moves along Line A, reflector size is 0.3m.
Figure 5.4 Propagation gain of the reflection path 3 (Fig. 2.1) as the receiver moves along Line A, reflector size is 3m.
5.3 Potential Benefits of Adaptive Transmission

In this section, we discuss the potential for adaptive transmission that utilizes different bands of the UWB spectrum as a function of the dominant propagation mechanism. These signaling methods are motivated by the large gap in the PG for pulses in two different bands of the FCC mask. Of course, it is well known that lower frequencies propagate better in the presence of obstacles [70]. However, our novel physics-based study provides quantitative comparison of distortion and channel gains required for the development and verification of the proposed adaptive transmission schemes. In particular, assuming an ideal single-path received signal affected by either LOS of diffraction, a simple adaptive signaling method is to utilize the 8th order Gaussian monocycle in LOS, but to switch to the 2nd order Gaussian monocycle when the receiver is in the shadow. The receiver has to send the feedback signal to the transmitter to switch pulses when the appropriate power threshold is detected. This approach can be viewed as a variable rate adaptive transmission method, since utilization of the wider pulse in the lower band would result in decreased transmission rate for practical modulation techniques relative to the case when the narrower pulse in the higher band is employed [4][7]. In addition to improved bandwidth and power efficiency relative to the conventional techniques that utilize fixed pulses in the higher band, the proposed adaptive approach relaxes the stringent timing requirements of UWB impulse radio systems and improves timing estimation due to longer duration of the lower band pulse [31][94][95]. This is especially beneficial for degraded channel conditions when diffraction is dominant. The key feature of the proposed adaptive transmission method is the ability to switch between
various bands depending on propagation conditions. This adaptive strategy can be applied to all UWB systems to improve the transmission bandwidth and power efficiency.

5.4 Summary

In this chapter, we quantified the frequency-dependent propagation gain associated with pulses from the two disjoint bands of the FCC spectral mask for variant channel conditions. Our results reveal that there exists a large gap in the propagation gain for pulses from the two different bands, e.g. diffracted ‘low band’ pulses are at least 13dB stronger than diffracted ‘high band’ pulses (Fig. 5.1). This large gap implies potential benefits of adapting the transmitted signal to the dominant propagation mechanism. As an example, a simple adaptive signaling method is proposed, i.e. to utilize the 8th order Gaussian monocycle in LOS, but to switch to the 2nd order Gaussian monocycle when the receiver is in the shadow. This adaptive transmission method can not only improve the system energy efficiency, but also relax the stringent timing requirements of UWB impulse radio systems and improve timing estimation due to longer duration of the lower band pulse. On the other hand, utilization of the wider pulse in the lower band would result in decreased transmission rate.
Chapter 6

CONCLUSIONS AND FUTURE WORK

6.1 Conclusions

This thesis concentrated on improving the UWB systems by taking into account their physical channel characteristics, e.g. the frequency dependency associated with UWB channels. We mainly focused on investigation of per-path distortion, correlation template design and the frequency-dependent propagation loss for various propagation mechanisms.

The physical UWB channel model we developed is based on a Fresnel diffraction augmentation of the method of images [76-78], and is able to give an accurate description of how the strength and shape of the received pulse changes with position in given local environment, even in the proximity of the reflectors. The modeled UWB channel frequency responses indicate that some propagation mechanisms, e.g. diffraction and reflection (by small objects) introduce frequency dependency of the received power. This implies the spectral characteristic of the transmitted signal would be changed and the received signal would be distorted. We observed that the slope of the amplitude of the channel transfer function resembles that of $H(f)=1/\sqrt{f}$ for deep shadowing, while the integrator with response $1/f$ provides satisfying intuitive explanation for the distortion associated with diffracted channels. For the reflection by small reflector, the low frequencies experience great loss, and the slope of the response is modeled well by that of $H(f)=\sqrt{f}$, while the derivative $H(f)=f$ is a simple approximation to this channel.
We employ the modeled frequency responses to investigate the effect of per-path distortion on several widely adopted families of pulses, i.e. Gaussian monocycles, MHP/MMHP and PSWF. We have found that while all pulses are distorted by these propagation mechanisms, i.e. diffraction and reflection (by small objects), the distortion is particularly dramatic for the diffracted MHP pulses of all orders and diffracted PWSF in 0~0.96GHz band. This is because these pulses are significant for very low frequencies where the frequency response of the diffracted path varies the most. Another observation is ‘pulse widening’ problem associated with pulses that contain dc component, which is due to the similarity of the diffracted channel to the integrator. This ‘pulse widening’ can cause interference in the receiver and reduce the data rate. All even-order MHP contain dc component and thus subject to this ‘pulse widening’ problem. To remedy this problem, Modulated Modified Hermite Pulse (MMHP) can be employed instead, the high-frequency carrier of the MMHP introduces odd symmetry and spectral shaping that is beneficial for the diffracted channel.

The per-path distortion poses difficulty for choosing an appropriate correlation template for the correlator and the ‘mismatch’ between the received pulse and the correlation template leads to low output SNR from the correlators. Based on the observations on characteristics of channel responses and pulse distortion, we proposed some suboptimal correlation templates, which are easy to generate and produce good results. We found when MMHP, PSWF in the 3.1~10.6GHz band and Gaussian monocycles are used as transmit pulses, the transmit pulse shape represents near-optimal template for both diffracted and reflected single-path conditions, resulting in $\text{SNR}_c$ within 1dB of the ideal (optimal) template.
This conclusion greatly simplifies the receiver design for channels with distortion, since accurate per-path channel response estimation is not required. Thus, these transmit pulses are ‘robust’ for outdoor UWB channels. On the other hand, our results indicate that for certain pulses (e.g. MHP or PSWF in 0~0.96GHz band), more accurate template design is required to obtain high energy capture.

Finally, we studied the frequency-dependent propagation gain associated with pulses from the two disjoint bands of the FCC spectral mask for variant channel conditions. Our results reveal that there exists a large gap in the propagation gain for pulses from the two different bands, e.g. diffracted ‘low band’ pulses are at least 13dB stronger than diffracted ‘high band’ pulses. This large gap implies potential benefits of adapting the transmitted signal to the dominant propagation mechanism. As an example, a simple adaptive signaling method was proposed, i.e. to utilize the 8th order Gaussian monocycle in LOS, but to switch to the 2nd order Gaussian monocycle when the receiver is in the shadow. This adaptive transmission method can not only improve the system energy efficiency, but also relax the stringent timing requirements of UWB impulse radio systems and improve timing estimation due to longer duration of the lower band pulse.
6.2 Future Work

Due to lack of channel measurements, the channel model for the lower band of FCC spectral mask is still absent in [68]. Extension of our single-path model to multipath channel model, where each path might experience per-path distortion specific to its propagation mechanism, and identification of dominant propagation conditions for this model is required to test the effectiveness of the proposed adaptive signaling techniques.

While we have previously concentrated on classical I-UWB systems that utilize carrier-less pulses, the future investigation should include recently proposed DS-UWB [8][28] and MB-OFDM [10][11] UWB systems. The comparison between these systems was carried out in [9] for the IEEE 812.15 UWB model, and can be extended to include the frequency-dependent per-path distortion. It is expected that the performance of these systems will be significantly affected by shadowing and other challenging propagation environments.

The research on adaptive transmission can be extended to multi-band and DS-UWB systems. In the current MB-OFDM proposal for these systems [9][10][11], several 500-MHz bands of the UWB spectrum from 3.1 to 10.6 GHz are utilized, and Frequency Hopping (FH) or Frequency Division Multiple Access (FDMA) techniques are employed to avoid the inter-piconet interference and to provide diversity. The proposed DS-UWB [9][13] system also uses two distinct spectral bands. In these proposed systems, adaptive FH between the bands of the FCC spectrum in the 0~0.96GHz and the 3.1~10.6 GHz regions can be employed to improve the system energy efficiency. This adaptive design should take into account dominant channel propagation environment and other channel impairments, interference, and
system constraints. Moreover, joint adaptive FH and channel loading approaches can be explored.
BIBLIOGRAPHY


