Ultra-Wideband Direct-Sequence Impulse Radio

Wireless Communications

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by

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To my dear parents and brother
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Ultra-Wideband Direct-Sequence Impulse Radio Wireless Communications

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Abstract

The potential strength of ultra-wideband (UWB) radio technology lies in its extremely wide transmission bandwidth, which results in desirable capabilities including high channel capacity, resistance against fading, and high multiple access capability. However, there are still challenges in making this technology live up to its full potential. The main challenges are on account of the very low allowed average power spectral density (PSD) and the very fine multipath delay resolution. In this thesis, we will focus on the design of UWB direct-sequence impulse radio wireless communication systems.

Since UWB wireless communication is very power-limited, we will investigate the potential of improving its power-efficiency by increasing the dimensionality of the signal waveforms. Using $N$ orthogonal pulse shapes and $M$ orthogonal code-words, we will propose a $2MN$-ary biquantum keying modulation and will show its power-efficiency improvement for direct-sequence multiple-access impulse radio systems, compared with bipolar pulse amplitude modulation and biquantum-code keying.

The overlap between pulses arriving at the receiver is generally ignored in impulse radio studies, but it occurs when the pulse duration is greater than the fine multipath delay resolution of the UWB channel, thereby reducing the power-efficiency of correlator-based UWB receivers. We will provide a performance analysis incorporating the effect of channel-induced pulse overlap for direct-sequence multiple-access impulse radio systems. The detailed analysis as well as extensive simulations will reveal that channel-induced pulse overlap has a significant impact on the system performance and
our analytical formula shows its accuracy and importance in performance analysis for direct-sequence impulse radio communications.

UWB communications with limited transmission power while signal energy dispersed by a large number of multipath components, require a practical Rake receiver that can provide a desirable output signal-to-noise ratio (SNR). We will propose a pilot-channel assisted (PCA) generalized selection combining (GSC) with log-likelihood ratio (LLR) threshold test per path (PCA-\(|LLR|\)-T-GSC) for UWB Rake receivers with limited number of fingers. We will show the good of PCA-\(|LLR|\)-T-GSC in providing desirable performance under channel time-variations and multi-user interference, while without sacrificing the system data rate.

Our future work will focus on investigating methods for increasing the achievable distance of UWB systems for a fixed average PSD, by trading lower spectral efficiency for increased power efficiency to achieve a desired rate/range operating point. We will also investigate practical algorithms for initiating and maintaining synchronization at the UWB receiver, to combat the challenges in timing synchronization induced by the stringent power constraints and the fine delay resolution of UWB signals.
Notations

Throughout this thesis, bold uppercase (lowercase) letters denote matrices (vectors).

Besides, the following notations are also used:

- \( \infty \) : infinity
- \( \in \) : belongs to
- \( \min[\cdot] \) : the minimum value of \( \cdot \)
- \( \max[\cdot] \) : the maximum value of \( \cdot \)
- \( \arg\min_x f(x) \) : the value of \( x \) (argument) that minimizes \( f(x) \)
- \( \arg\max_x f(x) \) : the value of \( x \) (argument) that maximizes \( f(x) \)
- \( |\cdot| \) : the norm of \( \cdot \)
- \( (\cdot)^* \) : (superscript only) complex conjugate
- \( (\cdot)^T \) : (superscript only) transpose
- \( (\cdot)^H \) : (superscript only) complex conjugate transpose
- \( \mathbf{I}_K \) : an identity matrix of size \( K \times K \)
- \( \mathbf{0}_{M \times N} \) : an all-zero matrix of size \( M \times N \)
- \( \delta(\cdot) \) : the Dirac delta function
- \( \mathbb{E}[\cdot] \) : expectation
- \( \lfloor \cdot \rfloor \) : integer floor
- \( \lceil \cdot \rceil \) : integer ceiling
- \( Q(\cdot) \) : the standard \( Q \)-function

\[ \sum_{i=1}^{n} x_i = x_1 + x_2 + \ldots + x_n, \text{ sum over } i (i = 1, 2, \ldots, n) \]

\[ \prod_{i=1}^{n} x_i = x_1 \cdot x_2 \cdot \ldots \cdot x_n, \text{ product over } i (i = 1, 2, \ldots, n) \]
Chapter 1

Introduction

In this chapter, we introduce the motivation and background of the thesis research, and outline the main contributions and the organization of this thesis.

1.1 Motivation

In the last few years, four trends have been driving the development of short-range wireless technologies [2]:

1. The growing demand for higher wireless data capability needed to support increasingly sophisticated broadband applications among networks of portable electronic devices.

2. Crowding in the spectrum that is segmented and licensed by regulatory authorities.

3. The growth of high-speed wired connections to the Internet in enterprises, homes, and public spaces and the fast-growing combination of wireless and wired Internet.

4. Shrinking semiconductor cost and power consumption for signal processing.

Trends 1 and 2 favor systems that offer higher data rates but with lower cost and lower power consumption than currently available, as well as a high spatial capacity (a metric measured in bits per second per square meter). As increasing numbers of broadband users gather in crowded spaces, like airports, hotels, convention centers and workplaces, the most critical parameter of a wireless system will be its spatial capacity [3]. Trend 3 makes it possible to provide communications between portable electronic devices and some kind of wired links to the Internet using short-range wireless standards like Bluetooth and IEEE 802.11. Trend 4 makes possible the use of signal processing techniques that would have been impractical only a few years ago.

However, many existing wireless technologies do not satisfy the demands of the broadband multimedia networks, nor do they provide the necessary performance to support multiple audio-video streams. The low-power Bluetooth standard, which is designed to replace the USB cables for passing data among closely located electronic
equipments, offers a maximum data rate of about 700 kbps over distances of up to 10 meters [3]. This is too little for even one digital TV channel, which needs at least 2 Mbps using heavy signal compression [3]. The IEEE 802.11b and 802.11a standards are established for wireless local area networks (WLANs), where the former provides maximum data rates of around 5.5 Mbps across distances of up to 100 meters [3] and the latter provides maximum data rates of 24 to 35 Mbps over distances of up to 50 meters [3]. Although IEEE 802.11 technologies offer higher data-rate and longer range, their fundamental flaw is that the power consumption requirements of around 1.5 to 2W make them unsuitable for battery dependent devices like PDAs and even laptops with short battery lives [2].

All these have led to the current interest in the use of ultra-wideband (UWB) radio technology for commercial applications. UWB is defined by the United States Federal Communications Commission (FCC) as any radio technology that occupies a bandwidth greater than 20% of its center frequency or a bandwidth of at least 500MHz. It is Trend 4 that makes UWB now commercially practical. The progress in microelectronics industry makes possible a hardware realization as well as high potentials of UWB radio systems. The FCC Report and Order (R&O) issued in February 2002 [4] allocated the 3.1-10.6GHz frequency band for unlicensed use of UWB devices under the Part 15 rules, which require the average power spectral density (PSD) to be less than –41.25dBm/MHz. The low PSD of UWB signals allows low average radiated power and possible coexistence with incumbent radio systems. The allowed bandwidth of UWB is much greater than the bandwidth ever used by any current technology for wireless communications, e.g., Bluetooth and IEEE 802.11b that operate in the unlicensed 2.400-2.483GHz band, and IEEE 802.11a operating in the 5.150-5.350GHz band.

UWB systems, occupying several-gigahertz bandwidth, have greater room for channel-capacity expansion than systems that are more bandwidth-constrained, because the capacity of a channel grows linearly with the available bandwidth [2]. As a result, UWB is superior in spatial capacity to other short-range wireless technologies, as shown in Figure 1.1. UWB is expected to provide very high data rates – 100 to 500Mbps across distances of 5 to 10 meters – make it favorite for audio/video distribution around the house, as well as cable replacement for USB and FireWire [3]. These high data rates will
also give rise to applications that are impossible using existing wireless standards. The recently established IEEE 802.15 Task Group 3a [5] is defining a PHY alternative to respond to demands in high-data-rate wireless personal area networks (WPANs) based on UWB radio technologies [3]. At the same time, the IEEE 802.15 Task Group 4a [6] is currently working on a standard specifying low-data-rate, low-power, and low-cost WPAN communications, based on UWB radio technologies as well.

![Spatial capacity comparison between IEEE 802.11, Bluetooth, and UWB](image)

**Figure 1.1** Spatial capacity comparison between IEEE 802.11, Bluetooth, and UWB [2].

UWB radio technology has become a leading candidate for providing short-range high-data-rate applications in WPANs, such as wireless USB and wireless IEEE1394, as well as low-data-rate, low-power-consuming and low-cost wireless applications. However, the same qualities that make UWB radio attractive also bring the design challenges. With very limited transmission power, UWB is facing the difficulties of quickly and accurately detecting signals below the noise floor and in the presence of multiple-access users and dense multipath. As the ultra-wide bandwidth offers excellent
frequency diversity and very fine delay resolution, the channel estimation and multipath combining tasks become correspondingly more challenging, because the signal energy is dispersed by a large number of multipath components and consequently the signal energy per path is significantly reduced [9]. For a correlator-based UWB receiver, the timing needs to be very accurate to properly detect the received short-duration pulses, while the fine multipath delay resolution of UWB signals will increase the sync acquisition time [10].

There are many methods for generating a UWB waveform, however, in this thesis, we will focus on the pulse-based UWB technology, a.k.a. impulse radio [2]. Impulse radio is typically implemented in a carrierless fashion, i.e. instead of using a continuous carrier, impulse radio emissions are composed of a series of intermittent extremely-short pulses. Information is coded into the pulse train by varying the pulses’ amplitudes, polarities, timing (pulse-to-pulse period) or other characteristics.

In this thesis, we focus on the design of power-efficient direct-sequence impulse radio UWB systems for use in wireless communications. Specifically, we will develop a multidimensional modulation to improve the power-efficiency of UWB impulse radio communications. We will propose a performance analysis approach for accurately evaluating direct-sequence impulse radio systems operating in multipath and multiple-access environments. Furthermore, we will investigate other important issues, including reliable channel estimation and tracking in time-varying channels, and path selection for Rake combining that can help balance receiver complexity with performance for UWB systems.

1.2 Background

Ultra-wideband (UWB) impulse radio systems employ trains of ultra-short pulses to convey information [4], [11]-[14]. The FCC currently allows UWB communications devices to operate under Part 15 rules within the band 3.1-10.6GHz with a maximum average power spectral density (PSD) of only -41.25dBm/MHz [4]. Therefore, UWB is primarily power constrained and bandwidth unconstrained, and power efficiency becomes the critical constraint behind choice of modulation approaches [16].

In previous works, binary modulations for impulse radio systems include binary
pulse position modulation (PPM) \cite{11, 14} and bipolar pulse amplitude modulation (PAM) \cite{17, 18}, among which bipolar PAM is the most power-efficient \cite{19}. $M$-ary PPM schemes \cite{20}, \cite{21} with sufficiently large $M$ can be more power-efficient than bipolar PAM, but the spectral lines or spikes in the signal PSD caused by PPM compel a reduction in the total transmission power in order to comply with the Part 15 constraints. The orthogonal pulse polarity modulated time-hopping (TH) UWB system \cite{21} uses Walsh codes to construct $M$-ary orthogonal signals, but the Walsh codes need to be elaborately modified to balance the negative and positive polarities of modulated pulses to smooth the envelop of signal PSD. Bipolar pulse waveform and position modulation \cite{22} and biorthogonal PPM \cite{23} both apply antipodal and pulse position modulations at the same time on Gaussian pulses, where the antipodal modulation offers a smooth PSD while the PPM may not be effective in Rake reception, because the delayed multipath arrivals may be confused with data modulation when the separation of delays is on the order of the PPM modulation index $\delta$ \cite{21}. $M$-ary pulse shape modulation (PSM) \cite{12, 24, 25} that adopts $M$ orthogonal pulse shapes can be less time-precision restricted and more multipath-resistant than $M$-ary PPM. However, PSM relies on orthogonal pulse-shape design.

There are several recent works on orthogonal pulse shaping for impulse radio. In \cite{24} and \cite{25}, orthogonal pulses without dc component were generated based on the combination of Hermite functions and the use of prolate spheroidal wave functions, respectively. Orthogonal modified Hermite pulses were proposed in \cite{12, 26}. But the above pulses occupy frequency bands not allowed by the current FCC spectrum regulations for UWB \cite{4}. The algorithm in \cite{27} uses eigenvectors of a Hermitian matrix to generate orthogonal pulses under the desired spectrum mask, but the pulse duration ($T_w$) increases with the number of orthogonal pulses generated – with a $T_w$ around 1ns only two orthogonal pulses can be obtained. In addition, it has been shown that higher order pulses may be more susceptible to any distortion in transmission \cite{28}.

Performance analysis for impulse radio systems has been reported by several authors. In \cite{10, 18, 22}, the analysis assumed channels without multipath fading. In \cite{13, 17, 29, 76}, the multipath delay resolution of the channel employed was set to be the duration ($T_w$) of the modulated pulses, i.e. multipath arrivals would be at integer
multiples of $T_w$ and would not overlap at the receiver. However, it is not the duration ($T_w$) but the bandwidth ($B$) of the transmitted pulses that decides the multipath resolvability of a wireless propagation channel, and the minimum path spacing ($\Delta \tau$) is approximately $1/B$ [30]. In fact, the fine multipath delay resolution provided by the ultra-wide bandwidth of impulse radio signals can affect the performance of a correlator-based receiver. That is, if $\Delta \tau < T_w$, pulses arriving at the receiver will overlap and the pulse correlator output will be corrupted by correlations of the template pulse with interfering pulses that arrive or have tails within the integration interval of the correlator. In [31], [77], modulations using a 1ns-duration pulse were examined under a 6GHz-bandwidth channel model, but the effect of pulse overlap on correlation reception was treated implicitly. Also note that almost all published UWB pulses to date [10], [12], [24], [25], [27] have effective time durations greater than the inverse of their frequency bandwidths, i.e. $T_w > 1/B \ (\approx \Delta \tau)$. Therefore, the failure of including the effect of channel-induced pulse overlap in performance analysis for impulse radio systems could be a significant issue. In addition, the particular modulation scheme used also affects the system performance, for which desirable characteristics include power-efficiency and a resultant power spectral density (PSD) without discrete spectral lines [19]. In such endeavors, bipolar pulse amplitude modulation (BPAM) [17], [18] was the most power-efficient binary modulation [19]. Among the published $M$-ary modulation schemes [12], [20]-[25], pulse shape modulations (PSM) [12], [22], [24], [25] using multiple orthogonal pulses show advantages in being less time-precision restricted and being more multipath-resistant [24].

With information conveyed by ultra-narrow pulses, UWB radio can resolve a large number of multipath components using a Rake receiver [32] to achieve diversity [2]. Although combining all resolvable multipath components with maximal-ratio combining (MRC) [32] is optimum in the perspective of performance [33], the number of multipath components that can be utilized by a Rake receiver is limited by hardware implementation, power consumption, and channel estimation [34]. On the other hand, the multipath power-delay profile is generally non-uniform [30], where weak paths contribute little energy to the combiner and are susceptible to channel estimation errors [16]. At the same time, relative performance improvement offered by diversity
diminishes with the increase of diversity order [33]. Therefore, reliable path selection for Rake combining at the receiver becomes an important issue in UWB system design.

Several selective Rake combining schemes have been proposed in the literature. Generalized selection combining (GSC) schemes [33], [35], [36] select paths that have instantaneous signal-to-noise ratio (SNR) exceeding a fixed absolute (or normalized) threshold and then combine the selected paths as MRC. However, a fixed threshold even if optimized through numerical simulations is unsatisfactory in severe fading situations [37]. The adaptive threshold for path selection [37], [38] takes into account the noise level estimated by non-coherently averaging multiple measured power delay profiles. In [39], a minimum number of paths are selected as long as the combined SNR maintains the bit error rate (BER) below a certain level, which relies on periodically measuring the BER to provide feedback information. Performance comparable to MRC is obtained in [40] by dividing all received paths into two complementary subgroups and choosing the subgroup providing the larger magnitude of log-likelihood ratio (LLR). However, the above schemes do not save hardware and are undesirable from implementation point of view, because the number of selected paths and hence the required number of Rake fingers may change with time as well as location in the range of 1 to \( N_r \) (i.e. the number of resolvable multipath components) [39], [41]. For a Rake receiver with \( L_c \) fingers, where \( L_c \) is fixed and less than \( N_r \), the SNR-GSC scheme [34], [35] combines the \( L_c \) paths with the largest instantaneous SNR using MRC, and the \( |LLR| \)-GSC scheme [41] selects the \( L_c \) paths with the largest magnitudes of LLR. Nevertheless, the sorting of all received paths by their magnitudes of SNR or LLR requires instantaneous and accurate estimation of the complete multipath channel, while UWB channel estimation is difficult due to the large number of resolvable multipath components and stringent channel estimation conditions [42].

Channel estimation for UWB wireless communications has been addressed by various authors. The maximum-likelihood (ML) approach [42] requires numerical searches over a multi-dimensional space spanned by the likelihood function and thus, is of high computational intensity. The subspace method exploiting the clustering property of the channel [43] or using the second order statistics of the received signal to explicitly remove interference [44] yields good performance at the expense of complexity. Pilot
assisted schemes [45]-[57] send pilot and data-bearing pulses sequentially and attempt to recover the continuous-time equivalent channel from the received pilot pulses for use in an analog correlator to detect the data. Pilot waveform assisted modulation (PWAM) [46] optimizes the number, placement, and energy allocation of pilot pulses assuming the knowledge of the channel coherence time, but incurs a discrepancy between data-bearing pulse energy and pilot pulse energy and the resulting uneven transmission power levels may challenge the power amplifier. It is worth noting that these pilot assisted approaches require the channel to be quasi-static [48], because the channel estimate obtained from the pilot pulses in a transmission burst is used for the detection of all data-bearing pulses in that burst. In addition, the implementation of a programmable analog correlator is of high complexity [48].

1.3 Thesis Research and Contributions

In this thesis, some preliminary knowledge of UWB radio technology and general UWB impulse radio systems are introduced in Chapter 2. In Chapter 3, a power-efficient 2MN-ary biorthogonal keying modulation scheme is developed for UWB direct-sequence impulse radio communications. Following that, performance analysis incorporating the effect of channel-induced pulse overlap for direct-sequence multiple-access impulse radio systems is provided in Chapter 4. In Chapter 5, a pilot-channel assisted modulation scheme, along with a generalized selection combining scheme with log-likelihood ratio threshold test per path, is proposed to provide reliable channel estimation and tracking, and proper path selection for UWB Rake receivers with limited number of Rake fingers. Finally, the conclusions and future plans are presented in Chapter 6. The thesis contributions in more details are given as follows.

In Chapter 3, we develop a multidimensional modulation for UWB impulse radio direct-sequence multiple-access communications under multipath fading. An important advantage of multidimensional modulation over unidimensional modulation is the possibility of higher power-efficiency and this may be critical in UWB wireless communications. For this reason, we propose a 2MN-ary biorthogonal keying scheme, in which N orthogonal pulse shapes in conjunction with M orthogonal code-words are employed to construct an MN-dimensional biorthogonal modulation. The direct-
sequence multiple-access performance of the proposed scheme is evaluated through simulations based on the channel model recommended by IEEE 802.15.3a. In addition, most published work on impulse radio research [11], [15] has focused on systems where the received pulse is the second derivative of a Gaussian function, which does not meet the FCC UWB indoor emission limits [49], i.e. the −10dB-bandwidth must be kept within the band from 3.1GHz to 10.6GHz. Since this thesis uses the current FCC UWB regulations as its guide, orthogonal pulse shapes that comply with the FCC UWB indoor emission limits are introduced and used in the simulations. Our results confirm the power efficiency improvement achieved by multidimensional modulation for UWB multiple-access communications under multipath fading.

In Chapter 4, we provide performance analysis for direct-sequence (DS) multiple-access impulse radio systems operating in multipath environments, where the effect of overlap between pulses arriving at the receiver is included. Pulse overlap is generally ignored in UWB studies but it occurs when the pulse width is larger than the fine multipath delay resolution of an ultra-wideband channel and is particularly important to consider in pulse-based multidimensional modulations. A closed-form bit error rate (BER) expression incorporating the channel-induced pulse-overlap is derived and compared with simulations that employ UWB signals complying with the FCC UWB indoor emission mask and the channel model recommended by IEEE 802.15.3a. The detailed theoretical analysis and simulations reveal that channel-induced pulse overlap has a significant impact on UWB system performance. The results also show that compared with binary modulations, multidimensional biorthogonal pulse keying (BOPK) can improve power efficiency for UWB impulse radio multiple-access systems under multipath fading, thereby providing an advantage for power-limited UWB wireless communications.

For UWB communications with signal energy dispersed by a large number of multipath components, the design of a Rake receiver that can provide a desirable output signal-to-noise-ratio (SNR) using only a moderate number of fingers becomes an important issue. In Chapter 5, we propose a generalized selection combining (GSC) scheme with log-likelihood ratio (LLR) threshold test per path (\(|LLR|-T\)-GSC) to provide reliable path selection for UWB Rake receivers with fixed number of fingers,
based on channel estimation provided by a pilot-channel assisted modulation (PCAM). The pilot channel is constructed with pulses that are orthogonal to the data-bearing pulses. Each pilot pulse is transmitted simultaneously with a data-bearing pulse. The PCAM parameters are optimized through jointly minimizing the channel estimation mean square error (MSE) and maximizing the average receiver output SNR. The proposed PCA-|LLR|-T-GSC scheme alleviates the need to estimate all resolvable multipath components for path selection, by setting an adaptive threshold for the magnitude of LLR per path. Extensive simulations confirm the good performance of PCA-|LLR|-T-GSC in providing desirable performance in fast-fading channels and in the presence of multi-user interference, without sacrificing the overall system data rate.

1.4 Publications

**Journal Papers:**


- **Xiaoli Chu, Ross D. Murch,** "Multidimensional modulation for ultra-wideband multiple-access impulse radio in wireless multipath channels," *IEEE Transactions on Wireless Communications,* accepted.

- **Xiaoli Chu, Ross D. Murch,** "Performance analysis of DS-MA impulse radio communications incorporating channel-induced pulse overlap," *IEEE Transactions on Wireless Communications,* accepted.

- **Xiaoli Chu, Ross D. Murch,** "Pilot-channel assisted generalized selection combining with log-likelihood ratio threshold test per path for Rake reception in ultra-wideband communications," *IEEE Transactions on Communications,* submitted and under review.

**Conference Papers:**

- **Xiaoli Chu, Ross D. Murch,** "Quadrature modulation for UWB wireless multipath


Chapter 2

Preliminaries

In this chapter, we first provide the essential aspects of knowledge of ultra-wideband (UWB) radio technology in Section 2.1, with particular emphasis on impulse radio techniques, and then discuss the main aspects in the design of a UWB impulse radio wireless communications system in Section 2.2, including the channel model that will be adopted throughout this thesis.

2.1 UWB Radio Technology

The use of ultra-wideband (UWB) radio technology dates back to the 1940s where it was initially used for radar systems and then matured in the 1970s for covert military communication. UWB is becoming a commercially applicable technology at this point in time because of three primary reasons [8]: 1) the U.S. FCC approved the technology in February 2002; 2) the technology can be done now, due to advancements in things like silicon germanium and high-speed switching technology, which also lead to lower prices; 3) higher data rates are in demand.

In February 2002, the FCC authorized the commercial deployment of UWB on an unlicensed basis in the 3.1-10.6GHz band subject to a modified version of Part 15.209 rules, where the maximum average power spectral density (PSD) of UWB is limited to –41.25dBm/MHz. The FCC created three classes of UWB devices, each with its own technical standards and operating restrictions [8]: 1) imaging systems including ground penetrating radars (GPRs), medical imaging, and surveillance devices; 2) vehicular radar systems; and 3) communications and measurement systems. Figure 2.1 shows the preliminary FCC and European Telecommunications Standards Institute (ETSI) UWB emission limits for indoor systems, which were designed to curtail the interference that UWB devices might have on GPS, military, ground/air navigation, or cellular applications. Currently, UWB is legal only in the United States. International regulatory bodies are considering possible rules and emission limits that would help enable
worldwide operation of UWB devices.

The FCC defines UWB as any wireless technology with a fractional bandwidth greater than 20% or an absolute bandwidth of at least 500MHz, regardless of the modulation type or transmission method [4]. The fractional bandwidth is defined by \(2(f_H - f_L)/(f_H + f_L)\), where \(f_H\) and \(f_L\) are the upper and lower frequencies at the -10dB emission point, respectively. Because of this broad definition of UWB promulgated by the FCC, the UWB implementation applies to any wireless technology using more than 500MHz bandwidth within the allowed spectral mask [1].

![UWB EIRP emission level diagram](image)

**Figure 2.1** The UWB signal emission spectral mask [8].

Impulse radio communicates with series of intermittent extremely-short pulses, instead of using a continuous carrier signal [10]. For impulse radio, the basic idea is to send a fixed pulse shape at a regular (or controlled) interval. For conventional carrier systems, the "pulse shape" at RF is not fixed due to the lack of fixed relationship between carrier phase and symbol clock phase. If properly designed, the benefits of conventional carrier-based systems and impulse radio systems can be combined – fixed phase relationship between carrier and symbol [82].
Impulse radio is a form of spread-spectrum communication, which is distinguished by the following characteristics: 1) the transmission bandwidth is much greater than the information bit rate and is independent of the information bit rate; 2) demodulation must be accomplished, in part, by correlation of the received signal with a replica of the signal used at the transmitter to spread the information signal. For impulse radio, the spectrum spreading is achieved by the use of pulses with very short duration, typically on the order of a nanosecond, thereby spreading the signal energy over a bandwidth of a few gigahertz. A UWB impulse radio system generates pulses that satisfy the spectral regulations and transmits these pulses at a pulse repetition frequency (PRF) that satisfies the data rate as well as range requirements. Information data can be modulated on the pulses in a number of ways, including amplitude, polarity, time position, and any combination of these. The techniques for introducing pseudorandomness in spread-spectrum signals, like direct sequence (DS), frequency hopping (FH) and time hopping (TH) [32], can also be used for impulse radio signals.

![Signals in Time](image1)

![Signals in Frequency](image2)

**Figure 2.2** Single-band and multi-band systems supporting 200Mb/s. The multi-band system cycles the signals through the 15 bands. Both systems have equivalent SNR [1].
Single-band impulse radio is the traditional approach for UWB implementation, which offers relatively simple radio designs but provides little flexibility for spectrum management [9]. An emerging multiband modulation approach splits the legally available UWB spectrum into multiple sub-bands, each occupying an instantaneous bandwidth of at least 500MHz. The multi-band UWB modulation approach increases flexibilities in modulation, transmission, coexistence, interference suppression, and the efficient use of the legally available spectrum [9]. For a multiband impulse radio system, the pulses occupying different sub-bands can be generated by modulating different "carrier" frequencies with a common baseband pulse that occupies a bandwidth of at least 500MHz [16]. The sub-bands are used sequentially for transmission and the PRF on each sub-band can be lower than the single-band approach. Two extreme systems are shown in Figure 2.2 [1]. The first one uses a single-band approach, where the signal is shaped to occupy the entire 7.5GHz band. The second one uses a multiband approach, where 15 signals, each occupying a 500MHz bandwidth, are shaped to cover the entire band. There are, of course, intermediate cases. The bandwidth and number of available sub-bands generate different performance tradeoffs and implementation challenges [1], [16].

Recently, a number of UWB variants of conventional wideband transmission strategies, like orthogonal frequency division multiplexing (OFDM) [50], [75] and direct-sequence code division multiple access (DS-CDMA) [51], are being considered for application to areas such as WPANs [5].

OFDM is a multi-carrier modulation technique that uses multiple orthogonal subcarriers to modulate data in parallel. One approach to design a UWB system based on OFDM is to combine the OFDM modulation technique with a multiband approach (MB-OFDM) [75], where the spectrum is divided into several sub-bands each with bandwidth of at least 500MHz and each sub-band employs OFDM modulation. The transmitted OFDM symbols are time-interleaved across the sub-bands.

Direct-sequence spread-spectrum technique generates a spread-spectrum signal by modulating a data-modulated carrier signal a second time using a wideband spectrum-spreading signal, which is obtained by using a pseudorandom sequence in conjunction with a phase modulation. CDMA is the type of digital communication in which each
user (transmitter-receiver pair) has a distinct pseudorandom sequence for transmitting information over a common channel bandwidth. Single- and multi-band DS-CDMA strategies have been considered for use in UWB communications [51]. In direct-sequence spread-spectrum impulse radio (DS-UWB) [82], data is spread over multiple pulses using a pseudo random code. Pulse waveform has a UWB spectrum. The chip rate of the spreading code does not need to be so high because the code is used only for user separation, not to spread the spectrum. The DS is also used to randomize the spectrum of the transmitted signal so as to avoid strong spectral lines associated with simple pulse repetition. DS-UWB is based on integer relationship between center frequency (carrier) and chip or symbol rate — makes frequency synthesis and implementation easier. In the recently proposed “DS-UWB physical layer submission to 802.15 task group 3a” [82], the pulse shape that always has the same phase relationship between carrier and pulse is maintained by an integer relationship between chip rate and center frequency (i.e. center frequency is always 3 times of the chip rate), so that the chip rate clock and the chip carrier shall be provided from the same source.

In this thesis and unless otherwise mentioned, we restrict out discussion to those systems that utilize direct-sequence impulse radio technologies.

2.2 Impulse Radio

In this section, we describe main aspects in the design of a UWB impulse radio system: 1) channel capacity; 2) pulse shaping and modulation; 3) multiple assessing; 4) coexistence with narrowband systems; 5) multipath channel characterization; 6) Rake demodulator; and 7) timing synchronization.

2.2.1 Channel Capacity

High channel capacity is one major advantage of UWB. This can be explained using Shannon’s capacity limit equation, which for additive white Gaussian noise (AWGN) channels is given by

\[ C = B \cdot \log_2(1+S/N) \]  

(2.1)

where \( C \) = maximum channel capacity (bits/sec),
\[ B = \text{channel bandwidth (Hz)}, \]
\[ S = \text{signal power (watts)}, \]
\[ N = \text{noise power (watts)}. \]

We can see that the channel capacity scales linearly with bandwidth but only logarithmically with power. Therefore, linear increase of information capacity requires linear increase in bandwidth, while similar capacity increase would require exponential increase in power [2]. This is why UWB technology is capable of transmitting very high data rates at very low power levels.

![Channel Capacity as a function of separation distance](image)

**Figure 2.3** UWB capacity vs. other WLAN technologies [9].

The theoretical capacity of UWB in AWGN channels can also be expressed as a function of range, with the received signal-to-noise ratio (SNR), \( SNR_r(d) \), given by

\[
SNR_r(d) (\text{dB}) = -41.25 - (\text{108}) - L - 20\log_{10}\left(\frac{4\pi f_c}{c}\right) - n10\log_{10}d \tag{2.2}
\]

where \( c = 3\times10^8 \text{m/s} \) is the speed of light, \( n \) is the exponent of path loss, \( d \) is the distance
from the transmitter to the receiver, $L$ (dB) subsumes additional implementation losses, $-41.25$ dBm/MHz is the allowed average emission PSD by Part 15 rules, and 6dB noise figure was applied to the $-114$ dBm/MHz standard thermal noise PSD [16]. With $B = 10.6 - 3.1 = 7.5$ GHz, $n$ equals to 2 (free space) up to 8m and 3.3 (as representative of some indoor wireless channels) beyond, and the center frequency $f_c = (3.1 + 10.6)/2 = 6.85$ GHz, the UWB capacity in an AWGN channel as a function of range is compared with IEEE 802.11a and 802.11g in Figure 2.3 [9]. This figure shows the distinct range-vs.-rate tradeoff provided by UWB systems, which offers the greatest promise for very high data rates with ranges below 10m. The limited range is mainly due to the low power levels allowed by the FCC for legal UWB operation.

### 2.2.2 Pulse Shaping and Modulation

Impulse radio UWB implementation directly modulates impulse-like waveforms. In earlier work, Gaussian monocycles obtained by differentiation of the standard Gaussian waveform have been frequently used as typical baseband UWB pulses for analytical evaluation of UWB systems. The waveform of an idealized monocycle is shown in Figure 2.4 [11]. However, such baseband pulses occupy frequency spectra from near dc up to a few gigahertz, which are unsuited to the current FCC spectral mask and point to the need for alternate pulse shapes [16].

![Figure 2.4](image_url)

**Figure 2.4** An idealized received monocycle $w_r(t)$ at the output of the antenna. The waveform is given by $w_r(t + 0.35) = [1 - 4\pi (t/\tau_m)^2]\exp[-2\pi (t/\tau_m)^2]$ with $\tau_m = 0.2877$ ns [11].
The most popular modulation schemes developed to date for UWB impulse radio include pulse-position modulation (PPM), pulse-amplitude modulation (PAM), on-off keying (OOK), and bipolar pulse modulation [1].

PPM encodes information with two or more positions in time relative to the nominal pulse position. Figure 2.5(a) shows a two-position modulation, where a pulse transmitted at the nominal position represents a 0, and a pulse transmitted after the nominal position represents a 1. Additional positions can be used to encode more bits in one symbol. The time between nominal positions is typically much longer than the pulse width to avoid interference between consecutive pulses.

![Figure 2.5](image)

**Figure 2.5** Four different modulation techniques for UWB pulses, from the top down: (a) PPM, (b) OOK, (c) PAM, and (d) bipolar pulse modulation [19].

PAM encodes information in the amplitudes of the pulses. Figure 2.5(c) shows a two-level amplitude modulation. OOK can be seen as a special case of two-level PAM, as shown in Figure 2.5(b). More amplitude levels can be used for PAM to encode more bits per symbol.

In bipolar pulse modulation, information is encoded with the polarities of the pulses, as shown in Figure 2.5(d). The polarity of the pulse is switched to encode a 0 or a 1.
2.2.3 Multiple Accessing

Thanks to the enormous bandwidth, a UWB system may accommodate many users, even in multipath environments [10]. UWB systems can utilize various methods of multiple-access that are commonly used in traditional wireless systems. For an impulse radio system operating in a fixed single band, either a time-division multiplexing (TDM) or a code-division multiplexing (CDM) multiple-access method can be applied [1]. The option of frequency division multiplexing (FDM) as a multiple-access method is possible with multiband UWB systems [1].

Two approaches to multiple accessing for impulse radio that have been widely discussed in the literature are based on time-hopping (TH) [11] and direct-sequence (DS) encoding [13]. In the DS approach, a binary (±1) user-specific pseudorandom code sequence \( c^k \) (with a period of \( N_x \)) for each particular \( k^{th} \) user is impressed on a train of \( N_x \) pulses, which are sent within a symbol interval. The TH approach shifts the time position of each pulse in units of a chip interval \( (T_c) \) according to the user-specific pseudorandom code sequence \( c^k \), for which each code element \( c^k_j \) is an integer satisfying \( 0 \leq c^k_j < N_b \) and \( N_b T_c \leq T_r \), where \( T_r \) is the nominal pulse repetition period. In this thesis, we will focus on the DS approach, because DS generally supports higher pulse repetition rates (thus higher data rates) and achieves better performance than time hopping (TH) [10] for the same pulse width and spreading code length in the same channel [18].

2.2.4 Coexistence with Narrowband Systems

UWB systems operating over the highly populated frequency range must contend with a variety of interfering signals from incumbent radio systems. According to the current FCC regulations on UWB, it is almost certain that UWB will be affected by 5GHz-band WLANs transmissions, i.e. IEEE 802.11a, which are up to 20dB higher in power than UWB transmissions [7]. The degree of that interference has yet to be determined. At the same time, UWB systems must insure that they do not interfere with narrowband radio systems operating in their dedicated bands. Due to the relatively small bandwidth of conventional radio systems compared with the UWB signal bandwidth, a narrowband
receiver will see the UWB signal as an additive noise with quite flat power spectral density (PSD) [52]. The ability of the UWB receiver to reject narrowband interference and the ability of the UWB transmitter to avoid interfering with other radio systems both depend on the PSD of the UWB signals [10].

2.2.5 UWB Channel Characterization

A reliable channel model, which captures the important characteristics of the channel, is a vital prerequisite for system design. There are distinct differences between UWB and narrowband wireless channels, especially with respect to fading statistics and time of multipath arrivals.

The assumption of flat fading is used when the system bandwidth is small enough that all the multipath components interfere constructively or destructively at the receiver and the delays of the individual multipath components do not impact the system performance. When the signal bandwidth gets larger, the different delays of multipath components may influence the system performance. For a wideband signal covering a bandwidth \( B \), which is larger than the coherence bandwidth of the channel, the channel becomes frequency-selective and the multipath components in the received signal are resolvable with a delay resolution of \( 1/B \) in time [32]. Thus, for system analysis, the delay axis is typically divided into bins with size comparable to the inverse of the signal bandwidth. If there is still interference between the multipath components that fall within each delay bin, the amplitude statistics of the delay bins can be modeled as Rayleigh or Rice. For UWB systems with extremely large bandwidths, only few or even no multipath components fall in a resolvable delay bin, so the central limit theorem is no longer applicable in predicting the fading statistics.

The IEEE 802.15 Task Group 3a has evaluated a number of popular indoor channel models to determine which model best fits the channel characteristics obtained from realistic UWB channel measurements. In particular, the channel measurements showed that multipath arrivals are in clusters rather than in a continuum. This is a result of the fine delay resolution provided by UWB waveforms. The UWB channel measurements also indicated that the fading amplitudes follow either a lognormal or a Nakagami distribution. Consequently, the final model adopted by the IEEE 802.15.3a committee
for the evaluation of UWB physical layer performance was based on a modified Saleh-Valenzuela (S-V) model [53], which also includes a shadowing term to account for the variation of received energy caused by blockage of the line-of-sight path. This section describes the modeling of ultra-wideband wireless propagation channels.

2.2.5.1 Multipath Model

The multipath model consists of the following discrete time impulse response [53]

\[ h_i(t) = X_i \sum_{k=0}^{K} \sum_{i=0}^{K} \alpha(i,k)^i \delta(t - T_i^i - t_{k,i}^i) \]  

(2.3)

where \( \{ \alpha(k,i) \} \) are the multipath gain coefficients, \( \{ T_i^i \} \) is the excess delay of the \( i^{th} \) cluster (i.e. the arrival time of the first path of the \( i^{th} \) cluster), \( \{ t_{k,i} \} \) is the delay of the \( k^{th} \) path within the \( i^{th} \) cluster relative to the cluster arrival time \( T_i^i \), obviously \( t_{0,i} = 0 \), \( \{ X_i \} \) represents the log-normal distributed shadowing, and \( i \) refers to the \( i^{th} \) realization.

There are seven key parameters that define this multipath channel model:

- \( \Lambda \) = cluster arrival rate
- \( \lambda \) = ray arrival rate within a cluster
- \( \Gamma \) = cluster decay factor
- \( \gamma \) = ray decay factor
- \( \sigma_1 \) = standard deviation of cluster lognormal fading term (dB)
- \( \sigma_2 \) = standard deviation of ray lognormal fading term (dB)
- \( \sigma_z \) = standard deviation of lognormal shadowing term for the whole multipath realization (dB)

The distributions of cluster and ray arrival times are given by

\[ P(T_i|T_{i-1}) = \Lambda \exp(-\Lambda(T_i - T_{i-1})) \quad i > 0 \]  

(2.4)

\[ P(t_{k,i}|t_{(k-1),i}) = \lambda \exp(-\lambda(t_{k,i} - t_{(k-1),i})) \quad k > 0 \]  

(2.5)

The log-normal shadowing term \( X_i \) is characterized by

\[ 20 \log_{10}(X_i) \sim \text{Normal}(0, \sigma_z^2) \]  

(2.6)

Independent fading is assumed for each cluster as well as each ray within a cluster. The fading coefficients are defined as follows

\[ \alpha_{k,i} = p_{k,j} \tilde{\xi}_k \beta_{k,i} \]  

(2.7)
20\log_{10}(\xi_\beta_{k,l}) \approx \text{Normal}(\mu_{k,l}, \sigma_1^2 + \sigma_2^2) \text{ or } |\xi_\beta_{k,l}| = 10^{(\mu_{k,l} + n_1 + n_2)/20} \quad (2.8)

\mathbb{E}[|\xi_\beta_{k,l}|^2] = \Omega_0 \exp(-T_0/\Gamma) \exp(-\tau_{k,l}/\gamma) \quad (2.9)

where \( p_{k,l} \) is equiprobably +1 or -1 accounting for signal inversion due to reflections, \( \xi_\beta \) reflects the fading associated with the \( l \)th cluster, \( \beta_{k,l} \) corresponds to the fading associated with the \( k \)th ray within the \( l \)th cluster, \( n_1 \approx \text{Normal}(0, \sigma_1^2) \) and \( n_2 \approx \text{Normal}(0, \sigma_2^2) \) are independent and correspond to the fading on each cluster and each ray, respectively, \( \Omega_0 \) is the mean energy of the first path of the first cluster, and \( \mu_{k,l} \) is given by

\[ \mu_{k,l} = \frac{10\ln(\Omega_0) - 10T_i / \Gamma - 10\tau_{k,l} / \gamma}{\ln(10)} - \frac{(\sigma_1^2 + \sigma_2^2)\ln(10)}{20} \quad (2.10) \]

Since the shadowing effect on the total multipath energy is captured by the term \( X_i \), the total energy contained in the terms \( \{X_{i,l} \} \) is normalized to unity for each realization of the channel impulse response. Also note that complex-valued taps were not adopted in this model, because the complex baseband model only fits for narrowband systems to capture channel behaviors independent of the carrier frequency, while carriers are not necessary for UWB impulse radio systems and thus, real-valued simulations at radio frequency may be more natural [53].

Table 2.1 shows the channel characteristics and corresponding parameters matching measurement results for several channel scenarios [53]. LOS refers to line-of-sight and NLOS refers to non-line-of-sight. Although the output of the above channel model yields continuous time samples (i.e. infinite bandwidth), the results in Table 2.1 are for a time resolution of 0.167ns, corresponding to the 6GHz bandwidth of the underlying measurements. However, different UWB waveforms may be proposed, which may have different bandwidths and thus, different multipath delay resolution. For UWB bandwidths less than 6GHz, to obtain a larger minimum-path-spacing, the channel realizations obtained from the above channel model could be low-pass filtered to the desired bandwidth and re-sampled with a reduced sampling rate. For UWB bandwidths greater than 6GHz, it was suggested to reduce the minimum path spacing to the inverse of the signal bandwidth and use the given model parameters, since the maximum bandwidth allowed by FCC is 7.5GHz, which is not significantly greater than the 6GHz bandwidth used to derive the above model parameters.
Table 2.1 Example multipath channel characteristics and corresponding model parameters.

<table>
<thead>
<tr>
<th>Target Channel Characteristics&lt;sup&gt;5&lt;/sup&gt;</th>
<th>CM 1&lt;sup&gt;1&lt;/sup&gt;</th>
<th>CM 2&lt;sup&gt;2&lt;/sup&gt;</th>
<th>CM 3&lt;sup&gt;3&lt;/sup&gt;</th>
<th>CM 4&lt;sup&gt;4&lt;/sup&gt;</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mean excess delay (nsec) (τ_m)</td>
<td>5.05</td>
<td>10.38</td>
<td>14.18</td>
<td></td>
</tr>
<tr>
<td>RMS delay (nsec) (τ_rms)</td>
<td>5.28</td>
<td>8.03</td>
<td>14.28</td>
<td>25</td>
</tr>
<tr>
<td>NP&lt;sub&gt;10&lt;/sub&gt;dB</td>
<td></td>
<td></td>
<td>35</td>
<td></td>
</tr>
<tr>
<td>NP (85%)</td>
<td>24</td>
<td>36.1</td>
<td>61.54</td>
<td></td>
</tr>
</tbody>
</table>

Model Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>CM 1&lt;sup&gt;1&lt;/sup&gt;</th>
<th>CM 2&lt;sup&gt;2&lt;/sup&gt;</th>
<th>CM 3&lt;sup&gt;3&lt;/sup&gt;</th>
<th>CM 4&lt;sup&gt;4&lt;/sup&gt;</th>
</tr>
</thead>
<tbody>
<tr>
<td>λ (1/nsec)</td>
<td>0.0233</td>
<td>0.4</td>
<td>0.0667</td>
<td>0.0667</td>
</tr>
<tr>
<td>γ</td>
<td>7.1</td>
<td>5.5</td>
<td>14.00</td>
<td>24.00</td>
</tr>
<tr>
<td>σ&lt;sub&gt;1&lt;/sub&gt; (dB)</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
</tr>
<tr>
<td>σ&lt;sub&gt;2&lt;/sub&gt; (dB)</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
</tr>
<tr>
<td>σ&lt;sub&gt;3&lt;/sub&gt; (dB)</td>
<td>3</td>
<td>3</td>
<td>3</td>
<td>3</td>
</tr>
</tbody>
</table>

Model Characteristics<sup>5</sup>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>CM 1&lt;sup&gt;1&lt;/sup&gt;</th>
<th>CM 2&lt;sup&gt;2&lt;/sup&gt;</th>
<th>CM 3&lt;sup&gt;3&lt;/sup&gt;</th>
<th>CM 4&lt;sup&gt;4&lt;/sup&gt;</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mean excess delay (nsec) (τ_m)</td>
<td>5.0</td>
<td>9.9</td>
<td>15.9</td>
<td>30.1</td>
</tr>
<tr>
<td>RMS delay (nsec) (τ_rms)</td>
<td>5</td>
<td>8</td>
<td>15</td>
<td>25</td>
</tr>
<tr>
<td>NP&lt;sub&gt;10&lt;/sub&gt;dB</td>
<td>12.5</td>
<td>15.3</td>
<td>24.9</td>
<td>41.2</td>
</tr>
<tr>
<td>NP (85%)</td>
<td>20.8</td>
<td>33.9</td>
<td>64.7</td>
<td>123.3</td>
</tr>
<tr>
<td>Channel energy mean (dB)</td>
<td>-0.4</td>
<td>-0.5</td>
<td>0.0</td>
<td>0.3</td>
</tr>
<tr>
<td>Channel energy std (dB)</td>
<td>2.9</td>
<td>3.1</td>
<td>3.1</td>
<td>2.7</td>
</tr>
</tbody>
</table>

<sup>1</sup> This model is based on LOS (0-4m) channel measurements.
<sup>2</sup> This model is based on NLOS (0-4m) channel measurements.
<sup>3</sup> This model is based on NLOS (4-10m) channel measurements.
<sup>4</sup> This model is generated to fit a 25ns RMS delay spread to represent an extreme NLOS multipath channel.
<sup>5</sup> These characteristics are based on a 0.167ns sampling time.

2.2.5.2 Path Loss Model

The current path-loss model adopted by the IEEE 802.15 Task Group 3a uses a simple free-space path-loss formula [53], i.e. \( PL = (4\pi d/\lambda)^2 \), where the wavelength \( \lambda \) is computed at the center frequency (geometrical mean of upper and lower −10dB cutoff frequency) of the system. More realistic path loss models are required to anticipate typical UWB system performance.

2.2.5.3 Time Variability of Channel

The above IEEE 802.15.3a channel model assumes that the channel stays either
completely static, or changes completely from one data burst (about 100µs) to the next. While this covers extreme cases, some important aspects, like adaptive channel estimation and interleaving, require a model incorporating time variance. The impulse response of a time-varying wireless channel can be modeled as [32]

\[ h(t, \tau) = \sum_{l} h(t) \delta(\tau - \tau_l) \]  

(2.11)

where \( \tau \) and \( h(t) \) are the delay and path gain of the \( l \)th path, respectively, and \( h(t, \tau) \) is the impulse response at time \( t \) due to an impulse applied at time \( \tau \). Assuming the total multipath spread is \( T_d \), for all practical purposes the tapped-delay-line model can be truncated at \( N_r = \lfloor T_d/B \rfloor + 1 \) taps [32].

To characterize the time variance of a channel, the Doppler power spectrum is useful only if \( h(t, \tau) \) can be modeled as a wide-sense stationary (WSS) random process in the variable \( t \) [54]. The time variance of a UWB channel due to the relative movement of transmitter and/or receiver can be treated using the wide sense stationary model with uncorrelated scattering (WSSUS). The Doppler spectrum is related to the angular power spectrum (APS) as follows [55]

\[ S_D = \Omega_0 \left[ \text{pdf}_G(\gamma)G(\gamma) + \text{pdf}_G(-\gamma)G(-\gamma) \right] \frac{1}{\sqrt{f_d^2 - f^2}}, \quad \text{for} \ |f| < f_d \]  

(2.12)

where \( \text{pdf}_G(\gamma) \) is the arriving angular power spectrum, i.e. the probability density function of power arriving at the angle \( \gamma \) with respect to the (quasi)-LOS component, \( G(\gamma) \) is the antenna pattern, and \( f_d \) is the maximum Doppler frequency. Since the shape of the angular power spectrum is not exactly known, it was suggested in [55] to model the angular spectrum by a simple rectangular function with a delay-dependent width as follows

\[ \text{pdf}_G(\gamma, \tau) = \begin{cases} \text{rect}(2\pi\tau / \tau_{\text{max}}), & 0 < \tau < \tau_{\text{max}} \\ 1/(2\pi), & \tau > \tau_{\text{max}} \end{cases} \]  

(2.13)

in which the parameter \( \tau_{\text{max}} \) depends on the power delay profile.

When the temporal variations stem primarily from the movement of scatterers, the WSSUS model can no longer be applied, because the assumption of stationarity is violated – the moving scatterer has a significant angular extent such that the channel may switch between LOS and NLOS characteristics. When the scatterers are moving,
the angular power spectrum becomes time-variant itself, and the description becomes considerably more complicated [55]. In that case a geometrical model (blocking off rays from a certain angular range) can be used for simulation [54].

2.2.5.4 Summary of Channel Model

The channel model suggested by the IEEE 802.15.3a committee was an important step for the understanding of UWB channels, but it is not a universal stochastic model of the UWB wireless propagation channel. A lot of work on channel modeling will have to be spent before our understanding of UWB wireless channels is complete.

2.2.6 Rake Demodulator

The transmitted energy of UWB signals is distributed among a large number of multipath components, implying that some form of energy combining is needed to capture a sufficient amount of signal energy for reliable detection [16].

For a frequency-selective channel that can be modeled by a tapped delay line with statistically independent tap weights, the optimum demodulator employs correlation (or matched filtering), where the locally generated reference signals are properly delayed and correlated with the received signal [32]. An alternative realization of the optimum demodulator employs a single delay line through which is passed the received signal, which is correlated at each tap with reference signals. The demodulator output is sampled at the symbol rate and the samples are passed to a decision circuit. In effect, the tapped delay line demodulator attempts to collect the signal energy from all the received multipath components that carry the same information and fall within the span of the delay line and thus, it is called “Rake” demodulator [32]. The Rake demodulator with perfect (noiseless) estimates of the channel tap weights is equivalent to a maximal ratio combiner, which weights the signal on each diversity path in proportion to the branch signal-to-noise-ratio (SNR) before combining [16].

For UWB systems, the ability to finely resolve individual multipath components and Rake combine the energy of them at the receiver can greatly boost the receiver output SNR [9]. However, this is achieved at the expense of increasing the receiver complexity.
The perfect Rake receiver with unlimited resources (correlators) and infinitely fast adaptability is not realizable. In fact, as the multipath delay resolution of a radio system becomes finer, performance approaching that of a perfect Rake demodulator becomes more difficult to achieve [10]. With increasing desired bit rates, the cost/complexity of the Rake will become an important determinant in the cost/complexity tradeoffs in the UWB transceiver design [16].

2.2.7 Timing Synchronization

Timing synchronization is another important issue in system and receiver design, in terms of the preamble resources and the receiver complexity devoted to it, respectively [9]. Timing synchronization is challenging for impulse radio signals due to the strict power limitation and the extremely short pulse duration [71].

The conventional technique to determine the relative delay of the received signal with respect to the receiver template signal, is the serial search algorithm [72]. The received signal is correlated with the template signal and the correlation output is compared with a threshold. If the output is lower than the threshold, the template signal is shifted by an amount in time comparable to the pulse duration and is correlated with the received signal again. This procedure continues until an output exceeds the threshold. However, this conventional synchronization technique based on pulse-rate sliding correlation is not only very slow to converge, due to the prohibitively large number of possible pulse positions must be searched in order to locate the narrow pulses employed in impulse radio UWB systems, but also suboptimal in the presence of dense multipath [71]. In order to shorten the acquisition time, parallel acquisition with multiple correlators has been proposed. Nevertheless, full parallel search schemes increase the hardware complexity and the power consumption of the receiver [73]. In view of this, hybrid schemes have been proposed to trade off the speed of parallel schemes with the simplicity of serial search schemes [74].
Chapter 3

Multidimensional Modulation for Impulse Radio

In this chapter, we investigate the potential of improving the power-efficiency of UWB impulse radio communications by increasing the dimensionality of signaling waveforms. Motivation for the use of multidimensional modulation stems from the fact that multidimensional signaling increases the minimum Euclidean distance and the cutoff rate by increasing the dimensionality of signals, for which the number of available dimensions increases linearly with the time-bandwidth product [57]. On the other hand, for UWB communications bandwidth is less of a concern and bandwidth efficiency (bits/s/Hz) can be relatively low. However, one challenge in increasing the modulation dimensionality is to design multiple orthogonal waveforms with short effective time durations and with frequency content contained to the legally allowed spectral band.

Our work is different in that we use orthogonal pulse shapes in conjunction with waveform coding to increase the modulation dimensionality and to reduce the restriction induced by orthogonal pulse shaping at the same time. We suggest waveform coding in the multidimensional modulation because the spread-spectrum nature of impulse radio enables the use of channel codes with relatively low coding rates [58]. Biorthogonal coding is chosen because of its superior performance to other waveform coding procedures [59] and its inherent antipodal signaling characteristics that can smooth the envelope of the signal PSD [19]. Making use of $N$ orthogonal pulses and $M$ ($M \geq 2$) orthogonal Walsh codes [60], we propose a $2MN$-ary biorthogonal keying (BOK) scheme, which antipodal modulates $MN$ orthogonal waveforms to represent $2MN$ information symbols. When $N = 1$ and $M = 1$, BOK becomes bipolar pulse amplitude modulation (PAM) [17]; when $N > 1$ and $M = 1$, BOK becomes a biorthogonal pulse keying (BOPK) scheme; when $N = 1$ and $M > 1$, BOK changes to the conventional biorthogonal-code keying (BOCK) [59]. The direct-sequence (DS) multiple-access (MA) performance of the proposed scheme is evaluated in indoor multipath environments. The power efficiency is measured in the required $E_b/N_0$ to achieve the desired
rate/performance operating point [16]. We also include conventional bipolar PAM [17], [18] and biorthogonal-code keying [59] in simulations for comparison. Orthogonal pulses are provided coinciding with the FCC UWB indoor mask [49] and are used in the simulations.

The remainder of this chapter is organized as follows. In Section 3.1, the 2MN-ary biorthogonal keying (BOK) scheme in its DS-MA format and the construction of orthogonal UWB pulses are presented. In Section 3.2, the UWB channel model and the corresponding receiver structure are introduced, followed by a discussion on the resultant multiple-access performance. Results of simulations are presented in Section 3.3. Conclusions are drawn in Section 3.4.

3.1 Multidimensional Modulation for Impulse Radio

3.1.1 Multidimensional Modulated Direct-Sequence Multiple-Access

We form orthogonal signaling waveforms by using \( N \) orthogonal pulse shapes (we discuss their construction in Section 3.1.2) in conjunction with \( M \) (\( M \geq 2 \)) orthogonal code words so that \( MN \) orthogonal-coded pulse sequences (each consisting of \( M \) pulses) can provide \( MN \) signal dimensions, where both \( M \) and \( N \) are positive integers. Using these \( MN \) orthogonal waveforms we wish to construct power-efficient modulation. Three candidates are orthogonal, biorthogonal and simplex modulations [59]. Of these we select biorthogonal modulation because the minimum distance between constellation points of biorthogonal modulation is larger than that of the others, when compared as a function of the energy per information bit (\( E_b \)). The inherent antipodal characteristic of biorthogonal signals that can smooth the envelope of the signal PSD [19] is an additional feature for the selection of biorthogonal modulation. With the above-mentioned \( MN \) orthogonal waveforms along with the negative version of each of them [59], we propose a 2MN-ary biorthogonal keying (BOK) scheme for DS-MA UWB impulse radio systems.

Specifically, we consider a DS-MA UWB system consisting of \( N_u \) asynchronous users. The 2MN-ary BOK waveform of a particular \( u^{th} \) user can be expressed as

\[
s^{(u)}_{t}(t) = \sqrt{\frac{E^{(u)}_{\text{r}}}{N_\text{r}}} \sum_{k=-\infty}^{\infty} c^{(u)}_{k} \sum_{k/N_u} w_{k}^{(u)} \langle \langle d^{(u)}_{k,N_u} | M \rangle \langle k | M \rangle \rangle P^{(u)}_{k,N_u} (t - kT_f) \tag{3.1}
\]
where $b_{k/N}^{(u)} \in \{\pm 1\}$ and $d_{k/N}^{(u)} \in \{0, 1, ..., MN-1\}$ represent the 2MN-ary data stream, $\langle k | M \rangle = k \mod M$, $p_m(t)$ $(n = 0, \ldots, N-1)$ are $N$ orthogonal unit-energy pulses occupied in $[0, T_w]$, $T_f \geq T_w$ is the pulse repetition time, $c_k^{(u)} \in \{\pm 1\}$ and $E_s^{(u)}$ are the pseudorandom signature sequence and symbol energy of the $u^{th}$ user, respectively. The orthogonal modulating codes $[w_{m,0}, \ldots, w_{m,M-1}]$ $(m = 0, 1, \ldots, M-1)$ are generated by selecting as code words the rows of an $M \times M$ Hadamard matrix and substituting 0 with +1 and 1 with $-1$ [60]. Thus, 2MN biorthogonal signals are composed by the $MN$ orthogonal waveforms, $w_{\tilde{m}}(t) = \sum_{k=0}^{M-1} w_{\tilde{m}|M} \cdot p_{\tilde{m}/M} \cdot (t - kT_f)$ $(\tilde{m} = 0, 1, \ldots, MN-1)$, along with their negative versions. These signaling waveforms are repeated $G$ times per symbol period to provide a processing gain and hence the number of pulses transmitted per symbol period is $N_s = G \cdot M$. Note that this $N_s$ factor is what allows impulse radio to operate at a very low average transmit power spectral density while still achieving useful throughput and range [2]. In addition, the low duty cycle (i.e. $T_f > T_w$) provides a pulse processing gain of $G_p = T_f/T_w$ [13]. With each 2MN-ary symbol transmitted over $N_s$, pulse-repetition intervals, the bit rate is $R_b = (\log_2 2MN)/(N_s T_f)$.

### 3.1.2 Generation of Orthogonal Pulses

Here we present two classes of orthogonal pulses that can accommodate the current FCC UWB regulations. First, UWB pulses have been widely modeled using Gaussian functions in the literature [11], [28], [61], and we can generate two orthogonal pulses based on time derivatives of a generic Gaussian function, the time- and frequency-domain functions of which are given by

$$p_0(t) = \exp\left[-2\pi \left(\frac{t}{\tau_m}\right)^2\right] \xrightarrow{FT} P_0(f) = \frac{\sqrt{2\pi} \tau_m}{2} \exp\left[-\frac{\pi}{2} (\tau_m f)^2\right]$$

(3.2)

where $\tau_m$ is a parameter that determines the pulse width. By taking two successive derivatives of $p_0(t)$, we get two pulses $p_n(t)$ and $p_{n+1}(t)$ being even (odd) and odd (even) functions respectively, where $p_n(t) = \frac{d^n p_0(t)}{dt^n}$ is the $n^{th}$ derivative of $p_0(t)$ and its Fourier
transform is \( P_n(f) = (j2\pi f)^n P_0(f) \). With properly selected pulse order \( n \) and \( \tau_m \), we can construct orthogonal-pulse pairs with frequency content contained in the FCC UWB indoor emission mask [62]. As an example, time-domain waveforms and frequency spectra of the two Gaussian pulses \( p_4(t) \) and \( p_5(t) \) for \( \tau_m = 0.182\text{ns} \) are illustrated in Figure 3.1(a) and Figure 3.1(b), respectively.

Second, nine orthogonal modified Hermite pulses have been proposed in [12]. Although these pulses occupy frequency bands not allowed by the current FCC UWB regulations [4], they are still potentially suitable for multi-dimensional modulations. We now shift them to the FCC permitted frequency band by first defining a pulse-width parameter \( \tau_m \) along with the following functions

\[
\tilde{v}_n(t) = (-1)^{n-1} e^{-\frac{(it \tau_m)^2}{4}} \frac{d^{n-1}}{dt^{n-1}} \left( e^{-\frac{(t \tau_m)^2}{4}} \right)
\]

\[
\tilde{v}_{n+1}(t) = \frac{t}{2\tau_m^2} \tilde{v}_n(t) - \frac{d\tilde{v}_n(t)}{dt} \xrightarrow{FT} \tilde{V}_{n+1}(f) = j \frac{1}{4\pi^2 \tau_m^2} \frac{d\tilde{V}_n(f)}{df} - j2\pi f \tilde{V}_n(f)
\]

where \( n (= 1, 2, \ldots) \) denotes the pulse order, \( \tilde{V}_n(f) \) is the Fourier transform of \( \tilde{v}_n(t) \).

Then frequency shift can be applied to get the time- and frequency-domain expressions as follows

\[
v_n(t) = \tilde{v}_n(t) \cos(2\pi f_c t) \xrightarrow{FT} V_n(f) = \frac{1}{2} \tilde{V}_n(f - f_c) + \frac{1}{2} \tilde{V}_n(f + f_c)
\]

where the nominal center frequency \( f_c \) and \( \tau_m \) are decided by the required spectral mask. For example, the four pulses \( \{v_1(t), v_2(t), v_3(t), v_4(t)\} \) with \( f_c = 6.55\text{GHz} \) and \( \tau_m = 0.1\text{ns} \) coincide with the FCC spectral requirements, as shown in Figure 3.2. Eight orthogonal pulses \( \{v_1(t), v_2(t), \ldots, v_8(t)\} \) in the FCC-permitted band can be obtained with \( f_c = 6.8\text{GHz} \) and \( \tau_m = 0.13\text{ns} \). Because the frequency shift is embedded in the pulse generation functions, in the lack of a continuous sinusoidal carrier the resulting pulses are still impulsive, but the receiver pulse generator may need a clock recovery circuit to perfectly reproduce the pulse.
Figure 3.1 (a) Time-domain waveforms of the Gaussian pulses $p_1(t)$ and $p_2(t)$ for $r_m = 0.182$ns, where the pulse amplitudes are normalized to give unity energy. (b) Power spectra of the Gaussian pulses $p_1(t)$ and $p_2(t)$ for $r_m = 0.182$ns, where the FCC UWB indoor emission mask is also plotted for comparison.
Figure 3.2 (a) Temporal waveforms of four orthogonal modified Hermite pulses \( \{v_1(t), \ldots, v_4(t)\} \) given by (3.5) with \( f_c = 6.55 \text{GHz} \) and \( \tau_m = 0.1 \text{ns} \), where the pulse amplitudes are normalized to give unit energy. (b) Frequency spectra of \( \{v_1(t), \ldots, v_4(t)\} \) with \( f_c = 6.55 \text{GHz} \) and \( \tau_m = 0.1 \text{ns} \), the FCC UWB indoor emission mask is also plotted.
For all these orthogonal pulse types, an important consideration is the orthogonality between the received pulses when they are time dispersed by the multipath channel, which will be taken into account in our numerical simulations. We will also provide detailed discussion on this in Chapter 4.

3.2 Multipath Propagation & Rake Reception

3.2.1 The UWB Indoor Multipath Channel

There are distinct differences between UWB and narrowband wireless channels, especially with respect to the fading statistics [30]. Due to the extremely large bandwidths of UWB signals, only few or even no multipath components fall in each delay bin, so the amplitude statistics of the delay bins are lognormal [30] rather than Rayleigh that is common to narrowband channels [32]. Lognormal fading leads to smaller variations of the instantaneous amplitudes than Rayleigh fading, and the UWB channel power delay profile is generally not monotonous [30]. These have significant impact on UWB system performance [29], [31], [76], [77].

The channel model used in this study is the one recommended by the IEEE 802.15.3a channel modeling subcommittee, as we have introduced in Section 2.2.5. This channel model is based on measurements in the 2-8GHz band and thus has a path resolution of $\Delta \tau = 0.167\text{ns}$ [30], for which the channel impulse response is defined by

$$h(t) = \sum_{j=0}^{L} \sum_{k=0}^{K} \alpha_{j,k} \delta(t - T_j - \tau_{j,k}) = \sum_{l=0}^{N_r-1} h_l \delta(t - l\Delta \tau)$$

where $\alpha_{j,k} = p_{j,k}^{\beta_{j,k}}$ are path gain coefficients, $p_{j,k} \in \{\pm 1\}$ denotes the random polarity with equal probability and the fading amplitude $\beta_{j,k}$ is lognormal-distributed. $T_j$ is the delay of the $j^{th}$ cluster, and $\tau_{j,k}$ is the delay of the $k^{th}$ path within the $j^{th}$ cluster relative to $T_j$. $T_j$ and $\tau_{j,k}$ are described with a double-Poisson process and all of them are rounded to integer multiples of the delay resolution $\Delta \tau$. Thus with $T_d$ denoting the multipath delay spread and $N_r = \lceil T_d / \Delta \tau \rceil$, $h_l$ is the sum of all $\alpha_{j,k}$ with $T_j + \tau_{j,k} = l\Delta \tau$. Due to the clustering of multipath components [30], the channel does not necessarily have multipath arrivals within each delay bin. This is accounted for by setting $h_l = 0$ for any delay bin $l\Delta \tau$ that

34
has no path arrives.

3.2.2 Receiver Structure

In the asynchronous multiple-access system under consideration, we assume that all the $N_u$ active users experience the same multipath environment characterized by the channel model $h(t) = \sum_{l=0}^{N_r-1} h_l \delta(t - l\Delta\tau)$, but the statistical behavior of the multipath experienced by each $u^{th}$ $(u = 1, \ldots, N_u)$ user is independent, identically distributed (i.i.d.) and can be represented by the vector $\mathbf{h}^{(u)} = [h_0^{(u)} \ h_1^{(u)} \ \ldots \ h_{N_r-1}^{(u)}]^T$. The signal at the output of the receiver antenna is given by

$$r(t) = \sum_{u=1}^{N_u} \sum_{l=0}^{N_r-1} h_l^{(u)} s_t^{(u)}(t - l\Delta\tau - \tau_u) + n(t)$$  \hspace{1cm} (3.7)

where the asynchronous multiple-access delays $\tau_u$ are independent for different users, $n(t)$ denotes the AWGN with double-sided PSD of $N_0/2$, and $s_t^{(u)}(t)$ has the same expression as $s_t(t)$ except that the transmitted pulse $p(t)$ has been changed to $p(t)$ when passing through the channel due to its wideband distorting effects and also the responses of the transmitter and receiver antennas. Suggested models for the distortion induced by antenna systems include differentiation [10], but other effects will also be present [78] making the exact distortion difficult to characterize.

In this thesis, we will assume that the orthogonality between the received pulse shapes can be retained by appropriately pre-distorting the transmitted signals [78], i.e.,

$$\int_0^T p_m(t)p_n(t)dt = \delta(m-n) \text{ for } m, n \in \{0, 1, \ldots, N-1\}.$$  \hspace{1cm} (3.8)

In practice, however, it is likely that the pre-distortion will only approximately produce the desired received pulses and there will likely be some loss in performance. Alternatively, the receiver may be able to use training information to better determine the effects of the channel and the antennas on the received pulse shapes [46]. Under the assumption that orthogonality is preserved among the received pulses, the receiver for the $2MN$-ary BOK consists of $MN$ correlators each using as correlation template one of the $MN$ possible received
waveforms \( w_{\tilde{m}}(t) = \sum_{k=0}^{M-1} w_{(\tilde{m}M),k} P_{(\tilde{m}M),k} (t - kT_f) \) \( (\tilde{m} = 0, 1, \ldots, MN-1) \). Decisions are made by selecting the correlator output with the largest magnitude and by acquiring the sign (positive or negative) information of the selected output. We also employ a maximum ratio combining (MRC) partial-Rake structure [56], which consists of a DS despreader that correlates the received signal with the desired user's appropriately delayed signature sequence and \( L_c \) fingers that lock the first \( L_c \) paths of the channel output from the desired user, and followed by a combining circuit. We assume that the receiver knows the exact received pulse shapes and has perfect estimates of the channel fading coefficients.

Note that the system complexity increases when more orthogonal waveforms are employed for 2MN-ary BOK. Therefore, the number of dimensions that can be used in signaling is restricted by practical implementation considerations. However, since the correlation output at each \((m+nM)^{th}\) \( (m = 0, \ldots, M-1) \) branch at the receiver is the sum of negative or positive outputs of the same pulse correlator that uses \( p_{n}(t) \) \( (n \in \{0, \ldots, N-1\}) \) as template, the receiver structure can be simplified such that only \( N \) pulse correlators and \( M \) pulse-polarity-control-and-sum circuits (corresponding to the \( M \) orthogonal Walsh codes) are required.

### 3.2.3 Multiple-Access Performance

For a specific user, we take the signals from other users as interference. Without loss of generality, we assume that the receiver has locked to the signal from the 1\( st \) user. That is, the receiver knows \( \tau_1 \) and has an appropriately delayed replica of the signature sequence \( \left\{_{(1)}^{(1)k} \right\} \). Assuming the \( q^{th} \) symbol \( d_q^{(1)} = 0 \) and at the branch of the receiver that uses \( w_{0r}(t) = \sum_{k=0}^{M-1} w_{0,k} p_{0r}(t - kT_f) \) as the correlation template waveform, the correlator output corresponding to the \( 1^{th} \) pulse-repetition interval and the \( j^{th} \) \( (j = 0, 1, \ldots, N_r-1) \) path is given by

\[
y_0(i,j) = \int_{0}^{T_f} r(t + iT_f + j\Delta \tau + \tau_1)_{(1)}^{(1)i} w_{0,(i|M)} p_{0r}(t) \, dt
\]  

(3.8)
where $\langle i|M \rangle = i \pmod{M}$ and we drop the super-index $^{(1)}$ from $y_0(i, j)$ to simplify notation.

Defining $y_0(i) = [y_0(i,0) \ y_0(i,1) \ \ldots \ y_0(i,N_r-1)]^T$ as the correlation output vector corresponding to the $i^{th}$ pulse-repetition interval, the MRC decision statistic for the $q^{th}$ symbol can be expressed by

$$y_0(q) = \sum_{i=qN_s}^{qN_s+N_s-1} [h^{(1)}_i]^H \Gamma y_0(i) = y_3(q) + y_{MP}(q) + y_{MUI}(q) + y_\eta(q) \tag{3.9}$$

in which $\Gamma$ represents the partial-Rake [56] that combines the first $L_c \leq N_r$ paths of the channel output and is defined by the following $N_r \times N_r$ matrix

$$\Gamma = \begin{bmatrix} I_{L_c} & 0 & \cdots & 0 \\ 0 & 0 & \cdots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ 0 & \cdots & 0 & 0 \end{bmatrix} \tag{3.10}$$

In (3.9), $y_3(q)$ and $y_{MP}(q)$ both result from the signal of the 1st user but denote the desired signal part and the multipath interference (MPI) part caused by the multipath propagation, respectively, $y_{MUI}(q)$ stands for the multi-user interference (MUI), and $y_\eta(q)$ is given by

$$y_\eta(q) = \sum_{i=qN_s}^{qN_s+N_s-1} [h^{(1)}_i]^H \Gamma \eta(i) \tag{3.11}$$

where $\eta(i) = [\eta(i,0) \ \eta(i,1) \ \ldots \ \eta(i,N_r-1)]^T$ is the output-noise vector corresponding to the $i^{th}$ pulse-repetition interval, in which $\eta(i, j) = \int_0^{T_s} n(t + iT_f + j\Delta r + r_1)X_0(i, j|\mathcal{M})P_{\text{re}}(t)dt$.

It is easy to show that $y_\eta(q)$ is a zero-mean Gaussian random process with variance of

$$\text{var}[y_\eta(q)] = \frac{N_0}{2} N_s \sum_{i=0}^{L_c-1} |h^{(1)}_i|^2 \tag{3.12}$$

Since there is some dependency among the components of the multipath interference (MPI) as well as the multi-user interference (MUI), it is not tractable to derive the exact expression of the error probability. It is natural to invoke Gaussian approximations on MPI and MUI to evaluate the bit error rate (BER), using the union bounding technique and the $Q$-function [32]. As equally-likely and equal-energy biorthogonal signals, it is
easy to show that the symbol error rate (SER) of the proposed 2MN-ary BOK conditioned on the channel parameters can be upper bounded as follows [59]

\[
P_{\epsilon, \text{BOK}} \leq (2MN - 2)Q \left( \frac{N_s E_s \left( \sum_{l=0}^{L-1} |h_l(0)|^2 \right)^2}{2 \text{var}[n_{\text{total}}]} \right) + Q \left( \frac{N_s E_s \left( \sum_{l=0}^{L-1} |h_l(0)|^2 \right)^2}{\text{var}[n_{\text{total}}]} \right) \tag{3.13}
\]

which becomes increasingly tight for fixed M and N as the signal-to-noise-ratio (SNR) increases [59], where \( n_{\text{total}} = y_{\text{MPI}} + y_{\text{MU1}} + y_{\eta} \) represents the total noise. The BER is a complicated function of SER for biorthogonal signals, but we can approximate it with the relationship \( P_{\epsilon, \text{BOK}} = \frac{1}{2} P_{\epsilon, \text{BOK}} \). This approximation is quite good for \( 2MN > 8 \) [59]. The unconditional BER can be approximated by averaging the conditional BER's over multiple channel realizations.

### 3.3 Numerical and Simulation Results

#### 3.3.1 Orthogonal Pulses

We generate two orthogonal Gaussian pulses that coincide with the FCC UWB indoor mask [49], by choosing the two pulses \( p_4(t) \) and \( p_5(t) \) and setting \( \tau_m = 0.182 \text{ns} \). Time-domain waveforms and frequency spectra of them are illustrated in Figure 3.1(a) and Figure 3.1(b), respectively. It can be observed that the durations of \( p_4(t) \) and \( p_5(t) \) (\( \tau_m = 0.182 \text{ns} \)) are around 0.5ns, while the occupied bandwidth of them is about 7GHz (corresponding to a multipath delay resolution of 0.143ns). Figure 3.3 shows correlations values of \( p_4(t) \) and \( p_5(t) \) sampled at integer multiples of the path resolution time (\( \Delta \tau = 0.167 \text{ns} \)) of the channel model recommended by IEEE 802.15.3a [30]. Comparing Figure 3.3(a) with Figure 3.3(b), we can see that \( p_5(t) \) appears preferable to \( p_4(t) \), because the sampled autocorrelation of \( p_5(t) \) is more peaky than that of \( p_4(t) \). Therefore, for modulations where only one pulse shape is required, e.g. bipolar PAM or biorthogonal-code keying [59], \( p_5(t) \) is used in simulations to model the received pulse.

For the four pulses \( \{v_1(t), v_2(t), v_3(t), v_4(t)\} \) with \( f_c = 6.55 \text{GHz} \) and \( \tau_m = 0.1 \text{ns} \) as shown in Figure 3.2, the sampled correlations of them are shown in Figure 3.4, where
Correlations values are also sampled at integer multiples of the multipath delay resolution (0.167ns). We can observe that the orthogonal modified Hermite pulses and their cross-correlations are non-zero for longer durations compared with the two Gaussian pulses $p_4(t)$ and $p_5(t)$ ($\tau_m = 0.182$ns). As we will show in Chapter 4, when the effective duration of the modulated pulse is larger than the multipath delay resolution of the channel, arriving pulses may overlap at the receiver. The longer the pulse duration is the more possible cross-correlations of the template waveform with overlapping interference pulses, which can degrade the system performance significantly. Considered in this context, only the orthogonal Gaussian pulses have sufficiently short durations and will be used in the numerical simulations here.

![Diagram](image)

**Figure 3.3** (a) Autocorrelation $R_{p_4}(t)$ of the Gaussian pulse $p_4(t)$ and cross-correlation $R_{p_5p_4}(t)$ between $p_5(t)$ and $p_4(t)$, sampled at integer multiples of the multipath delay resolution (0.167ns). (b) Autocorrelation $R_{p_5}(t)$ of $p_5(t)$ and cross-correlation $R_{p_4p_5}(t)$ between $p_4(t)$ and $p_5(t)$, sampled at integer multiples of 0.167ns.
Figure 3.4 (a) Autocorrelation $R_{v_1}(t)$ of the modified Hermite pulse $v_1(t)$ and cross-correlations between $v_1(t)$ and $\{v_2(t), v_3(t), v_4(t)\}$, sampled at integer multiples of the multipath delay resolution (0.167ns). (b) Autocorrelation $R_{v_2}(t)$ of $v_2(t)$ and cross-correlations between $v_2(t)$ and $\{v_1(t), v_3(t), v_4(t)\}$. (c) Autocorrelation $R_{v_3}(t)$ of $v_3(t)$ and cross-correlations between $v_3(t)$ and $\{v_1(t), v_2(t), v_4(t)\}$. (d) Autocorrelation $R_{v_4}(t)$ of $v_4(t)$ and cross-correlations between $v_4(t)$ and $\{v_1(t), v_2(t), v_3(t)\}$.

3.3.2 Simulation Results

Now we present simulation results based on the channel model recommended by IEEE 802.15.3a and with the received pulses as shown in Figure 3.1. Channel model parameters are chosen corresponding to a NLOS channel CM3 (with mean excess delay $\tau_M = 14.18\text{ns}$, root mean square (RMS) delay $\tau_{RMS} = 14.28\text{ns}$, and number of paths capturing 85% of the channel energy NP(85%) = 61.54) given in Table 2 of [30]. We assume that the Rake receiver [56] perfectly synchronizes to the signal from the desired
user and collects all resolvable multipath components. The channel coefficients are normalized so that the total energy of the resolved paths is unity in order to remove the path loss and the loss in Rake combining. Strict power control is assumed. The elements of signature sequences $\{c_k^{(u)}\}$ for $u = 1, \ldots, N_u$ and any $k$ were generated as i.i.d. random variables uniformly distributed in $\{\pm 1\}$. The asynchronous delays $\tau_u$ ($u = 1, \ldots, N_u$) were i.i.d. random variables with $\tau_u \text{mod } T_j$ uniformly distributed on $[0, T_j)$. Possible cross-correlations between the received pulses when they are time-dispersed by the multipath channel are taken into account in the simulations. The BER is obtained by averaging conditional BER's over 1000 channel realizations. For a fair comparison of different modulation schemes, $E_b/N_0$ is used.

Figure 3.5 shows the BER performance versus $E_b/N_0$ for $2MN$-ary BOK with $N = 2$ (i.e. two orthogonal pulses are employed) and various values of $M$ (the size of orthogonal codes). Figure 3.6 presents results for $2M$-ary BOK (i.e. $N = 1$, which is equivalent to the conventional biorthogonal-code keying [59] applied on a single UWB pulse shape). Both Figure 3.5 and Figure 3.6 also include bipolar PAM [17] (with $N_s = 64$ and $T_j = 5$ns) for comparison, under the same condition of single-user ($N_u = 1$) and $R_b = 3.125$Mbps. We choose $N_s = 64$ for all of the $2MN$-ary BOK modulations and their values of $T_j$ are determined accordingly by $R_b = (\log_2 2MN)/N_s T_j = 3.125$Mbps. Figure 3.7 and Figure 3.8 show the BER versus $E_b/N_0$ for $2MN$-ary BOK with $N = 2$ and with $N = 1$ (biorthogonal-code keying), respectively, under the same conditions as those of Figure 3.5 and Figure 3.6, except for the presence of $N_u = 40$ asynchronous users.
Figure 3.5 BER vs. $E_b/N_0$ of $2MN$-ary BOK modulations that use $M$ Walsh codes and $N = 2$ orthogonal pulses, for $N_x = 1$ and $R_b = 3.125$Mbps ($N_z = 64$). The results assumed $p_d(t)$ and $p_f(t)$ in Figure 3.1 as the received pulses. The simulated BER of bipolar PAM is also plotted for comparison.

Figure 3.6 BER vs. $E_b/N_0$ of $2M$-ary BOK modulations that use $M$ Walsh codes on a single pulse shape, for $N_x = 1$ and $R_b = 3.125$Mbps ($N_z = 64$). The results assumed $p_f(t)$ in Figure 3.1 as the received pulse. The simulated BER of bipolar PAM is also plotted for comparison.
Figure 3.7 BER vs. $E_b/N_0$ of $2M$-ary BOK modulations that use $M$ Walsh codes and $N = 2$ orthogonal pulses, for $N_a = 40$ and $R_b = 3.125$Mbps ($N_r = 64$). The results assumed $p_d(t)$ and $p_b(t)$ in Figure 3.1 as the received pulses. The simulated BER of bipolar PAM is also plotted for comparison.

Figure 3.8 BER vs. $E_b/N_0$ of $2M$-ary BOK modulations that use $M$ Walsh codes on a single pulse shape, for $N_a = 40$ and $R_b = 3.125$Mbps ($N_r = 64$). The results assumed $p_b(t)$ in Figure 3.1 as the received pulse. The simulated BER of bipolar PAM is also plotted for comparison.
According to the simulation results provided in Figures 3.5 to 3.8, Table 3.1 gives the required $E_b/N_0$ (dB) for a target BER of $10^{-3}$ and $R_b = 3.125$Mbps ($N_z = 64$). The values of $T_f$ corresponding to different modulations are also shown. We can observe that in both cases of $N_u = 1$ and $N_u = 40$, all the listed multidimensional modulations achieve better power efficiencies than bipolar PAM. For $2MN$-ary BOK, the modulation with larger value of $MN$ provides better performance. This is due to its inherent orthogonal signaling properties, for which the BER can be made as small as desired by increasing the number of orthogonal waveforms as long as $E_b/N_0 > -1.59$dB [32]. Comparing the $E_b/N_0$ values of the $N_u = 1$ case with those of the $N_u = 40$ case, we can observe that $2MN$-ary BOK ($N = 2$) modulations suffer from less degradation caused by MUI, compared with the conventional biorthogonal-code keying (i.e. $2MN$-ary BOK with $N = 1$) modulations and bipolar PAM. This is due to the fact that for $2MN$-ary BOK ($N = 2$) the interference caused by pulses that are orthogonal to the desired pulse is reduced by the pulse correlation process at the receiver through the use of $N = 2$ orthogonal pulses. Although the orthogonality between the pulses may be lost when they are time dispersed by the channel, the effect of channel induced pulse overlap (or the loss of orthogonality) is not significant. This is because the pulse duration ($T_w = 0.5$ns) of $p_4(t)$ and $p_5(t)$ is not very large compared with the multipath delay resolution ($\Delta \tau = 0.167$ns) and the auto- and cross-correlations of $p_4(t)$ and $p_5(t)$ appear relatively desirable, as shown in Figure 3.3. In addition, the resistance to MUI is more evident for higher-level multidimensional modulations. One reason for this is that since these higher-level modulations carry more bits per symbol longer pulse repetition time $T_f$ can be used – makes the signal more multipath resistant and reduces collisions from other users. This also explains why 4-ary BOK performs slightly better than bipolar PAM.

From the above results, we can see that the advantages of the proposed $2MN$-ary BOK scheme include: more power efficient than bipolar PAM [17]; more power efficient than conventional biorthogonal-code keying [59] in the presence of multiple-access interference, specifically, $2MN$-ary BOK ($N = 2$) that employs Walsh codes of size $M$ and two orthogonal pulse shapes is more power-efficient than $4M$-ary biorthogonal-code keying that uses Walsh codes of size $2M$. However, in the use of both $N$ (number of orthogonal pulses) and $M$ (size of orthogonal codes) to increase the
modulation dimensionality, larger values of $N$ are restricted by the challenge in orthogonal pulse shaping and possible difficulties in synchronization and channel estimation. In addition, since the receiver requires $N$ pulse correlators and $M$ pulse-polarity-control-and-sum circuits, the level of $2MN$-ary BOK modulation is also restricted by practical implementation considerations.

Table 3.1 $E_b/N_0$ Required for BER = $10^{-3}$ and $R_b = 3.125$Mbps ($N_r = 64$), Obtained through Simulations. (The results assumed $p_d(t)$ and $p_s(t)$ in Figure 3.1 as the received pulses for $2MN$-ary BOK ($N = 2$), and $p_s(t)$ as the received pulse for $2MN$-ary BOK ($N = 1$) and bipolar PAM, where $M$ is the size of orthogonal Walsh codes and $N$ is the number of orthogonal pulses.)

<table>
<thead>
<tr>
<th>Modulation Schemes</th>
<th>$E_b/N_0$ (dB)</th>
<th>$T_f$(ns)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$N_s = 1$</td>
<td>$N_s = 40$</td>
</tr>
<tr>
<td>$2MN$-ary BOK</td>
<td></td>
<td></td>
</tr>
<tr>
<td>($N = 2$)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$M = 2$</td>
<td>$2MN = 8$</td>
<td>5.8</td>
</tr>
<tr>
<td>$M = 4$</td>
<td>$2MN = 16$</td>
<td>5.0</td>
</tr>
<tr>
<td>$M = 16$</td>
<td>$2MN = 64$</td>
<td>4.2</td>
</tr>
<tr>
<td>$M = 64$</td>
<td>$2MN = 256$</td>
<td>3.7</td>
</tr>
<tr>
<td>$2MN$-ary BOK</td>
<td></td>
<td></td>
</tr>
<tr>
<td>($N = 1$)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$M = 2$</td>
<td>$2MN = 4$</td>
<td>6.7</td>
</tr>
<tr>
<td>$M = 4$</td>
<td>$2MN = 8$</td>
<td>5.9</td>
</tr>
<tr>
<td>$M = 8$</td>
<td>$2MN = 16$</td>
<td>5.1</td>
</tr>
<tr>
<td>$M = 32$</td>
<td>$2MN = 64$</td>
<td>4.2</td>
</tr>
<tr>
<td>Bipolar PAM</td>
<td>6.8</td>
<td>9.4</td>
</tr>
</tbody>
</table>

3.4 Conclusions

We have proposed a multidimensional modulation scheme to efficiently utilize the UWB spectrum. The analysis and simulation results show that the proposed $2MN$-ary BOK scheme can provide more power-efficient modulation for UWB impulse radio DS-MA communications under multipath fading, compared with conventional bipolar PAM and biorthogonal-code keying. Using a small size (e.g. $M = 2$ or 4) of orthogonal codes and $N = 2$ orthogonal pulse shapes, the $2MN$-ary BOK scheme shows promise for both performance and practical implementation.
Chapter 4

Performance Analysis Incorporating Channel-Induced Pulse Overlap for DS-MA Impulse Radio Systems

In this chapter, we provide performance analysis for direct-sequence (DS) multiple-access (MA) UWB systems operating in multipath environments when the effect of overlap between pulses arriving at the receiver is included. Pulse overlap is generally ignored in UWB studies but it occurs when the pulse width is larger than the multipath delay resolution of the UWB channel. For binary-modulated UWB signals we will show that the effect of pulse overlap can be made insignificant by using narrow pulses, but for multi-dimensional pulse modulations where longer-duration pulses cannot be avoided we will show that it is very important to consider the effect of pulse overlap. We also use these results to show that compared with binary modulations, multidimensional modulations can achieve power efficiency improvement for multiple-access UWB systems under multipath fading, thereby providing an advantage for power-limited UWB communications. This study focuses on DS multiple access, because DS generally supports higher pulse repetition rates (thus higher data rates) and achieves better performance than time hopping (TH) [10] for the same pulse width and spreading code length in the same channel [18].

Our work is different in that we are concerned about the possible pulse overlap caused by the fine multipath delay resolution of impulse radio signals, which are generated in our study using pulses coinciding with the FCC UWB indoor emission mask [49]. Our analysis approach permits explicit studies of the impact of channel-induced pulse overlap on DS-MA system performance. The modulation scheme under consideration is $2M$-ary biorthogonal pulse keying (BOPK) based on the use of $M$ orthogonal pulses, because the BOPK achieves power efficiency [57] as well as a smooth envelope of PSD [19], but our analysis is not restricted to this. Including the loss of orthogonality between pulses when they are time dispersed by the multipath channel, we derive a closed-form bit error rate (BER) expression for BOPK. The analytical
results are then compared with results of extensive simulations, under the channel model recommended for use in IEEE 802.15.3a evaluations [30].

The remainder of this chapter is organized as follows. In Section 4.1, the BOPK modulated DS-MA impulse radio system and the corresponding receiver structure are presented. Section 4.2 provides the detailed performance analysis. Simulations results are presented in Section 4.3. Conclusions are drawn in Section 4.4.

4.1 DS-MA Impulse Radio System

4.1.1 DS-MA Signal Format

We consider a $2M$-ary biorthogonal pulse keying (BOPK) scheme, based on the use of $M$ orthogonal pulses (we have discussed their construction in Section 3.1.2) along with their negative versions, because BOPK provides power efficiency [57] and a smooth envelope of PSD [19]. Bipolar PAM is a special version of BOPK when $M = 1$. For a DS-MA UWB system, where each of $N_u$ users is assigned with a unique pseudorandom sequence, the $2M$-ary BOPK signal waveform of a particular $u^{th}$ ($u = 1, ..., N_u$) user is given by

$$s_u(t) = \frac{E_u}{N_s} \sum_{k=-\infty}^{\infty} c_{u,k} b_{u,k} w_{d_{u,k}}(t - kT_f) \quad (4.1)$$

where we define $\tilde{k} = \lfloor k / N_s \rfloor$, $b_{u,k} \in \{\pm 1\}$ and $d_{u,k} \in \{1, 2, ..., M\}$ represent the $2M$-ary data stream, $c_{u,k} \in \{\pm 1\}$ and $E_u$ are the signature sequence and symbol energy of the $u^{th}$ user, respectively, $w_{m}(t)$ ($m = 1, ..., M$) are $M$ unit-energy orthogonal pulses that occupy $[0, T_w]$, these signaling pulses are repeated $N_s$ times per symbol period to provide a processing gain, $T_f$ ($\geq T_w$) is the pulse-repetition time, the low duty cycle (i.e. $T_f > T_w$) provides a pulse processing gain [13], [22], and the bit rate is $R_b = (\log_2 2M) / (N_s T_f)$.

4.1.2 Receiver Structure

In the multiple-access system under consideration, the channel for all users is characterized by (3.6) in Section 3.2.1, but the multipath fading experienced by each
user is independent, identically distributed (i.i.d.) [63] and is represented by the vector \( \mathbf{h}_u = [h_{u,0}, h_{u,1}, \ldots, h_{u,N_u-1}]^T \) \( (u = 1, \ldots, N_u) \). For high-rate indoor communications, we can assume the channel is stationary over many symbols for each user [63]. The composite signal at the output of the receiver is given by

\[
    r(t) = \sum_{u=1}^{N_u} \frac{E_u}{N_s} \sum_{l=0}^{N_s-1} h_{u,l} \sum_{k=-\infty}^{\infty} c_{u,k} b_{u,k} w_{d_{u,k}} (t - kT_f - l\Delta \tau - \tau_u) + n(t) \tag{4.2}
\]

where \( \tau_u \) is the multiple-access delay of the \( u \)-th user, \( n(t) \) denotes the additive white Gaussian noise (AWGN) with a double-sided PSD of \( N_0/2 \). Note that the pulse shapes may change when passing through the channel due to its wideband distorting effects and also the responses of the transmitter and receiver antennas. Suggested models for the distortion include differentiation [10], but other effects will also be present [78] making the exact distortion difficult to characterize. Here we will assume that the orthogonality between the received pulse shapes can be retained by appropriately pre-distorting the transmitted signals [78]. i.e.,

\[
    \int_0^\infty w_m(t) w_n(t) dt = \delta(m-n) \text{ for } m, n \in \{1, \ldots, M\}.
\]

In practice, however, it is likely that the pre-distortion will only approximately produce the desired received pulses and there will likely be some loss in performance. Alternatively, the receiver may be able to use training information to better determine the effects of the channel and the antennas on the received pulse shapes [46].

With the received pulses \( \{w_1(t), \ldots, w_M(t)\} \) being a set of orthogonal pulses, the receiver for \( 2M \)-ary BOPK signals consists of a Rake structure [32] and \( M \) correlators each using one of the \( M \) possibly received pulses \( w_m(t) \) \( (m = 1, \ldots, M) \) as correlation template. Decisions are made by selecting the correlator output with the largest magnitude and acquiring the sign (positive or negative) information of the selected output. Conventional Rake structures [32] are based on the path resolvability assumption that the minimum path spacing is larger than the signal autocorrelation time, but path resolvability cannot be ensured in UWB indoor communications due to the relatively small multipath delay resolution [30]. Therefore, we employ a Rake structure with resolution reduction (RR) of the fingers [79], which consists of \( L_c \) fingers that are spaced less than one pulse duration apart and synchronize with the first \( L_c \) paths of the channel.
output from the desired user, and a DS despreader that correlates the received signal with an appropriately delayed signature sequence of the desired user. Since the system complexity increases when more orthogonal pulse shapes are used, the level of BOPK modulation is restricted by practical implementation considerations.

### 4.2 Performance Analysis

#### 4.2.1 Output of Rake

Without loss of generality, we assume the 1\textsuperscript{st} user is the desired user and the receiver has synchronized with signals from the 1\textsuperscript{st} user, i.e. the receiver knows $\tau_1$ and has an appropriately delayed replica of \{$c_{1,k}$\}, and its $q$\textsuperscript{th} transmitted symbol $d_{1,q} = n$ ($n \in \{1, \ldots, M\}$). At the receiver branch that uses $w_n(t)$ as correlation template, the Rake output for the $q$\textsuperscript{th} transmitted symbol is

$$ y_n(q) = S_n(q) + \text{MPI}_n(q) + \text{MUI}_n(q) + n_n(q) $$ (4.3)

where $y_n(q) = [y_n(q,0), y_n(q,1), \ldots, y_n(q,L_c-1)]^T$ is the Rake output vector corresponding to the $q$\textsuperscript{th} transmitted symbol, $y_n(q, j) = \sum_{i=qN_i}^{qN_i+N_i-1} \int_{t} E_b \int_{t}^{t_j} r(t + jT_f + j\Delta \tau + \tau_i) c_{1,i} w_n(t) dt$ for $j = 0, 1, \ldots, L_c-1$, and in the absence of AWGN $y_n(q,j)$ is given by

$$ \frac{qN_i+N_i-1}{N_i} \sum_{i=qN_i}^{qN_i+N_i-1} \sum_{k=0}^{N_i-1} h_{u,l} c_{1,k} b_{u,k} w_{d_{u,k}} \left( t - (k-i)T_f - (l-j)\Delta \tau - (\tau_u - \tau_i) \right) c_{1,i} w_n(t) dt $$

$$ \sqrt{E_b} \sum_{i=qN_i}^{qN_i+N_i-1} h_{j,k} c_{1,k} b_{j,k} w_{d_{j,k}} \left( t - (k-i)T_f - (l-j)\Delta \tau \right) w_n(t) dt $$

$$ \sqrt{E_b} \sum_{i=qN_i}^{qN_i+N_i-1} \sum_{l=1}^{L_c-1} h_{u,l} c_{1,l} b_{u,l} w_{d_{u,l}} \left( t - (k-i)T_f - (l-j)\Delta \tau \right) w_n(t) dt $$

which shows that the template pulse $w_n(t)$ can be correlated with not only the pulses to which the receiver is synchronized but also other pulses from the desired user and pulses from other users. Thus in (4.3), $S_n(q)$ and $\text{MPI}_n(q)$ both result from the signal of the 1\textsuperscript{st} user but denote the desired signal part and the part caused by multipath interference.
(MPI), respectively, \( \text{MUI}_n(q) \) is due to multiuser interference (MUI), and \( \eta_n(q) \) is due to the AWGN \( n(t) \).

Due to the resolution reduction (RR), the interference and noise components at the Rake output are correlated [79]. In this case, the optimal combiner is the minimum mean-square-error combiner (MMSEC), for which the combining coefficients are given by \( \mathbf{h}^H_{1:t_e} \mathbf{R}^{-1} \) [79], where \( \mathbf{R} \) is the correlation matrix of the interference and noise components of the Rake output \( y_n(q) \) in (4.3), and is defined by \( \mathbf{R} = \mathbb{E}\{\mathbf{Z}\mathbf{Z}^H\} \) and \( \mathbf{Z} = \text{MPI}_n(q) + \text{MUI}_n(q) + \eta_n(q) \). However, the correlation matrix \( \mathbf{R} \) is not available to the receiver in practice. The detection technique for fading channels with unresolved multipath components proposed in [80] consists of a decorrelation stage and then implements the decision rule for a resolved multipath channel using the parameters of the transformed signal and channel, but the decorrelation operation is too complex to implement for UWB channels that have a large number of multipath components. Therefore, we will focus our analysis on the use of maximum ratio combining (MRC) Rake receiver with resolution reduction [79].

Denoting by \( y_n(q) \) the MRC decision variable for the \( q^{th} \) transmitted symbol, we obtain that

\[
y_n(q) = \mathbf{h}^H_{1:t_e} y_n(q) = S_n(q) + \text{MPI}_n(q) + \text{MUI}_n(q) + \eta_n(q)
\]  

(4.5) 

in which \( \mathbf{h}_{1:t_e} = [h_{1,0} \ h_{1,1} \ \cdots \ h_{1,t_e-1}]^T \) contains the \( L_c \) perfectly estimated channel coefficients of the 1st user for the \( L_c \)-finger MRC Rake with resolution reduction. Given \( d_{1,q} = n \) is transmitted, we calculate that

\[
S_n(q) = \mathbf{h}^H_{1:t_e} \sqrt{\frac{E_1}{N_s}} \sum_{l=qN_l}^{qN_l+N_l-1} \mathbf{h}_{1,t_e} b_{l,n} \int_0^{T_f} w_{d_{1,l}}(t) w_n(t) dt
\]

\[
= b_{1,n} \sqrt{N_s E_1} \sum_{l=0}^{L_c-1} h_{1,l}^2
\]

(4.6) 

At the receiver branches that use correlation templates other than \( w_n(t) \), e.g. \( w_x(t) \) (\( x \in \{1, \ldots, M\} \) and \( x \neq n \)), the MRC decision variable \( y_x(q) \) is given by

\[
y_x(q) = \mathbf{h}^H_{1:t_e} y_x(q) = \text{MPI}_x(q) + \text{MUI}_x(q) + \eta_x(q)
\]

(4.7)
in which \( y_x(q, j) = \sum_{i=qN_x}^{qN_x+N_x-1} \int_0^{\tau_i} r(t + iT_j + j\Delta \tau + \tau_1) c_{1_x} w_x(t) dt \), and \( S_x(q) = 0 \) for \( \int_0^{\tau_i} w_n(t) w_x(t) dt = 0 \).

### 4.2.2 Discrete-Time System Model

In accordance with the discrete channel model where all multipath components arrive at integer multiples of the multipath delay resolution \( \Delta \tau \), as described in Section 3.2.1, we build a discrete model for the DS-MA UWB system by assuming that all multiple-access delays (\( \tau_u \) for \( u = 1, 2, \ldots, N_u \)) and the pulse repetition time \( T_f \) are integer multiples of \( \Delta \tau \). Specifically, we set \( T_f = N_f \Delta \tau \) with \( N_f \) being a positive integer.

If the channel delay spread \( T_d \) is larger than \( T_f \), channel responses to consecutively transmitted pulses will overlap. In order to describe the effect of channel-response overlap, we construct the discrete-time channel matrix for the \( u^{th} \) user as follows

\[
H_{u,0} = \begin{bmatrix}
    h_{u,0} & \cdots & h_{u,N_f} & h_{u,0} & \cdots & 0 \\
    \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
    h_{u,N_x-1} & \cdots & h_{u,N_f} & h_{u,0} & \cdots & 0 \\
0 & h_{u,2N_f} & h_{u,N_f} & h_{u,0} & \cdots & \cdots \\
    \vdots & \vdots & \ddots & \vdots & \ddots & \vdots \\
    h_{u,3N_f} & \cdots & h_{u,2N_f} & h_{u,0} & \cdots & \cdots \\
    \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & h_{u,0} & \cdots & \cdots \\
    \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & h_{u,N_x-1} & \cdots & \cdots \\
    \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & h_{u,N_x-1} & \cdots & \cdots \\
\end{bmatrix}
\]  

(4.8)

where the size of \( H_{u,0} \) is \( N_r \times (2\alpha+1) \), \( \alpha = \left\lceil \frac{N_r}{N_f} \right\rceil - 1 \) is the single-side span of the channel-response overlap, and \( N_r = \left\lceil \frac{T_d}{\Delta \tau} \right\rceil \) is assumed the same for all \( N_u \) users for simplicity. The middle column of \( H_{u,0} \) is the channel response to the currently considered pulse transmitted from the \( u^{th} \) user, the left \( \alpha \) columns represent tails of channel responses to \( \alpha \) prior pulses transmitted from the \( u^{th} \) user, and the right \( \alpha \) columns are due to \( \alpha \) posterior pulses that have channel responses occur within the delay spread of the currently considered pulse. Each row of \( H_{u,0} \) contains multipath
components from the \( u \)th user that arrive at the same time. Since the 1st user is the desired user, the Rake receiver has been synchronized with the middle column of \( \mathbf{H}_{1,0} \), and \( \mathbf{H}_{u,0} \) contains multipath arrivals from the \( u \)th user occurring at the same time instants as the multipath components from the 1st user that are synchronized by the Rake receiver.

At the same time, according to (4.4), the output of the pulse correlator depends on the delay differences between the template pulse and the received pulses, which may be less than the pulse duration \( T_w \) and lead to pulse overlap. Letting \( N_w = \lceil T_w/\Delta \tau \rceil \), we define the following matrices, \( \mathbf{H}_{u,j} \) (\( j \) is an integer, \( 1-N_w \leq j \leq N_w-1 \) and \( j \neq 0 \)), to formulate the pulse overlap at the receiver:

1) For \( 1 \leq j \leq N_w-1 \), \( \mathbf{H}_{u,j} \) contains multipath arrivals from the \( u \)th user occurring \( j \Delta \tau \) later in time than the multipath components that are synchronized by the Rake receiver but within the correlator integration time of the synchronized pulses, and is obtained by up-shifting the components of \( \mathbf{H}_{u,0} \) by \( j \) steps and zero-padding as follows

\[
\mathbf{H}_{u,j} = \begin{bmatrix}
    h_{u,0\times-j} & \cdots & h_{u,N_{j-1}} & h_{u,j} & 0 & \cdots & 0 \\
    \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\
    h_{u,N_{j-1}} & h_{u,N_{j-1}} & 0 & \cdots & 0 \\
    0 & h_{u,2N_{j}} & h_{u,N_{j}} & h_{u,0} & \cdots & 0 \\
    \vdots & \vdots & \vdots & \ddots & \ddots & \ddots \\
    0 & \cdots & 0 & h_{u,\alpha_{N_{j}}} & h_{u,(\alpha-1)N_{j}} & \cdots & h_{u,0} \\
    \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
    0 & \cdots & 0 & h_{u,N_{j-1}} & h_{u,\alpha_{N_{j}}-N_{j}} & \cdots & h_{u,\alpha_{N_{j}}-N_{j-1}} \\
    \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
    0 & \cdots & 0 & 0 & h_{u,\alpha_{N_{j}}-N_{j}+j} & \cdots & h_{u,\alpha_{N_{j}}-N_{j}+j-1}
\end{bmatrix}
\] (4.9)

2) For \( 1-N_w \leq j \leq -1 \), \( \mathbf{H}_{u,j} \) consists of multipath arrivals from the \( u \)th user occurring \( j \Delta \tau \) earlier in time than the multipath components that are synchronized by the Rake receiver but have tails within the correlator integration time of the synchronized pulses, and is obtained by down-shifting the components of \( \mathbf{H}_{u,0} \) by \( j \) steps and zero-padding as follows
\[
H_u,j = \begin{bmatrix}
    h_{u,0N_j + j} & \cdots & h_{u,N_j + j} & 0 & 0 & \cdots & 0 \\
    \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \vdots \\
    h_{u,0N_j - 1} & \cdots & h_{u,N_j - 1} & 0 & 0 & \cdots & 0 \\
    \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \vdots \\
    h_{u,0N_j} & \cdots & h_{u,N_j} & h_{u,0} & 0 & \cdots & 0 \\
    \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \vdots \\
    0 & \cdots & h_{u,2N_j} & h_{u,N_j} & h_{u,0} & \ddots & \vdots \\
    \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \vdots \\
    0 & \cdots & 0 & h_{u,0N_j} & h_{u,(\alpha-1)N_j} & \cdots & h_{u,0} \\
    \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \vdots \\
    0 & \cdots & 0 & h_{u,N_j + j - 1} & h_{u,N_j - N_j + j - 1} & \cdots & h_{u,N_j - \alpha N_j + j - 1}
\end{bmatrix}
\]

Using the above definitions, we obtain the following expression for MPI\(_n(q)\) in (4.5)

\[
MPI_n(q) = \sqrt{\frac{E_r}{N_S}} \sum_{i=qN_1}^{qN_x+qN_1-1} h_{1,i,L_c}^H \sum_{j=1-N_w}^{N_1-1} H_{1,j,L_c} d_1(i, t - j\Delta\tau) c_{1,j} w_n(t) dt
\]

\[
= \sqrt{\frac{E_r}{N_S}} h_{1,i,L_c}^H \sum_{j=1-N_w}^{N_1-1} H_{1,j,L_c} \sum_{i=qN_1}^{qN_1+N_w-1} c_{1,j} \int_0^{T_c} d_1(i, t - j\Delta\tau) w_n(t) dt
\]

\[
= \sqrt{\frac{E_r}{N_S}} h_{1,i,L_c}^H \sum_{j=1-N_w}^{N_1-1} H_{1,j,L_c} \sum_{i=qN_1}^{qN_1+N_w-1} c_{1,j} g_{1,n}(i, j)
\]

where \(H_{u,j,L_c}\) is the \(L_c\times(2\alpha+1)\) matrix composed of the upper \(L_c\) rows of \(H_{u,j}\), we set the middle column of \(H_{1,0,L_c}\) to an all-zero vector for the calculation of MPI, and define the \((2\alpha+1)\times1\) vector

\[
d_u(i, t) = \begin{bmatrix}
c_{u,i-a} b_{u,(i-a)} w_d(i-a) \cdots c_{u,i+\alpha} b_{u,(i+\alpha)} w_d(i+\alpha) \quad \cdots 
\end{bmatrix}^T
\]

(4.12)

to represent the DS-spread data that contribute to the channel-response overlap and thus

\[
g_{u,n}(i, j) = \begin{bmatrix}
c_{u,i-a} b_{u,(i-a)} R_{a,n}(j\Delta\tau) \cdots c_{u,i+\alpha} b_{u,(i+\alpha)} R_{a,n}(j\Delta\tau) \quad \cdots 
\end{bmatrix}^T
\]

(4.13)
in which \(R_m(\tau) = \int_0^\infty w_m(t - \tau) w_n(t) dt\) is the correlation function of two overlapping pulses.

The expression for the \(MU_l\) component in (4.5) is derived in Appendix A and is given as below.
\[ MUI_n(q) = \mathbf{h}_{1,I_u}^H N_s \sum_{u=2}^{N_u} \left( E_u \sum_{j=1}^{N_s-1} q^{N_s-1-j} \sum_{i=qN_s}^{i+j-1} c_{i,j} \int_{i+qN_s}^{i+j} f_{u,i,j} \left( i, t - j\Delta \tau \right) w_n(t) dt \right) \] (4.14)

in which \( f_{u,i,j} \left( i, t \right) \) is given by (A.4) in Appendix A. \( MPI_n(q) \) and \( MUI_n(q) \) of (4.7) have the same expressions as \( MPI_n(q) \) and \( MUI_n(q) \), respectively, but with the subscript "n" replaced by "s".

### 4.2.3 Bit Error Probability

Now we derive the BER expression for DS-MA BOPK systems that use MRC Rake with resolution reduction (RR) at the receiver. The BER is usually obtained by averaging the conditional BER (conditioned on the instantaneous signal-to-interference-and-noise-ratio (SINR) at the receiver output) over the probability density function (PDF) of the instantaneous SINR. Since the UWB channel fading coefficient amplitudes are lognormal distributed [30], the exact closed-form expression for the PDF of the receiver output SINR does not exist [76], but under the assumption of multipath resolvability and with the absence of MPI and MUI, the instantaneous receiver output SNR can be considered as a sum of independent lognormal random variables and hence its PDF can be obtained by approximating it as another lognormal random variable [76]. However, in channels with unresolved multipath components and in the presence of MPI and MUI, the instantaneous receiver output SINR is a rather complicated function of lognormal random variables as we will show in the following, and its PDF is hard to obtain. In this case, we will derive the BER expression conditioned on a fixed set of \( \{ h_u \} \) \( u = 1, \ldots, N_u \), and the unconditional BER can be obtained by averaging the conditional BERs over multiple channel realizations.

Given the symbol "+n" \( n \in \{ 1, \ldots, M \} \) is sent, the conditional symbol error rate (SER) can be bounded using the union bounding technique [32]

\[
P_s(e|+n,h) = P_s(e|h_{1,q} = +1, d_{1,q} = n, h) \leq 2 \sum_{x=1, x\neq n}^{M} \left( P(y_x(q) > y_n(q), y_n(q) > 0, h) + P(y_n(q) < 0| h) \right)
\] (4.15)

where the factor 2 accounts for the biorthogonal modulation.
First, we model $y_n(q)$ as a Gaussian random variable to calculate the probability $P(y_n(q)<0 \mid \mathbf{h})$. The validity of the Gaussian assumption is maintained because, the DS-MA system under consideration is based on the use of long pseudorandom sequences and hence $c_{u,k} \in \{\pm 1\}$ for any $u$ and $k$ can be modeled as i.i.d. random variables [18], data symbols are independent and equiprobable, each transmitted pulse has a large number of multipath components, each combined path experiences different multipath interference as shown by $\mathbf{H}_{u,k} \mathbf{d}_u(i,t)$, and the Rake fingers are uncorrelated with all interfering multipath arrivals, thus, $MPI_n(q)$ in (4.11) and $MUI_n(q)$ in (4.14) are the sums of $N_s L_w (2N_w-1)(2\alpha+1)$ and $(N_s-1)N_s L_w (2N_w-1)(2\alpha+1)$ independent random variables, respectively. It is possible to show that $N_s L_w (2N_w-1)(2\alpha+1)$ is of a sufficiently large value in the DS-MA UWB systems under consideration and the Liapounoff condition [81] is satisfied by both $MPI_n(q)$ and $MUI_n(q)$. Therefore, $MPI_n(q)$ and $MUI_n(q)$ can be well approximated as Gaussian random variables following the Liapounoff theorem [81].

The validity of the Gaussian assumptions on the MPI and MUI terms has also been justified through numerical simulations in [77]. In addition, it is easy to show that $\eta_n(q)$ is zero-mean Gaussian with $\text{var}[\eta_n(q)] = \frac{N_s}{2} \sum_{i=0}^{L-1} h_i^2$.

For $MPI_n(q)$, we find that $E[\text{MPI}_n(q)] = 0$ and its variance is given by

$$\text{var}[\text{MPI}_n(q)] = \frac{E[\text{MPI}_n(q)]}{N_s} h_{i,t}^H \left( \sum_{j=1}^{N_s-1} \mathbf{H}_{i,j} \mathbf{H}_{i,j}^H \right) h_{i,t}$$

(4.16)

Since $d_{i,v} = n$, for $v \in \{i-\alpha, \ldots, i, \ldots, i+\alpha\}$ and $i \in \{qN_s, \ldots, qN_s+N_s-1\}$, $d_{i,v} = n$ if $\tilde{\nu} = \lfloor \nu/N_s \rfloor = q$, otherwise $E[d_{i,v}] = \sum_{m=q}^{m=M} \frac{M}{m}$, and according to the expression of $\mathbf{g}_{1,n}(i,j)$ in (4.13), $\mathbf{E} \left[ \sum_{i=qN_s}^{i=qN_s+N_s-1} \mathbf{g}_{1,n}(i,j) \mathbf{g}_{1,n}(i,j)^H \right]$ is calculated to be a $(2\alpha+1) \times (2\alpha+1)$ diagonal matrix $\mathbf{G}(n,j)$, the diagonal elements $(k \in \{1, \ldots, 2\alpha+1\})$ of which are given by

$$G_{kk}(n,j) = \begin{cases} (N_s - k) R_{nn}^2 (j \Delta \tau) + \frac{k}{M} \sum_{m=1}^{M} R_{mn}^2 (j \Delta \tau) & \text{if } k = |k - \alpha - 1| < N_s, \\ \frac{N_s}{M} \sum_{m=1}^{M} R_{mn}^2 (j \Delta \tau) & \text{if } k = |k - \alpha - 1| \geq N_s. \end{cases}$$

(4.17)

Invoking average over the $2\alpha+1$ diagonal elements of $\mathbf{G}(n,j)$, we further
approximate that

\[ G(n,j) = \begin{cases} 
\frac{N_s^2}{2\alpha + 1} R_{nn}(j\Delta \tau) + N_s \left( 1 - \frac{N_s}{2\alpha + 1} \right) \frac{1}{M} \sum_{m=1}^{M} R_{mm}(j\Delta \tau) \right) & \text{if } \alpha \geq N_s, \\
\left( N_s - \frac{\alpha(\alpha + 1)}{2\alpha + 1} \right) R_{nn}(j\Delta \tau) + \frac{\alpha(\alpha + 1)}{2\alpha + 1} \frac{1}{M} \sum_{m=1}^{M} R_{mm}(j\Delta \tau) \right) & \text{if } \alpha < N_s.
\end{cases} \quad (4.18) \]

Let \( \zeta(\alpha, N_s) = \frac{N_s}{2\alpha + 1} \) for \( \alpha \geq N_s \) and \( \zeta(\alpha, N_s) = 1 - \frac{\alpha(\alpha + 1)}{N_s(2\alpha + 1)} \) for \( \alpha < N_s \), substitute (4.18) into (4.16), we have

\[
\var[MPI_n(q)] = \frac{E_1}{N_s} \mathbf{h}_1^H \mathbf{h}_1 \mathbf{I}_{t_c} \left[ N_s \zeta(\alpha, N_s) \sum_{j=1-N_u}^{N_s-1} R_{nn}(j\Delta \tau) + (1 - \zeta(\alpha, N_s)) \frac{N_s}{M} \sum_{m=1}^{M} \sum_{j=1-N_u}^{N_s-1} R_{mm}(j\Delta \tau) \right] \mathbf{h}_1 \mathbf{I}_{t_c}
\]

\[
= \frac{E_1}{N_f} \left( \zeta(\alpha, N_s) \sum_{j=1-N_u}^{N_s-1} R_{nn}(j\Delta \tau) + [1 - \zeta(\alpha, N_s)] \frac{1}{M} \sum_{m=1}^{M} \sum_{j=1-N_u}^{N_s-1} R_{mm}(j\Delta \tau) \right) \mathbf{h}_1^H \mathbf{h}_1 \sum_{l=0}^{L-1} \mathbf{I}_{t_c}^l
\]

(4.19)

where the approximation \( \mathbf{H}_{1,j,l_c} \mathbf{H}_{1,j,l_c}^H \approx \frac{1}{N_f} \mathbf{h}_1^H \mathbf{h}_1 \mathbf{I}_{t_c} \) is based on the characteristic that the polarities of multipath components are independent and equally likely to be positive or negative [30], thus the off-diagonal elements of \( \mathbf{H}_{1,j,l_c} \mathbf{H}_{1,j,l_c}^H \) can be neglected for sufficiently large value of \( \alpha \).

For \( MUL_n(q) \), we find that \( \mathbb{E}[MUL_n(q)] = 0 \) and its variance is given by

\[
\var[MUL_n(q)] = \mathbf{h}_1^H \sum_{u=2}^{N_u} \sum_{j=1-N_u}^{N_u-1} \mathbb{E} \left[ \sum_{i=q_{N_f}}^{q_{N_f}+N_u-1} \int_0^{T_u} f_{u,j,l_c}(i,t-j\Delta \tau) \omega_n(t) dt \int_0^{T_u} f_{u,j,l_c}^T(i,t-j\Delta \tau) \omega_n(t) dt \right] \mathbf{h}_1 \mathbf{I}_{t_c}
\]

\[
= \mathbf{h}_1^H \sum_{u=2}^{N_u} \sum_{j=1-N_u}^{N_u-1} \left[ \frac{N_s}{M} \sum_{m=1}^{M} R_{mm}(j\Delta \tau) \left( \frac{1}{N_f} \mathbf{h}_1^H \mathbf{h}_1 \right) \mathbf{I}_{t_c} \right] \mathbf{h}_1 \mathbf{I}_{t_c}
\]

\[
= \frac{1}{N_f M} \sum_{u=2}^{N_u} \sum_{j=1-N_u}^{N_u-1} R_{nn}(j\Delta \tau) \sum_{u=2}^{N_u} \mathbf{h}_1^H \mathbf{h}_1 \sum_{l=0}^{L-1} \mathbf{I}_{t_c}^l
\]

(4.20)

where \( \mathbb{E} \left[ \sum_{i=q_{N_f}}^{q_{N_f}+N_u-1} \int_0^{T_u} f_{u,j,l_c}(i,t-j\Delta \tau) \omega_n(t) dt \int_0^{T_u} f_{u,j,l_c}^T(i,t-j\Delta \tau) \omega_n(t) dt \right] \) is approximated by a \( L_c \times L_c \) diagonal matrix \( \frac{N_s}{M} \sum_{m=1}^{M} R_{mm}(j\Delta \tau) \left( \frac{1}{N_f} \mathbf{h}_1^H \mathbf{h}_1 \right) \mathbf{I}_{t_c} \) according to the definition of \( f_{u,j,l_c}(i,t) \) in (A.4) and because \( \mathbb{E}[d_{u,k}^T] = \sum_{m=1}^{M} m \) for \( u \in \{2, \ldots, N_u\} \) and any \( k \).
Since $\text{MPI}(q)$, $\text{MUI}(q)$ and $\eta(q)$ are zero-mean and independent, $y_n(q)$ is Gaussian with $E[y_n(q)] = S_n(q)$ and $\text{var}[y_n(q)] = \text{var}[\text{MPI}(q)] + \text{var}[\text{MUI}(q)] + \text{var}[\eta(q)]$. Therefore,

$$P(y_n(q) < 0 \mid h) = \frac{1}{\sqrt{2\pi \text{var}[y_n(q)]}} \int_{-\infty}^{0} \exp \left( - \frac{(\lambda - S_n(q))^2}{2 \text{var}[y_n(q)]} \right) d\lambda = Q\left( \frac{S_n(q)}{\sqrt{\text{var}[y_n(q)]}} \right) \quad (4.21)$$

Similarly, we find that $y_x(q)$ ($x \in \{1, \ldots, M\}$ and $x \neq n$) is a zero-mean Gaussian random variable with $\text{var}[y_x(q)] = \text{var}[\text{MPI}(q)] + \text{var}[\text{MUI}(q)] + \text{var}[\eta(q)]$, where $\text{var}[\text{MPI}(q)]$, $\text{var}[\text{MUI}(q)]$ and $\text{var}[\eta(q)]$ have the same expressions as $\text{var}[\text{MPI}(q)]$, $\text{var}[\text{MUI}(q)]$ and $\text{var}[\eta(q)]$, respectively, but with $R_{mn}(\tau)$ replaced by $R_{mx}(\tau)$ ($m \in \{1, \ldots, M\}$). Letting $Z_{x\tau}(q) = \text{MPI}(q) + \text{MUI}(q) + \eta(q) - \text{MPI}(q) - \text{MUI}(q) - \eta(q)$, we can show that

$$2 \sum_{x=1, x \neq n}^{M} P(y_x(q) > y_n(q) \mid y_n(q) > 0 \mid h) = 2 \sum_{x=1, x \neq n}^{M} P(Z_{x\tau}(q) > S_n(q) \mid h) \quad (4.22)$$

where $Z_{x\tau}(q)$ is also a zero-mean Gaussian random variable with variance $\text{var}[Z_{x\tau}(q)] = \text{var}[\text{MPI}(q) - \text{MPI}(q)] + \text{var}[\text{MUI}(q) - \text{MUI}(q)] + \text{var}[\eta(q) - \eta(q)]$, because $\text{MPI}(q) - \text{MPI}(q), \text{MUI}(q) - \text{MUI}(q)$ and $\eta(q) - \eta(q)$ are zero-mean and independent to each other. Following similar steps as in deriving $P(y_n(q) < 0 \mid h)$, we obtain that

$$2 \sum_{x=1, x \neq n}^{M} P(Z_{x\tau}(q) > S_n(q) \mid h) = 2 \sum_{x=1, x \neq n}^{M} \frac{1}{\sqrt{2\pi \text{var}[Z_{x\tau}(q)]}} \int_{S_n(q)}^{\infty} \exp \left( - \frac{\lambda^2}{2 \text{var}[Z_{x\tau}(q)]} \right) d\lambda$$

$$= 2 \sum_{x=1, x \neq n}^{M} Q\left( \frac{S_n^2(q)}{\sqrt{\text{var}[Z_{x\tau}(q)]}} \right) \quad (4.23)$$

in which for $\text{var}[Z_{x\tau}(q)]$, it is easy to show that

$$\text{var}[\text{MPI}(q) - \text{MPI}(q)] \
= \frac{E_1}{N_f} \sum_{j=1}^{N_u} \left[ R_{mx}(j\Delta \tau) - R_{mn}(j\Delta \tau) \right]^2 + \frac{1 - \zeta(\alpha, N_s)}{M} \sum_{m=1}^{N_u} \sum_{j=1}^{N_u} \left[ R_{mx}(j\Delta \tau) - R_{mn}(j\Delta \tau) \right]^2$$

$$\approx h_{ii} h_{ii} \sum_{j=0}^{L-1} h_{ij}^2 \quad (4.24)$$

$$\text{var}[\text{MUI}(q) - \text{MUI}(q)] = \frac{1}{N_f M} \sum_{m=1}^{N_u} \sum_{j=1}^{N_u} \left[ R_{mx}(j\Delta \tau) - R_{mn}(j\Delta \tau) \right]^2 \sum_{u=2}^{N_u} h_{uu} h_{uu} \sum_{j=0}^{L-1} h_{ij}^2 \quad (4.25)$$
\[ \text{var}[\eta_s(q) - \eta_n(q)] = N_0 N_s \sum_{i=0}^{L-1} h_{i,j}^2 \] (4.26)

For biorthogonal signals, BER is a complicated function of SER, we approximate it as follows [59]

\[
P_b(\varepsilon|+n,h) \leq \frac{2M-1}{2(2M-2)} \sum_{x=1, x \neq n}^{M} P(y_x(q) > y_n(q), y_n(q) > 0, h) + \frac{1}{2} P(y_n(q) < 0, h)
= \frac{2M-1}{2M-2} \sum_{x=1, x \neq n}^{M} \left( Q\left( \frac{S_n^2(q)}{\text{var}[Z_{x,n}(q)]} \right) + \frac{1}{2} Q\left( \frac{S_n^2(q)}{\text{var}[y_n(q)]} \right) \right) \] (4.27)

With the 2M-ary symbols being equally likely, the overall conditional BER is obtained by

\[
P_b(\varepsilon|h) = \frac{1}{2M} \sum_{m=1}^{M} [P_b(\varepsilon|+n,h) + P_b(\varepsilon|-n,h)]
\] (4.28)

where \( P_b(\varepsilon|-n,h) = P_b(\varepsilon|+n,h) \), for \( P(y_x(q)<y_n(q) \mid y_n(q)<0, h) = P(y_x(q)>y_n(q) \mid y_n(q)>0, h) \) and \( P(y_n(q)>0 \mid h) = P(y_n(q)<0 \mid h) \), based on the Gaussian assumptions of \( y_n(q) \) and \( y_x(q) \).

The above results apply to cases of \( M \geq 2 \). If only one pulse \( w_1(t) \) is used in the modulation, i.e. \( M = 1 \), the BOPK scheme reduces to bipolar PAM [17], for which the BER is given by

\[
P_b(\varepsilon|h) = \frac{1}{2} \left[ P_b(\varepsilon|+1,h) + P_b(\varepsilon|-1,h) \right] = Q\left( \frac{S_n^2(q)}{\text{var}[y_1(q)]} \right)
\] (4.29)

### 4.3 Numerical and Simulation Results

#### 4.3.1 PSD of DS-MA BOPK Signals

In order to make the PSD of DS-MA BOPK signals comply with the FCC UWB emission mask [49], we investigate the PSD of the signal \( s_u(t) \) in (4.1). The PSD of \( s_u(t) \) is computed in Appendix B and is given by

\[
P_s(f) = \frac{E_u}{N_s T_f} \frac{1}{M} \sum_{m=1}^{M} |W_m(f)|^2
\] (4.30)

where \( E_u (N_s T_f) \) is the power of the \( u \)th user, and \( W_m(f) \) is the Fourier transform (FT) of the pulse \( w_m(t) \) (\( m = 1, \ldots, M \)). We can see that for DS-MA BOPK signals, the PSD
depends only on the PSDs of the pulses. Therefore, if the PSD of each pulse complies with the FCC UWB emission mask [49], the PSD of the DS-MA BOPK signal can comply with the FCC UWB regulation.

To coincide with the FCC UWB indoor emission mask [49], we use the orthogonal Gaussian pulses and modified Hermite pulses [12] discussed in Section 3.1.2 to model the received pulses in our simulations. In this context, 4-ary BOPK and bipolar PAM will employ the two Gaussian pulses $p_4(t)$ and $p_5(t)$ as shown in Figure 3.1. It can be observed that the durations of $p_4(t)$ and $p_5(t)$ are around 0.5ns, while the $-10\text{dB}$-bandwidth of them is about 7GHz (corresponding to a multipath delay resolution of 0.143ns). 8-ary BOPK will use the four modified Hermite pulses $\{v_1(t), \ldots, v_4(t)\}$ as shown in Figure 3.2, from which we can see that both the pulse duration and bandwidth increase with the pulse order, but the maximum $-10\text{dB}$-bandwidth occupied by them is also approximately 7GHz.

For all these pulses, our simulations will use the channel model recommended by IEEE 802.15.3a, which is based on 6GHz-bandwidth measurements and has a multipath resolution of 0.167ns [30]. Channel model parameters are chosen corresponding to a NLOS channel CM3 (with mean excess delay $\tau_M = 14.18$ns, RMS delay $\tau_{RMS} = 14.28$ns, and number of paths capturing 85% of the channel energy $NP(85\%) = 61.54$) given in Table 2 of [30]. For any channel realization, the channel coefficients are normalized so that the total energy of the resolved paths is unity, so as to remove the path loss and the loss in partial-Rake combining. Perfect timing sync and strict power control are assumed. The elements of signature sequences $c_{u,k} \in \{\pm 1\}$ ($u = 1, \ldots, N$, and any $k$) were generated as i.i.d. random variables. The asynchronous delays $\tau_u$ ($u = 1, \ldots, N$) were also i.i.d. random variables with $\tau_u (\text{mod} \ T_f)$ uniformly distributed on $[0, T_f]$. The BER is averaged over 1000 channel realizations.

### 4.3.2 Impact of Pulse Overlap

Figure 4.1 shows BER versus $E_b/N_0$ for 4-ary BOPK using $\{p_4(t), p_5(t)\}$, 8-ary BOPK using $\{v_1(t), v_2(t), v_3(t), v_4(t)\}$ and bipolar PAM using $p_4(t)$ or $p_5(t)$, for a single-user system with $R_b = 500\text{Mbps}$ and $N_s = 1$. Figure 4.2 shows BER performance of these
modulations for a single user with $R_b = 50$Mbps and $N_x = 10$. The pulse repetition time $T_f$ is determined accordingly by $R_b = (\log_22M)/(N_xT_f)$. In both Figure 4.1 and Figure 4.2, the corresponding BER curves obtained by assuming that the multipath delay resolution $\Delta\tau$ equals the pulse duration $T_w$ (dashed lines) and those calculated using our BER formulas (dotted lines) are also plotted for comparison. We observe that the performance of each modulation under the realistic channel model [30] is always degraded compared with its performance under the assumption of $\Delta\tau = T_w$ (i.e., no pulse overlap). This reveals that the fine multipath delay resolution provided by the wide bandwidth of UWB signals causes pulse overlap at the receiver and this degrades the system performance. Our BER formulas show their accuracy and importance in the performance evaluation of UWB systems operating in multipath environments. Comparing Figure 4.1 with Figure 4.2, for each modulation, the performance gap between the results obtained with and without the assumption of no pulse overlap increases with the increase of data rate, i.e., the impact of pulse overlap on performance is more significant at higher data rates. This is mainly due to the use of larger values of the processing gain $N_x$ ($T_f$ is fixed for each modulation) at lower data rates [13].

![Figure 4.1 BER vs. $E_b/N_0$ of Bipolar PAM that uses $p_d(t)$ or $p_5(t)$, 4-ary BOPK that uses $\{p_d(t), p_5(t)\}$, and 8-ary BOPK that uses $\{v_1(t), \ldots, v_4(t)\}$, for a single user with $R_b = 500$Mbps and $N_x = 1$. The corresponding formula BER curves (dotted lines) are plotted for comparison.](image)

Figure 4.1 BER vs. $E_b/N_0$ of Bipolar PAM that uses $p_d(t)$ or $p_5(t)$, 4-ary BOPK that uses $\{p_d(t), p_5(t)\}$, and 8-ary BOPK that uses $\{v_1(t), \ldots, v_4(t)\}$, for a single user with $R_b = 500$Mbps and $N_x = 1$. The corresponding formula BER curves (dotted lines) are plotted for comparison.
Figure 4.2 BER vs. $E_b/N_0$ of Bipolar PAM that uses $p_4(t)$ or $p_5(t)$, 4-ary BOPK that uses $\{p_4(t), p_5(t)\}$, and 8-ary BOPK that uses $\{v_1(t), \ldots, v_4(t)\}$, for a single user with $R_s = 50\text{Mbps}$ and $N_s = 10$. The corresponding formula BER curves (dotted lines) are plotted for comparison.

Figure 4.3 BER vs. $E_b/N_0$ of Bipolar PAM that uses $p_4(t)$ or $p_5(t)$, 4-ary BOPK that uses $\{p_4(t), p_5(t)\}$, and 8-ary BOPK that uses $\{v_1(t), \ldots, v_4(t)\}$, for a single user with $T_f = 2\text{ns}$ and $N_s = 1$.  

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Figure 4.3 shows BER versus $E_b/N_0$ of 4-ary BOPK, 8-ary BOPK and bipolar PAM for a single user with $T_f = 2$ns and $N_t = 1$. We can see that, with the assumption of no pulse overlap, 8-ary BOPK offers the best performance and achieves the highest data rate (1.5Gbps) at the same time. However, since the multipath channel causes the four Hermite pulses to overlap at the receiver, the actual performance of 8-ary BOPK degrades catastrophically. For modulations using the two Gaussian pulses, the performance degradation caused by pulse overlap is much less significant. This can be explained by comparing the sampled correlations of \{v_1(t), v_2(t), v_3(t), v_4(t)\} in Figure 3.4 with those of \{p_4(t), p_5(t)\} in Figure 3.3, where correlations values are sampled at integer multiples of the delay resolution (0.167ns). The comparison reveals that orthogonal Hermite pulses are non-zero for longer durations than Gaussian pulses and their sampled correlations do not appear as desirable as Gaussian pulses. In addition, we can observe in Figure 4.1 to Figure 4.3 that bipolar PAM using $p_5(t)$ performs better than bipolar PAM using $p_4(t)$. This is because the sampled autocorrelation of $p_5(t)$ has a narrower peak than that of $p_4(t)$ as shown in Figure 3.3. Hence, for modulations requiring only one pulse shape, e.g. bipolar PAM, $p_5(t)$ is preferable to $p_4(t)$.

As we have shown in Section 4.2.3, the longer the pulse duration $T_w$, the more possible correlations of the template pulse with overlapping pulses, which can degrade the performance. Therefore, to fully exploit the power efficiency promised by higher-level BOPK, as indicated by the dashed curve of 8-ary BOPK in Figure 4.3, the choice of signaling pulses is very important. Specifically, we would prefer $T_w B = 1$ where $B$ is the bandwidth of the UWB pulse, and good correlations characteristics represented by

$$\sum_{j=1}^{[T_w B]^{-1}} R_{nn}^2 \left( \frac{j}{B} \right) \to 1 \quad \text{and} \quad \sum_{m=1}^{M} \sum_{j=1}^{[T_w B]^{-1}} R_{mn}^2 \left( \frac{j}{B} \right) \to 1.$$  

Figure 4.4 and Figure 4.5 show the BER versus $E_b/N_0$ in the presence of $N_u = 10$ and $N_u = 20$ users, respectively, for each user $R_b = 50$Mbps and $N_t = 10$. By comparing the gap between corresponding solid curves and dashed curves in Figure 4.2, Figure 4.4 and Figure 4.5, we observe that the impact of channel-induced pulse overlap becomes more significant when the number of users increases. This is because more users will cause more interfering pulses overlapping at the receiver and thus corrupt the output of the pulse correlators. Our analytical formula again shows accuracy and importance in
performance evaluation for multiple-access impulse radio systems.

![Graph showing BER vs. E_b/N_0](image)

**Figure 4.4** BER vs. E_b/N_0 of Bipolar PAM that uses p_4(t) or p_5(t), 4-ary BOPK that uses {p_4(t), p_5(t)}, and 8-ary BOPK that uses {v_1(t), ..., v_4(t)}, for a system consisting of N_u = 10 users each with R_b = 50Mbps and N_i = 10.

### 4.3.3 Performance of Biorthogonal Pulse Keying (BOPK)

The results in Figure 4.1 and Figure 4.2 also show that 8-ary BOPK provides the best performance for both data rates considered, with the advantage in power-efficiency of BOPK over bipolar PAM more evident for the higher data rate. One reason for this is the inherent orthogonal signaling property of BOPK, i.e. the BER can be reduced with the use of more orthogonal waveforms [32]. Another reason is that since higher-level modulation carries more bits per symbol, longer pulse repetition time T_r can be used, leading to less interference caused by overlaps between channel responses to consecutively transmitted pulses.

According to the simulation results in Figure 4.2 and Figure 4.4, Table 4.1 gives the required E_b/N_0 (dB) for a target BER of 10^{-3}. For both cases of single-user and multi-user, 8-ary BOPK achieves the best power efficiency among all the listed modulations.
Comparing the $E_b/N_0$ values of the $N_u = 1$ case with those of the $N_u = 10$ case, we can see that 8-ary BOPK suffers from less performance degradation caused by multi-user interference, compared with 4-ary BOPK and bipolar PAM. This is due to the fact that for BOPK the interference caused by pulse shapes that are orthogonal to the desired pulse can be reduced by the pulse correlation process at the receiver, and this advantage becomes more evident when more orthogonal pulses are used, as shown by our analysis in Section 4.2.3. Another reason is that the longer pulse repetition time $T_f$ used by higher-level BOPK can also reduce pulse collisions from other users.

![Figure 4.5 BER vs. $E_b/N_0$ of Bipolar PAM that uses $p_d(t)$ or $p_3(t)$, 4-ary BOPK that uses $\{p_d(t), p_3(t)\}$, and 8-ary BOPK that uses $\{v_1(t), ..., v_d(t)\}$, for a system consisting of $N_u = 20$ users each with $R_u = 50$Mbps and $N_s = 10$.](image)

### 4.4 Conclusions

We have provided a detailed performance analysis for a DS-MA impulse radio system employing BOPK for data modulation. Specifically, our analysis investigates the impact of pulse overlap that results from the fine multipath delay resolution provided by the ultra-wide bandwidth of impulse radio signals. Simulations have been performed in the
channel model recommended by IEEE 802.15.3a, using pulses complying with the current FCC regulations. The differences between the actually simulated BER curves and the ones obtained by assuming no pulse overlap show that, channel-induced pulse overlap can degrade the performance significantly and the results of this chapter are important for system design and computation of performance for DS-MA impulse radio systems. Furthermore, $2M$-ary BOPK shows its advantage in providing power-efficient modulation for power-limited UWB communications, but the level of BOPK modulation is restricted by practical implementation considerations as the receiver complexity increases when more orthogonal pulses are employed.

Table 4.1 $E_s/N_0$ Required for BER $= 10^{-3}$ and $R_b = 50$Mbps ($N_s = 10$), Obtained through Simulations. (The results assumed the pulses in Figure 3.1 and Figure 3.2 as the received pulses.)

<table>
<thead>
<tr>
<th>Modulation Schemes</th>
<th>$E_s/N_0$ (dB)</th>
<th>$T_f$(ns)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$N_s = 1$</td>
<td>$N_s = 10$</td>
</tr>
<tr>
<td>Bipolar PAM (using $p_a(t)$)</td>
<td>7.24</td>
<td>16.2</td>
</tr>
<tr>
<td>Bipolar PAM (using $p_b(t)$)</td>
<td>7.17</td>
<td>14.5</td>
</tr>
<tr>
<td>4-ary BOPK (using $p_a(t)$ and $p_b(t)$)</td>
<td>7.3</td>
<td>9.9</td>
</tr>
<tr>
<td>8-ary BOPK (using $v_1(t)$, ... , $v_4(t)$)</td>
<td>6.3</td>
<td>8.4</td>
</tr>
</tbody>
</table>

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Chapter 5

Pilot Channel Assisted Generalized Selection Combining with Log-Likelihood Ratio Threshold Test for UWB Rake Receivers

In this chapter, we propose a pilot-channel assisted generalized selection combining scheme with log-likelihood ratio threshold test per path (PCA-$|\text{LLR}|$-T-GSC), which is tailored for UWB Rake receivers having limited number (e.g. $L_c$) of fingers. Our work is different in that, we release the path selection from the necessity of estimating all resolvable multipath components by setting an adaptive threshold of the magnitude of LLR per path, which takes into account the noise level, channel estimation errors, and the number of available Rake fingers. The use of LLR in path selection and data detection is motivated by the fact that hard decision based on LLR is optimal in the sense of minimizing BER [41]. Channel estimation is obtained by using a pilot-channel assisted modulation (PCAM), where the pilot channel is constructed with a pulse shape that is orthogonal to the pulse shape for data conveying and thus, can be separated from the data channel at the receiver using two corresponding pulse correlators [11]. Each pilot pulse is transmitted simultaneously with a data-bearing pulse. With the use of a tracking filter on the pilot channel and parameters optimized through jointly minimizing the estimation mean square error (MSE) and maximizing the average receiver output SNR, PCAM provides reliable estimates of the time-variant channel without sacrificing the overall throughput. Since path gains are estimated separately [16] and paths with smaller delays are generally stronger [39], the path search for PCA-$|\text{LLR}|$-T-GSC starts with the path arriving with the smallest delay and proceeds to later arriving paths one by one. A path is chosen only if its estimated magnitude of LLR exceeds the threshold. Once $L_c$ paths have been selected, the path search as well as the channel estimation process ends. Thus, PCA-$|\text{LLR}|$-T-GSC only needs to estimate a first portion of the channel impulse response. We employ direct-sequence (DS) bipolar pulse amplitude modulation (PAM) [64] for analytical convenience, while the extension to other modulations is possible. The performance of PCA-$|\text{LLR}|$-T-GSC is evaluated through
various simulations, where the IEEE 802.15.3a channel model [30], [55] and FCC-compliant orthogonal pulses [64] are used.

The remainder of the chapter is organized as follows. A time-variant model for the UWB multipath channel is introduced in Section 5.1. In Section 5.2, we propose the PCA-|LLR|-T-GSC scheme. Simulation results and conclusions are provided in Section 5.3 and Section 5.4, respectively.

5.1 UWB Channel with Time Variance

The propagation of UWB signals is characterized by dense multipath [30]. The multipath channel model used in this chapter is the one recommended by the IEEE 802.15.3a channel modeling subcommittee [30], as introduced in Section 2.2.5 and Section 3.2.1. This channel model assumes that the channel stays either completely static, or changes completely from one data burst to the next [30]. However, the channel is in general continuously time-varying due to the Doppler effects arising from relative motion between the transmitter and the receiver or the movement of scatterers in the environment [55]. Also, the temporal characteristics of the channel are important when the channel estimation relies on the information provided by previously received signals.

It was shown in [65] that the first-order auto-regressive (AR) model provides a sufficiently accurate model for time-selective fading channels. Therefore, we re-express the channel impulse response given by (3.6) using a time-variant tapped delay line as follows

\[ h(t, \tau) = \sum_{l=0}^{N_t-1} h_l(t) \delta(\tau - l\Delta\tau) \]  \hspace{1cm} (5.1)

where \( t \) and \( \tau \) denote the time and delay variables, respectively, and the time-varying dynamic of each particular \( l \)th path, \( h_l(t) \), is characterized by the first-order AR model as follows

\[ h_l(iT) = \alpha h_l((i-1)T) + \mu_l(i) \]  \hspace{1cm} (5.2)

in which \( \mu_l(t) \) is a zero-mean Gaussian driving noise statistically independent of \( h_l(t) \), with \( \text{var}[\mu_l(t)] = (1-\alpha^2)E[|h_l(t)|^2] \), and for a sampling rate of \( 1/T \), the fading correlation coefficient \( \alpha \) is given by
\[ \alpha = \frac{\mathbb{E}[h_i(t)h_j[(i-1)T]]}{\mathbb{E}[h_i^2(t)]} = J_0(2\pi f_d T) \]  

(5.3)

with \( J_0(\cdot) \) denoting the 0th-order Bessel function of the first kind and \( f_d \) being the maximum Doppler frequency. \( \alpha \) defines the scale of the channel time-variations and may vary between one (for a static channel) and zero (for a fast fading channel that changes independently from one sampling instant to the next). Since \( \alpha \) can be estimated as in [66], we assume henceforth that \( \alpha \) is known.

5.2 PCA - \(|\text{LLR}| - T\)-GSC

In this section, we propose the generalized selection combining with log-likelihood ratio threshold test per path (\(|\text{LLR}| - T\)-GSC) for UWB Rake receivers with limited number of fingers, with channel estimation provided by a pilot-channel assisted modulation (PCAM).

5.2.1 Pilot Channel Assisted Modulation & Receiver Structure

Proper path selection relies on accurate channel estimation [38]. We propose a pilot-channel assisted modulation (PCAM) for channel estimation in pulse-based UWB systems. Motivated by the fact that sending single pilots frequently on the pilot channel is better for channel-state tracking performance than sending larger pilot clusters with low frequency [67], we design the pilot channel to be parallel with the data channel, i.e. the pilot channel is constructed using a pulse shape that is orthogonal to the pulse shape for data conveying and each pilot pulse is transmitted simultaneously with a data-bearing pulse. Thus, the pilot channel experiences exactly the same fading as the data channel. The generation of orthogonal UWB pulses has been discussed in Section 3.1.2. With respect to the energy allocation among the transmitted pulses, we make all pilot pulses equi-energy and all data-bearing pulses equi-energy, respectively, because given the number of pilot (data-bearing) pulses and the total energy assigned to them, equi-energy pilot pulses minimize the channel estimation MSE and equi-energy data-bearing pulses maximize the average system capacity [46]. Since the pilot channel is transmitted
parallel with the data channel, the resulting instantaneous transmission power levels are even, despite the possible discrepancy between data-bearing pulse energy and pilot pulse energy.

Considering a UWB system employing bipolar pulse amplitude modulation (PAM) for data transmission, we define the waveforms for the data conveying channel and for the pilot channel, respectively, as follows

\[ s_{\text{data}}(t) = \sum_{i=-\infty}^{\infty} \sqrt{\frac{(1-\beta)E_b}{N_s}} c_i d_{\lfloor i/N_s \rfloor} w_{\text{dr}}(t - iT_f) \]  \hfill (5.4)

\[ s_{\text{pilot}}(t) = \sum_{i=-\infty}^{\infty} \sqrt{\frac{\beta E_b}{N_s}} c_i w_{\text{pr}}(t - iT_f) \]  \hfill (5.5)

where \( E_b \) is the energy per information bit, the energy allocation factor \( \beta \) (\( 0 < \beta < 1 \)) represents the percentage of total transmission energy assigned to pilot pulses, \( d_{\lfloor i/N_s \rfloor} \in \{\pm 1\} \) denotes the binary data stream, \( c_i \in \{\pm 1\} \) is the pseudorandom spreading sequence that is imposed on the data channel and the pilot channel simultaneously, \( w_{\text{dr}}(t) \) and \( w_{\text{pr}}(t) \) are two unit-energy orthogonal pulses both occupying on \( [0, T_w] \), these pulses are repeated \( N_s \) times per symbol period to provide a processing gain, \( T_f \geq T_w \) is the pulse-repetition time, and the bit rate is \( R_b = 1/(N_sT_f) \). Thus, the transmitted signal is given by \( s(t) = s_{\text{data}}(t) + s_{\text{pilot}}(t) \). With \( T_d \) denoting the maximum delay spread with respect to the first arriving path, we choose the frame time \( T_f > T_d + T_w \) so that channel responses to consecutively transmitted pulses will not overlap.

With the above definitions and the channel model defined in Section 5.1, the composite signal at the output of the receiver antenna is given by

\[ r(t) = \sqrt{\frac{E_b}{N_s}} \sum_{i=0}^{N_s-1} h_i(t) \sum_{i=-\infty}^{\infty} c_i \sqrt{\frac{1-\beta}{N_s}} d_{\lfloor i/N_s \rfloor} w_{\text{dr}}(t - iT_f - l\Delta \tau) + \sqrt{\beta} w_{\text{pr}}(t - iT_f - l\Delta \tau) \]  \hfill (5.6)

where \( \eta(t) \) denotes the AWGN with a double-sided power spectral density (PSD) of \( N_0/2 \), and the transmitted pulses \( w_{\text{dr}}(t) \) and \( w_{\text{pr}}(t) \) in (5.4) and (5.5) have been changed into \( w_{\text{dr}}(t) \) and \( w_{\text{pr}}(t) \), respectively, due to the wideband distorting effects of the channel and also the responses of the transmitter and receiver antennas. Suggested models for the distortion include differentiation [10], but other effects will also be present [78] making
the exact distortion difficult to characterize. Here we will assume that the orthogonality between the received pulse shapes can be retained by appropriately pre-distorting the transmitted signals [78]. In practice, however, it is likely that the pre-distortion will only approximately produce the desired received pulses and there will likely be some loss in performance. Alternatively, the receiver may be able to use training information to better determine the effects of the channel and the antennas on the received pulse shapes [46].

The receiver performs both channel estimation and data detection based on the received signal in (5.6). We consider a Rake receiver implemented by correlating the received signal with delayed versions of the template pulse waveform [32]. Specifically, the received signal \( r(t) \) is passed through a tapped-delay-line consisting of \( L_c \) delay taps \( \{l_1 \Delta \tau, l_2 \Delta \tau, \ldots, l_{L_c} \Delta \tau \} \), assuming the indices of the selected paths are given by \( \{l_1, l_2, \ldots, l_{L_c} \} \). At each of the \( L_c \) delays, \( r(t) \) is correlated with two locally generated templates \( w_{dl}(t) \) and \( w_{pr}(t) \), for the data-conveying channel and the pilot channel, respectively, and the outputs are sampled and combined. For purposes of analysis, we assume that the shapes of the received pulses are known at the receiver [42] and coarse timing has been acquired as in [68]. Since each pair of \( w_{dl}(t) \) and \( w_{pr}(t) \) in (5.6) are synchronized and \( \int_0^\infty w_{pr}(t) w_{dr}(t) dt = 0 \), the outputs of the two pulse-correlators corresponding to the \( l \)th path and the \( i \)th transmitted frame are given by

\[
x_l(i) = \int_0^\infty r(t + iT_f + i \Delta \tau) C_i w_{dr}(t) dt = \sqrt{\frac{(1 - \beta)E_b}{N_s}} h_l(iT_f) d_{l[i/N_s]} + \eta_{dl}(i) \tag{5.7}
\]

\[
y_l(i) = \int_0^\infty r(t + iT_f + i \Delta \tau) C_i w_{pr}(t) dt = \sqrt{\frac{\beta E_b}{N_s}} h_l(iT_f) + \eta_{pl}(i) \tag{5.8}
\]

where \( \eta_{dl}(i) = \int_0^\infty \eta(t + iT_f + i \Delta \tau) C_i w_{dr}(t) dt \) and \( \eta_{pl}(i) = \int_0^\infty \eta(t + iT_f + i \Delta \tau) C_i w_{pr}(t) dt \) are the sampled noises induced by the AWGN. An important consideration is the cross-correlations that may cause loss of orthogonality between the received pulses \( w_{dl}(t) \) and \( w_{pr}(t) \) when they are time dispersed by the multipath channel. We will take into account this in the numerical simulations.
5.2.2 Data Detection & Channel Estimation

The correlator outputs of the data-bearing channel, i.e., $x_i(i)$ in (5.7), are for use in data detection. Since hard decision based on the log-likelihood ratio (LLR) is optimal in the sense of minimizing BER [41], we propose to select paths for combining using a threshold of the magnitude of LLR per path ($|LLR|\cdot T$-GSC) and makes data-detection decisions based on the LLR provided by the selected paths. For a Rake receiver with $L_c$ fingers, the input to the detector is the LLR of the $L_c$ selected paths over a symbol period. Assuming the indices of the selected paths are given by $\{l_1, l_2, \ldots, l_{L_c}\}$, and independent fading and uncorrelated white Gaussian noise between paths, the decision statistic for a particular $q$th transmitted symbol ($d_q$) is given by

$$\Lambda = \sum_{i=qN_s}^{qN_s+N_s-1} \ln \frac{\Pr\left[ d_{[i/N_s]} = +1 | h_{l_1}(iT_f), x_{l_1}(i), \ldots, h_{l_{i_L}}(iT_f), x_{l_{i_L}}(i) \right]}{\Pr\left[ d_{[i/N_s]} = -1 | h_{l_1}(iT_f), x_{l_1}(i), \ldots, h_{l_{i_L}}(iT_f), x_{l_{i_L}}(i) \right]}$$

$$= \sum_{i=qN_s}^{qN_s+N_s-1} \ln \frac{\prod_{k=1}^{L_c} \Pr\left[ d_{[i/N_s]} = +1 | h_{l_k}(iT_f), x_{l_k}(i) \right]}{\prod_{k=1}^{L_c} \Pr\left[ d_{[i/N_s]} = -1 | h_{l_k}(iT_f), x_{l_k}(i) \right]}$$

$$= \sum_{i=qN_s}^{qN_s+N_s-1} \ln \frac{\prod_{k=1}^{L_c} \Pr\left[ x_{l_k}(i) | h_{l_k}(iT_f), d_{[i/N_s]} = +1 \right]}{\prod_{k=1}^{L_c} \Pr\left[ x_{l_k}(i) | h_{l_k}(iT_f), d_{[i/N_s]} = -1 \right]}$$

$$= \sum_{i=qN_s}^{qN_s+N_s-1} \ln \left[ \frac{\exp \left[ -\frac{\sqrt{1-\beta}E_b}{N_s h_{l_k}(iT_f)} \right] / N_0}{\exp \left[ -\frac{\sqrt{1-\beta}E_b}{N_s h_{l_k}(iT_f)} \right] / N_0} \right]$$

$$= \frac{4}{N_0 \sqrt{N_s}} \sum_{i=qN_s}^{qN_s+N_s-1} \ln \left( \frac{\exp \left[ -\sqrt{1-\beta}E_b / N_s h_{l_k}(iT_f) \right] / N_0}{\exp \left[ -\sqrt{1-\beta}E_b / N_s h_{l_k}(iT_f) \right] / N_0} \right)$$

(5.9)

where $\Pr\left[ d_{[i/N_s]} = \pm1 | h_{l_k}(iT_f), x_{l_k}(i), \ldots, h_{l_{i_L}}(iT_f), x_{l_{i_L}}(i) \right]$ is the likelihood function of the $[i/N_s]$th transmitted data symbol conditioned on the $L_c$ selected path gains and correlator outputs of the data-bearing channel for the $i$th transmitted frame. The sign of $\Lambda$ is the hard decision value, i.e. the detector decides $d_q = 1$ if $\Lambda > 0$ and decides $d_q = -1$ otherwise. It is obvious that the hard decision based on the combined LLR ($\Lambda$) requires the knowledge of path gains of the selected paths.
The correlator outputs of the pilot channel, i.e. \( y(i) \) in (5.8), are used for channel estimation. Since UWB signals are of high multipath diversities, the channel estimation is very difficult because the SNR per path is very low while the estimation for each path is performed separately [16]. In view of this, we employ a tracking filter [69] operating on the pilot channel at the receiver. Consequently, the estimate of the \( l \)th path for the \( i \)th transmitted frame is given by

\[
\hat{h}_l(iT_f) = (1 - \rho) \sqrt{\frac{N_s}{\beta E_b}} y(i) + \rho \hat{h}_l[(i-1)T_f]
\]  

(5.10)

in which the forgetting factor \( \rho \) is close to but less than one, and the current measurement \( y(i) \) is multiplied by a constant \( \sqrt{N_s/(\beta E_b)} \) to guarantee the unbiasedness of the estimator [46], where \( \beta, E_b \) and \( N_s \) are assumed to be available at the receiver as side information. Although this tracking filter is not of a high order, it still allows a reasonable amount of averaging to provide a desirable SNR per path for channel estimation [69].

With the estimates of path gains obtained from the pilot channel, i.e. \( \hat{h}_l(iT_f) \) given by (5.10), the LLR decision statistic that is actually used for a particular \( q \)th transmitted symbol is given by

\[
\Lambda = \frac{4}{N_0} \sqrt{\frac{(1-\beta)E_b}{N_s}} \sum_{i=qN_s}^{qN_s+N_s-1} \sum_{l=1}^{L} \hat{h}_l(iT_f) x_l(i)
\]  

(5.11)

in which \( \hat{h}_l(iT_f) \) (\( l \in \{l_1, l_2, \ldots, l_L\} \)) is given by (5.10). We can see that the LLR per path for bipolar PAM signals is proportional to the product of the estimated path gain and the corresponding correlator output of the data-bearing channel.

### 5.2.3 Optimization of PCAM Parameters

To optimize the tracking filter design, as given by (5.10), our task is to minimize the channel estimation mean square error (MSE). We define the channel estimation error as

\[
e_l(iT_f) = \hat{h}_l(iT_f) - h_l(iT_f)
\]  

(5.12)

for \( l = 0, \ldots, N_r-1 \), and assume that \( e_l(t) \) is statistically independent of \( h_l(t) \) [46].
Substituting (5.2), (5.8) and (5.10) into (5.12), we obtain that

\[ e_l(t_f) = \rho e_l[0] + \rho (1 - \alpha) h_l[0] (1 - \rho) \sqrt{\frac{N_s}{\beta E_b}} \eta_{p_l}(i) \]  

(5.13)

where \( \mu_l(t) \) is independent of \( h_l(t) \) and is zero-mean with \( \text{var}[\mu_l(t)] = (1 - \alpha^2) \text{E}[|h_l(t)|^2] \) [65], the sampled noise \( \eta_{p_l}(i) \) is approximately zero-mean white Gaussian [46]. According to (5.13) and the fading characteristics of the channel defined in [30] and [65], it is easy to show that \( \text{E}[e_l(t)] = 0 \) and the estimation MSE is given by

\[ \sigma_{e_l}^2(l, \beta, \rho) = \text{E}\left[ |e_l(t_f)|^2 \right] = 2(1 - \alpha) |h_l|^2 \frac{\rho^2}{1 - \rho^2} + \frac{N_s N_0}{2 \beta E_b} \frac{1 - \rho}{1 + \rho} \]  

(5.14)

in which \( |h_l|^2 = \text{E}[|h_l(t)|^2] \) for \( l \in \{0, ..., N_r - 1\} \).

Now, optimizing the tracking filter is equivalent to minimizing \( \sigma_{e}^2(l, \beta, \rho) \) as a function of \( \rho \). To find the \( \rho_{\text{opt}}(l, \beta) = \arg \min_{\rho} \sigma_{e}^2(l, \beta, \rho) \), we solve \( \frac{d \sigma_{e}^2(l, \beta, \rho)}{d \rho} = 0 \) as a function of \( \rho \) and obtain that

\[ \rho_{\text{opt}}(l, \beta) = 1 + (1 - \alpha) |h_l|^2 \frac{2 \beta E_b}{N_s N_0} - \sqrt{\frac{1 + (1 - \alpha) |h_l|^2}{N_s N_0} \frac{2 \beta E_b}{N_s N_0}} - 1 \]  

(5.15)

Substituting (5.15) into (5.14) and performing some algebraic manipulations, it turns out that

\[ \sigma_{e_{\text{opt}}}^2(l, \beta) = \left[ \frac{\beta E_b}{N_s N_0} + \sqrt{\frac{\beta E_b}{N_s N_0} \left( \frac{\beta E_b}{N_s N_0} + \frac{1}{(1 - \alpha) |h_l|^2} \right) \frac{2 \beta E_b}{N_s N_0} \right]^{-1} \]  

(5.16)

which implies that as \( \beta \) increases, the channel estimation MSE decreases monotonically, which facilitates the collection of multipath diversity. On the other hand, for a fixed bit energy \( E_b \), increasing \( \beta \) will decrease the energy assigned to the data-bearing pulses and thus will haul back the bit error rate (BER) performance. The optimal \( \beta \) indicating the optimal allocation of the transmission energy among information and pilot pulses is not obvious without taking into account the receiver detection performance. Since it is the SNR of the decision statistics that determines the data-detection performance [46], we now optimize \( \beta \) by maximizing the average receiver output SNR-per-bit subject to a fixed \( E_b \) and given receiver structure.
According to (5.7), (5.12) and (5.16), the average SNR of the LLR decision statistic in (5.11) is given by

\[
\overline{\gamma}_b(\beta) = \frac{2(1-\beta)\sum_{k=1}^{L_c}|h_k|^2 E_b}{N_0} \frac{2(1-\beta)\sum_{k=1}^{L_c}\sigma_{\text{opt}}^2(l_k, \beta) |h_k|^2}{N_0} + \frac{L_c}{\sum_{k=1}^{L_c}|h_k|^2} + 1
\]

(5.17)

where \(|h_l|^2 = E[|h_l|^2]\) for \(l \in \{l_1, l_2, ..., l_{L_c}\}\), and we can observe again that as \(\beta\) increases, although the estimation MSE decreases (according to (5.16)) and thus enhances the SNR, \(1-\beta\) decreases and reduces the effective SNR. Following the steps as fully established in Appendix C, we find that the \(\beta_{\text{opt}} = \arg \max_{\beta} \overline{\gamma}_b(\beta)\) is given by

\[
\beta_{\text{opt}} = \sqrt{\frac{(1-\alpha)N_sN_0}{4E_b} \frac{\sum_{k=1}^{L_c}|h_k|^2}{\sum_{k=1}^{L_c}|h_k|^3}} + \sqrt{\frac{(1-\alpha)E_b}{N_sN_0} \frac{\sum_{k=1}^{L_c}|h_k|^2}{3\sum_{k=1}^{L_c}|h_k|^3}} + 1
\]

(5.18)

Substituting (5.18) into (5.15), the optimal design of the tracking filter for a particular \(l^{th}\) \((l = 0, 1, ..., N_t-1)\) path is finally given by

\[
\rho_{\text{opt}}(l) = 1 + (1-\alpha)|h_l|^2 \frac{2\beta_{\text{opt}}E_b}{N_sN_0} - \sqrt{1 + (1-\alpha)|h_l|^2 \frac{2\beta_{\text{opt}}E_b}{N_sN_0}} - 1
\]

(5.19)

The above optimal solutions for the energy allocation factor \(\beta\) used in PCAM transmission and the forgetting factor \(\rho\) of the tracking filter require the knowledge of the channel fading statistics of the selected paths, which however, are usually unknown at the transmitter and/or the receiver. Also note that for a dynamic selective Rake receiver, the transmitter has no knowledge of which paths to be combined at the receiver. In view of this, we derive the following suboptimal designs for the PCAM and the tracking filter, respectively

\[
\beta_{\text{subopt}} = \sqrt{\frac{(1-\alpha)L_cN_sN_0}{4E_b}} + \sqrt{\frac{(1-\alpha)E_b}{N_sL_cN_0}} - \sqrt{\frac{(1-\alpha)L_cN_sN_0}{4E_b}} - \frac{2}{3} \sqrt{\frac{(1-\alpha)E_b}{N_sL_cN_0}}
\]

(5.20)
\[ \rho_{\text{subopt}} = 1 + (1 - \alpha) \frac{2 \beta_{\text{subopt}} E_b}{L_c N_s N_0} - \sqrt{\left[ 1 + (1 - \alpha) \frac{2 \beta_{\text{subopt}} E_b}{L_c N_s N_0} \right]^2 - 1} \]  

(5.21)

where we approximate \( |h_l|^2 = 1/L_c \) for \( l \in \{l_1, l_2, \ldots, l_{L_c}\} \) with the assumption that the total energy of the channel is unity and the \( L_c \) selected paths are strong enough to contain almost all the energy conveyed by the channel, i.e. \( \sum_{k=1}^{L_c} |h_{l_k}|^2 = \sum_{l=0}^{N_r-1} |h_l|^2 = 1 \). These suboptimal solutions are insensitive to the channel fading statistics and thus, are more suitable for practical implementations. For this reason, in the following and unless otherwise mentioned, the PCAM parameters \( \beta \) and \( \rho \) are generated according to (5.20) and (5.21), respectively.

### 5.2.4 LLR Threshold for Path Selection

Since the magnitude of \( \hat{A} \) in (5.11) represents the reliability of the LLR-based hard decision [40], paths with larger values of \( |\hat{h}_l(T_f)x_i(l)| \) \( (l \in \{0, \ldots, N_r-1\}) \) are more desirable to use in combining, under the assumption that the channel estimates provided by \( \hat{h}_l(T_f) \) are sufficiently accurate. In order to avoid sorting all the resolvable multipath components by their instantaneous \( |\hat{h}_l(T_f)x_i(l)| \) \( (l = 0, \ldots, N_r-1) \), which requires estimation of the complete channel, we propose to set a threshold of \( |\hat{h}_l(T_f)x_i(l)| \) for path selection. A path is chosen for combining only if the corresponding \( |\hat{h}_l(T_f)x_i(l)| \) \( (l \in \{0, \ldots, N_r-1\}) \) exceeds the threshold. Since path gains are estimated separately and paths with smaller delays are generally stronger [39], the path search starts with the path arriving with the smallest delay and proceeds to later arriving paths one by one. Once \( L_c \) paths have been selected, the path search process as well as the channel estimation process stops. Therefore, the need to estimate all resolvable multipath components is alleviated.

Now, our objective is to provide a proper threshold setting so as to make the best use of the \( L_c \) available Rake fingers. With \( \hat{h}_l(T_f) = h_l(T_f) + e_l(T_f) \) and according to (5.7), we calculate the expectation of the squared magnitude of \( \hat{h}_l(T_f)x_i(l) \) as follows
\begin{equation}
E\left[ |\hat{h}_l(iT_f)x_l(i)|^2 \right] = \frac{(1-\beta)E_b}{N_s} |h_l|^4 + \frac{N_0}{2} |h_l|^2 + \frac{(1-\beta)E_b}{N_s} \sigma_e^2(t) |h_l|^2 + \frac{N_0}{2} \sigma_e^2(t) \tag{5.22}
\end{equation}

where $|h_l|^2 = E[|h(t)|^2]$, $e_l(t)$ is zero-mean and statistically independent of $h_l(t)$, $\sigma_e^2(t) = E[|e_l(t)|^2]$. We observe that the power of $|\hat{h}_l(iT_f)x_l(i)|$ ($l = 0, ..., N_r-1$) depends not only on the path gain $h_l(iT_f)$ but also on the noise variance and the channel estimation error, all of which need to be taken into account for the threshold setting for $|\hat{h}_l(iT_f)x_l(i)|$.

Since the receiver does not know the noise variance $N_0/2$ a priori, it has to be estimated. According to (5.8), it is easy to show that $E[\eta_p(i)] = 0$ and $\text{var}[\eta_p(i)] = N_0/2$, therefore, we can estimate the noise variance based on the correlator outputs from the pilot channel, i.e. $y_l(i)$ in (5.8). For a time period of $NT_f$ that is shorter than the channel coherence time, we collect all the correlator outputs $\{y_l(1), ..., y_l(N)\}$ of the $l$th path and calculate their mean as $\bar{y}_l = \frac{1}{N} \sum_{i=1}^{N} y_l(i)$, then the noise variance can be estimated by $\hat{\sigma}_n^2 = \frac{1}{N} \sum_{i=1}^{N} (y_l(i) - \bar{y}_l)^2$, in which $\hat{\sigma}_n^2$ denotes the estimate of $N_0/2$. The accuracy of the estimation can be improved by repeating the above procedure on different paths to obtain an average of the estimated noise variance.

We then take into account the channel estimation errors by computing the expectation of the squared magnitudes of the channel gain estimates, according to $\hat{h}_l(iT_f) = h_l(iT_f) + e_l(iT_f)$, as follows
\begin{equation}
E\left[ |\hat{h}_l(iT_f)|^2 \right] = E\left[ |h_l(iT_f)|^2 \right] + E\left[ |e_l(iT_f)|^2 \right] + 2E[h_l(iT_f)e_l(iT_f)] = |h_l|^2 + \sigma_e^2(t) \tag{5.23}
\end{equation}
which shows that the estimated power of each path is biased by $\sigma_e^2(t)$ ($l = 0, ..., N_r-1$).

As a result, weak paths that contribute little energy to the combiner are susceptible to estimation errors, which can degrade the receiver detection performance [16]. Taking this into consideration, we would expect the selected paths are strong enough to conform to the following requirement
\begin{equation}
|h_l|^2 = K\sigma_e^2(t) \tag{5.24}
\end{equation}
where the scaling factor $K$ is larger than unity and will be described in detail below.

Substituting (5.16) into (5.24), it turns out that
\[ |h_i|^2 = \frac{K^2 (1-\alpha)N_s N_0}{(1+2K-2K\alpha)\beta E_b} \]  

(5.25)

Substituting (5.16) and (5.25) into (5.22) and taking the square root, we obtain the threshold used in path selection for LLR-based combining as follows

\[ |h_{\text{th}}| = \sqrt{\frac{[(1-\beta)(1-\alpha)2K^4 + (1-\alpha)2K^3 + (3-2\alpha)\beta K^2 + \beta K(1-\alpha)N_s N_0]^2}{[(1-\alpha)^2 4K^2 + (1-\alpha)4K + 1]2\beta^2 E_b}} \]  

(5.26)

where the noise variance and the channel estimation errors have been taken into account.

The value of the threshold must also adapt to the number of available Rake fingers. Given \( L_c \) Rake fingers, if the threshold \( |h_{\text{th}}| \) is too large, the number of paths with \( \hat{h}_i(T_f)x_r(i) \) (\( i \in \{0, \ldots, N_r-1\} \)) exceeding \( |h_{\text{th}}| \) might be less than \( L_c \) and some of the Rake fingers would be wasted; on the other hand, if the threshold \( |h_{\text{th}}| \) is too small, the number of selected paths might have already reached \( L_c \) even before the truly strong paths are examined, because the path search process starts with the first arriving path and proceeds to the later arriving paths one by one. For a given \( L_c, f_d \), and given parameters of PCAM, Figure 5.1 shows that the threshold in (5.26) increases with \( K \) monotonically. Consequently, the path selection process consists of two stages. In the first stage, the value of \( K \) is adjusted subject to the given \( L_c \) and other operational parameters through several iterations of the path search process starting with a certain value of \( K \) and a search grid \( \Delta \) for \( K \), hence the threshold could be set up once. If a \((L_c+1)^{\text{th}}\) path is found to exceed the current threshold \( |h_{\text{th}}| \), then the current value of \( K \) is increased by an amount of \( \Delta \) and the path search process starts over again from the first arriving path with the updated \( |h_{\text{th}}| \). If the number of selected paths at the end of a path search process is less than \( L_c \), then the current value of \( K \) is decreased by an amount of \( \Delta \) in the next iteration. Otherwise, the adjusting process stops and the current value of \( K \) is used to calculate \( |h_{\text{th}}| \) according to (5.26). This adjustment for the threshold ensures that only the paths providing the largest \( \hat{h}_i(T_f)x_r(i) \) (\( i \in \{0, \ldots, N_r-1\} \)) are selected. In the second stage, path selection is performed using the threshold \( |h_{\text{th}}| \) provided by the first stage.
Figure 5.1 Threshold vs. $K$ calculated according to (5.26) and (5.20), with $L_c = 10$, $f_d = 20$Hz, $T_f = 100$ns, and $N_r = 10$.

5.3 Simulation Results

In this section, we present simulation results to validate the proposed PCAM and PCA-LLR-T-GSC schemes. In the simulations, we use the channel model recommended by IEEE 802.15.3a [30], where parameters are chosen corresponding to a NLOS channel CM3 given in Table 2 of [30]. The channel time-variations are modeled following the time-domain filtering implementation suggested by [55]. We first create multiple realizations of the channel impulse response according to the selected CM3 [30]. The autocorrelation function (ACF) of each multipath delay is obtained by inverse Fourier-transforming its Doppler spectrum, which is computed from the small-scale averaged azimuthal power spectrum [55]. The samples for each separate delay are then filtered by the corresponding ACF. We normalize the total energy of each resulting channel impulse response to unity, to eliminate the path loss and shadowing effect [47]. In order to coincide with the FCC UWB indoor mask [49], we model the received pulse shapes using the two orthogonal Gaussian pulses $p_4(t)$ and $p_5(t)$ ($\tau_m = 0.182$ns) as shown in

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Figure 3.1, for both of which the pulse durations are \( T_w = 0.5\text{ns} \) and the occupied bands are 3.5-10.5GHz with a geometric center frequency [30] of \( f_c = 6\text{GHz} \). We assume the Rake receiver can resolve multipath components with delays differing by at least 0.167ns, i.e. the path resolution of the channel suggested by [30]. Possible cross-correlations between the received pulses when they are time-dispersed by the multipath channel are considered in the simulations. The BER is averaged over 1000 channel realizations.

We first compare the performance of our proposed PCAM with the optimal pilot waveform assisted modulation (PWAM) [46], because it has been shown in [46] that the optimal PWAM always outperforms TR schemes [45], [47] and the Equi-SNR PWAM [46] that features data-bearing and pilot pulses with equal SNR. For the optimal PWAM, we will consider both distributed PWAM that evenly distributes the pilot pulses throughout each transmission burst and preamble PWAM that gathers all pilot pulses at the beginning of each transmission burst. Figure 5.2 illustrates the BER vs. \( E_b/N_0 \) for the PCAM, distributed PWAM and preamble PWAM schemes. The BER curve with ideal channel estimate is also plotted for comparison. We assume the Rake receiver collects all resolvable multipath components unless otherwise mentioned. This allows better insights into the performance of the proposed PCAM and fair comparisons with the PWAM [46], because the correlation receiver in [46] employs an integration interval covering the whole channel delay spread. All results are obtained with the maximum Doppler frequency \( f_d = 20\text{Hz} \) corresponding to \( f_c = 6\text{GHz} \) and \( v = 1\text{m/sec} \) (i.e. the speed of relative motions between the transmitter and the receiver [55]), \( T_f = 200\text{ns} \), and \( N_r = 100 \). Thus, the data rate of PCAM is 50kbps and that of PWAM is 49.9kbps (calculated according to [46]). Figure 5.3 shows BER performance of the same considered schemes under the same condition as Figure 5.2 except for a higher maximum Doppler frequency \( f_d = 202\text{Hz} \) (for \( f_c = 6\text{GHz} \) and \( v = 10\text{m/sec} \)). In this case, the data rate of PCAM is still 50kbps while that of PWAM reduces to 49.5kbps, because larger percentage of the total transmitted pulses is used as pilots [46]. For both values of \( f_d \), our proposed PCAM performs close (within 2dB) to the case of ideal channel estimate. The performance of the PWAM schemes is substantially worse with the distributed PWAM performs slightly better than the preamble PWAM at the expense of a larger detection delay. Comparing
Figure 5.3 with Figure 5.2, we can observe that the performance improvement of the proposed PCAM over the PWAM schemes is more evident in the faster time-varying channel \( f_d = 202\text{Hz} \), which indicates that PCAM is more capable in tracking channel variations than PWAM, while without sacrificing the data rate.

![Graph](image)

**Figure 5.2** BER vs. \( E_b/N_0 \) with \( f_d = 20\text{Hz} \), \( T_f = 200\text{ns} \) and \( N_t = 100. R_b = 50\text{kbps} \) for PCAM, and \( R_b = 49.9\text{kbps} \) for distributed PWAM and preamble PWAM.

Figure 5.4 shows the values of \( \beta \) used in the above simulations as a function of \( E_b/N_0 \), which are calculated according to (5.20) and (5.3). We observe that \( \beta \) becomes larger for accommodating the faster fading channel. This is because a larger \( f_d \) (hence a smaller \( \alpha \)) will lead to a shorter forgetting factor \( \rho \) and therefore \( \beta \) needs to be larger. We also observe that \( \beta \) decreases as \( E_b/N_0 \) increases. This can be explained by referring to (5.17) and (5.16). At low \( E_b/N_0 \), the output SNR in (5.17) is dominated by the channel estimation MSE \( \sigma_e^2 \) and larger \( \beta \) can reduce \( \sigma_e^2 \) as indicated by (5.16), thus large \( \beta \) at low \( E_b/N_0 \) is required for a desirable output SNR; while at high \( E_b/N_0 \), the channel estimation MSE \( \sigma_e^2 \) in (5.16) and the output SNR in (5.17) are both dominated by \( E_b/N_0 \), smaller \( \beta \) can increase \( (1-\beta) \) so as to increase the output SNR.

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Figure 5.3 BER vs. $E_b/N_0$ with $f_o = 202$Hz, $T_f = 200$ns and $N_s = 100$. $R_b = 50$kbps for PCAM, and $R_b = 49.5$kbps for distributed PWAM and preamble PWAM.

Figure 5.4 $\beta$ vs. $E_b/N_0$ with $T_f = 200$ns, $N_s = 100$, and $L_c = 20$. 
We have assumed that coarse timing is acquired [68], but imperfect timing may induce erroneous channel estimates and thereby affect the performance. Figure 5.5 illustrates the BER performance of the PCAM and the distributed PWAM with uniformly distributed random timing offsets, under the same condition as Figure 5.2. We observe that the two schemes show comparable tolerances to the effect of timing offsets. For typical timing offset values that are on the order of 10ps [70], the performance degradation is negligible compared with the ideal case of accurate timing. However, timing offsets that are uniformly distributed on [-25ps, 25ps] incur substantial performance losses. This is due to the use of ultra-narrow \(T_w = 0.5\text{ns}\) pulses.

Figure 5.5 BER vs. \(E_b/N_0\) for PCAM and distributed PWAM, with random timing offsets uniformly distributed on [-10ps, 10ps] and on [-25ps, 25ps], \(f_d = 20\text{Hz}, T_f = 200\text{ns}\), and \(N_t = 100\).

Figure 5.6 compares the BER performance of \(|LLR|\)-T-GSC, \(|LLR|\)-GSC [41], and SNR-GSC [34] for Rake receivers with \(L_c = 5, 10, 20\) fingers. MRC is also included for comparison. Channel estimations for all the considered combining schemes are obtained using the proposed PCAM with \(T_f = 100\text{ns}\) [46] and \(N_t = 10\). Among the three GSC schemes, for each given value of \(L_c\), \(|LLR|\)-GSC provides the best performance, closely
followed by $|\text{LLR}|$-T-GSC, while SNR-GSC performs substantially worse. It is worth noting that both $|\text{LLR}|$-GSC and SNR-GSC require the estimation of the complete multipath channel so that they can perform path selection by ranking all received paths by their magnitudes of LLR or SNR, whereas $|\text{LLR}|$-T-GSC only needs to estimate a first portion of the channel impulse response. As the number of Rake fingers increases, the performance loss of the GSC schemes compared with MRC decreases. With $L_c = 20$, the performance of $|\text{LLR}|$-T-GSC is reasonably close to MRC—performance loss is on the order of 2dB. It is also shown that $|\text{LLR}|$-T-GSC with $L_c = 5$ outperforms SNR-GSC with $L_c = 10$, and $|\text{LLR}|$-T-GSC with $L_c = 10$ performs comparably to SNR-GSC with $L_c = 20$. The performance improvement of $|\text{LLR}|$-T-GSC over SNR-GSC is more evident at smaller value of $L_c$.

![Graph showing BER vs. $E_b/N_0$](image)

**Figure 5.6** BER vs. $E_b/N_0$ with $f_c = 20$Hz, $T_f = 100$ms, and $N_t = 10$, where channel estimation is obtained using the proposed PCAM for all the GSC schemes and MRC.

To better appreciate the efficiency of the proposed adaptive threshold $|h_t|_m$, Table 5.1 provides the average values of the number of estimated paths ($L_c$), the scaling factor
$K$, and the number of iterations for adjusting $K$, obtained from the simulations of Figure 5.6, where the adjustment of $K$ starts with $K = 30$dB and $\Delta = 1$dB. We can see that for $|\text{LLR}|$ -T-GSC, the number of paths required to estimate is much less than the number of resolvable multipath components, which is in general on the order of one thousand for the 3.5-10.5GHz band considered here [30]. For each listed value of $L_c$, the number of estimated paths decreases as $E_b/N_0$ increases. This is because stronger paths are generally with smaller delays [39] and channel estimates are less biased at higher SNR according to (5.16) and (5.23), consequently $|\text{LLR}|$ -T-GSC can find the strong paths more efficiently at higher SNR by estimating less paths. The value of $K$ increases with $E_b/N_0$. Keeping in mind that $K$ represents the ratio of path power gain to channel estimation MSE as indicated by (5.24) and the estimation MSE decreases as SNR increases, we can understand that larger $K$ for higher SNR guarantees desirable path gains of the selected paths.

Table 5.1 Average Values of the Number of Estimated Paths ($L_c$), the Scaling Factor ($K$), and the Number of Iterations for Adjusting $K$ with an Original $K$ of 30dB and $\Delta = 1$dB, Obtained from the Simulations of Figure 5.6.

| $E_b/N_0$ (dB) | $L_c = 5$ | | $L_c = 10$ | | $L_c = 20$ |
|---|---|---|---|---|---|---|---|
| | $L_c$ | $K$ (dB) | Iterations | $L_c$ | $K$ (dB) | Iterations | $L_c$ | $K$ (dB) | Iterations |
| 0 | 76.1 | 23.9 | 9.1 | 103.8 | 23.2 | 9.8 | 148.1 | 22.2 | 10.8 |
| 2 | 75.9 | 24.6 | 8.4 | 103.7 | 23.9 | 9.1 | 147.3 | 22.9 | 10.1 |
| 4 | 75.2 | 25.3 | 7.7 | 103.4 | 24.6 | 8.4 | 146.7 | 23.6 | 9.4 |
| 6 | 75.1 | 26.0 | 7.0 | 103.0 | 25.3 | 7.7 | 146.6 | 24.2 | 8.8 |
| 8 | 74.9 | 26.8 | 6.2 | 102.8 | 26.0 | 7.0 | 145.8 | 24.9 | 8.1 |
| 10 | 74.6 | 27.6 | 5.4 | 102.0 | 26.8 | 6.2 | 145.1 | 25.7 | 7.3 |
| 12 | 73.6 | 28.4 | 4.6 | 101.4 | 27.5 | 5.5 | 145.0 | 26.4 | 6.6 |
| 14 | 72.0 | 29.3 | 3.7 | 99.0 | 28.4 | 4.6 | 144.1 | 27.2 | 5.8 |
| 16 | 70.6 | 30.3 | 3.3 | 97.5 | 29.3 | 3.7 | 143.7 | 28.0 | 5.0 |
| 18 | 68.7 | 31.3 | 4.3 | 94.5 | 30.2 | 3.2 | 142.4 | 28.9 | 4.1 |
| 20 | 66.3 | 32.3 | 5.3 | 90.2 | 31.2 | 4.2 | 140.0 | 29.8 | 3.2 |

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Our design is carried out for a single-user system, however, the proposed PCA-|LLR|-T-GSC scheme can be applied to multi-user systems by treating the multi-user interference (MUI) as additive noise [11]. Figure 5.7 compares the BER performance of PCA-|LLR|-T-GSC (for \( L_c = 5, 10, 20 \)), PCA-MRC, distributed PWAM, and preamble PWAM, with the presence of 10 asynchronous users for \( T_f = 100 \text{ns} \), \( N_s = 10 \), and \( f_d = 202 \text{Hz} \). It is obvious that the pilot-channel-assisted (PCA) combining schemes exhibit superior resistance to MUI to the PWAM schemes. This is because the small number of pilot pulses used in PWAM [46] makes it vulnerable to MUI. With the number of Rake fingers \( L_c > 5 \), PCA-|LLR|-T-GSC significantly outperforms the PWAM schemes.

![Figure 5.7 BER vs. \( E_b/N_0 \) in the presence of 10 asynchronous users, for \( f_d = 202 \text{Hz} \), \( T_f = 100 \text{ns} \), and \( N_s = 10 \).](image)

**5.4 Conclusions**

We have proposed a PCA-|LLR|-T-GSC scheme to provide proper path selection for UWB Rake receivers with limited number of fingers. The PCAM for channel estimation optimally balances the channel estimation and receiver detection performance, without
incurring any loss in the overall system data rate. The performance of PCA-$|\text{LLR}|$-T-GSC has been evaluated through extensive simulations. The simulation results show that PCAM significantly outperforms the PWAM [46] in time-variant multipath channels and $|\text{LLR}|$-T-GSC significantly outperforms SNR-GSC [34] and performs comparably to $|\text{LLR}|$-GSC [41], both of which require the estimation of all received paths for path selection, whereas $|\text{LLR}|$-T-GSC only needs to estimate a first portion of the channel impulse response. Simulation results also reveal the superior resistance of PCAM to multi-user interference and its comparable tolerance to timing offsets, compared with PWAM.
Chapter 6

Conclusions and Future Work

In this chapter, we provide some concluding remarks on this thesis and discuss the potential directions for future work.

6.1 Conclusions

In this thesis, we have investigated several important issues in UWB direct-sequence impulse radio wireless communications. The main contributions of this thesis are summarized as follows.

We first investigated the potential to increase the power efficiency in UWB impulse radio wireless communications by employing multidimensional modulation. Toward this end, we proposed a $2MN$-ary biorthogonal keying (BOK) scheme, using both $N$ (number of orthogonal pulses) and $M$ (size of orthogonal code words) to increase the modulation dimensionality. Extensive simulations have confirmed that our proposed $2MN$-ary BOK scheme is capable of providing more power-efficient modulation for UWB impulse radio direct-sequence multiple-access communications in dense multipath environments, compared with unidimensional bipolar pulse amplitude modulation (PAM) and conventional biorthogonal-code keying. Using a small size (e.g. $M = 2$ or 4) of orthogonal codes and two ($N = 2$) orthogonal pulse shapes, the $2MN$-ary BOK scheme shows promise for the practical and power-efficient implementation of power-limited UWB communications.

Following that, we provided a detailed performance analysis for direct-sequence multiple-access (DS-MA) impulse radio systems. Specifically, our analysis approach explicitly examines the impact of pulse overlaps between pulses arriving at the receiver, which results from the fine multipath delay resolution of UWB impulse radio signals. A closed-form bit error rate (BER) expression incorporating the channel-induced pulse-overlap is derived and compared with simulations that employ UWB signals complying
with the FCC UWB indoor emission mask and the channel model recommended by IEEE 802.15.3a. The great gaps between the actual BER curves obtained through realistic simulations and the ones obtained by assuming no pulse overlap induced by the channel show that, channel-induced pulse overlap can degrade the system performance significantly and the results of our performance analysis are important for system design and performance evaluation of DS-MA impulse radio systems.

Finally, we proposed a pilot-channel assisted (PCA) generalized selection combining (GSC) scheme with log-likelihood ratio (LLR) threshold test per path (PCA-LOOR-T-GSC), to provide reliable channel estimation and tracking and proper path selection for UWB Rake receivers with limited number of Rake fingers. Extensive simulations have shown that our proposed PCA modulation for channel estimation significantly outperforms the PWAM scheme [46] in time-varying multipath channels, and the proposed LOOR-T-GSC markedly outperforms SNR-GSC [34] and performs comparably to LOOR-GSC [41], both of which require the estimation of all received paths for path selection, whereas LOOR-T-GSC only needs to estimate a first portion of the channel impulse response. Simulation results also reveal the superior resistance of PCA modulation to multi-user interference and its comparable tolerance to residual timing offsets, compared with PWAM. The relatively low complexity of PCA-LOOR-T-GSC and its superior capability to deal with rapidly fluctuating channels make it desirable for practical selective-Rake receiver implementation in UWB impulse radio communications.

6.2 Future Work

There are several possible directions that can follow the work done in this thesis.

6.2.1 Range Extension

It is worth noting that although the current UWB regulations valid in the United States and envisaged in Europe constrain long-range deployment of UWB devices [2], this situation might change, because it is anticipated that future UWB regulations will likely
evolve to an even greater extent than those experienced for other wireless technologies. The low average transmission power-spectral-density (PSD) naturally limits UWB to short ranges, however, UWB represents a tradeoff between lower spectral efficiency for increased power efficiency to achieve a desired rate/range operating point while maintaining a constant average PSD [16]. For example, increasing the occupied bandwidth, reducing the pulse repetition frequency, or using more pulses to send one information bit, the distance achieved by the UWB system can be increased for a fixed average PSD. It is thus of interest for us to investigate methods for increasing the communication range of a UWB system, and the feasibility and realization of a UWB radio system that can dynamically trade data rate, power consumption, and range.

6.2.2 Timing Synchronization

We have proposed in Chapter 5 a pilot-channel assisted modulation scheme for channel estimation in UWB DS impulse radio wireless communications, while intertwined with channel estimation is the issue of timing recovery. The same properties that give prominence to UWB impulse radio technology also lead to challenges in timing synchronization, as we have discussed in Section 2.2.7. There are two major issues that make synchronization one of the main challenges of UWB. Firstly, the search space becomes very large, and secondly, very high precision (down to the nanosecond level) is required. A full serial search can cause the acquisition time to be unacceptably long due to the fine multipath delay resolution and the stringent power constraints of UWB communications. However, a full parallel search shortens the acquisition time at the expense of increased hardware complexity and power consumption of the receiver. Therefore, it is worthwhile to study a practical algorithm for initiating and maintaining synchronization at the UWB receiver.

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Appendix A

Derivation of the Expression of $MUI_n(q)$ in (4.14)

For DS-MA BOPK signals, the $MUI_n(q)$ term in (4.5) can be expressed as

$$
MUI_n(q) = \sum_{u=2}^{N_u} \frac{E_u}{N_s} \sum_{i=q}^{N_u-1} h_{1,t_e}^{H} b_0 \sum_{j=1}^{N_x} \sum_{k=-\infty}^{\infty} H_{u,j,t_e} d_u \left( k, t - (k-i)T_f - (\tau_u - \tau_1) - j\Delta\tau \right) w_n(t) dt
$$

(A.1)

Let $-(k_u - i)T_f - (\tau_u - \tau_1) - l_u \Delta\tau = 0$, in which $l_u$ is an integer satisfying $0 \leq |l_u| \leq N_f$, and

$$
k_{u,i} = \begin{cases} i + \left[ \frac{\tau_1 - \tau_u}{T_f} \right] - 1 & \text{if } \tau_u < \tau_1, \\ i & \text{if } \tau_u = \tau_1, \\ i + \left[ \frac{\tau_1 - \tau_u}{T_f} \right] + 1 & \text{otherwise.} \end{cases}
$$

(A.2)

$$
MUI_n(q) = h_{1,t_e}^{H} \sum_{u=2}^{N_u} \frac{E_u}{N_s} \sum_{j=1}^{N_x} \sum_{k=-\infty}^{\infty} \sum_{i=q}^{N_u-1} c_{1,i} b_0 H_{u,j,t_e} d_u \left( k_{u,i} + l_u \Delta\tau - j\Delta\tau \right) w_n(t) dt
$$

(A.3)

where $f_{u,j,t_e}(i,t)$ is the $L_c \times 1$ vector composed of the upper $L_c$ components of $f_{u,j}(i,t)$ given below

$$
f_{u,j}(i,t) = \begin{bmatrix} 0_{L_c \times (N_u-1)} I_{u} H_{u,j} d_u (k_u, t) \\ I_{L_c} \delta_{l_u} H_{u,j} d_u (k_u, t) \\ 0_{L_c \times (N_u-1)} I_{u} H_{u,j} d_u (k_u, t) \\ I_{L_c} \delta_{l_u} H_{u,j} d_u (k_u, t) \end{bmatrix}$$

(A.4)

otherwise.

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Appendix B

PSD of DS-MA BOPK Signals

The PSD of the signal $s_n(t)$ in (4.1) can be computed by first computing the autocorrelation function of $s_n(t)$ and then taking the Fourier transform.

The autocorrelation function of $s_n(t)$ is

$$R_n(t + \tau, t) = \mathbb{E}[s_n(t) s_n(t + \tau)]$$

$$= \frac{E_u}{N_s} \sum_{k=\infty}^{\infty} \sum_{l=\infty}^{\infty} \mathbb{E}[e^{-j\omega T_f} b_{u,k} b_{u,l} w_{d,u,k}(t - kT_f) w_{d,u,l}(t + \tau - kT_f)]$$

$$= \frac{E_u}{N_s} \sum_{k=\infty}^{\infty} \mathbb{E}[w_{d,u,k}(t - kT_f) w_{d,u,k}(t + \tau - kT_f)]$$

$$= \frac{E_u}{MN_s} \sum_{m=1}^{M} \sum_{k=\infty}^{\infty} \mathbb{E}[w_m(t - kT_f) w_m(t + \tau - kT_f)]$$

(B.1)

in which $d_{u,k} \in \{1, 2, \ldots, M\}$ for different $\tilde{k}$ are independent and equiprobable, and $R_n(t + \tau, t)$ is periodic in the $t$ variable with period $T_f$ [32]. The dependence of $R_n(t + \tau, t)$ on the $t$ variable can be eliminated by averaging $R_n(t + \tau, t)$ over a single period [32]. Thus,

$$R_n(\tau) = \frac{1}{T_f} \int_0^{T_f} R_n(t + \tau, t) dt$$

$$= \frac{E_u}{MN_s} \sum_{m=1}^{M} \mathbb{E} \left[ \sum_{k=\infty}^{\infty} \frac{1}{T_f} \mathbb{E} \left[ w_m(t - kT_f) w_m(t + \tau - kT_f) dt \right] \right]$$

(B.2)

$$= \frac{E_u}{MN_s T_f} \sum_{m=1}^{M} \int_{-\infty}^{\infty} w_m(t) w_m(t + \tau) dt$$

$$= \frac{E_u}{MN_s T_f} \sum_{m=1}^{M} R_{mm}(\tau)$$

where $R_{mm}(\tau)$ is the autocorrelation function of the pulse $w_m(t)$ ($m = 1, \ldots, M$).

The Fourier transform of $R_n(\tau)$ yields the (average) PSD of $s_n(t)$ in the form of

$$P_s(f) = \frac{E_u}{N_s T_f} \frac{1}{M} \sum_{m=1}^{M} |W_m(f)|^2$$

(B.3)

where $W_m(f)$ is the Fourier transform of the pulse $w_m(t)$ ($m = 1, \ldots, M$).
Appendix C

Derivation of the Expression in (5.18)

In order to find the $\beta_{\text{opt}} = \arg \max_{\beta} \bar{y}_b(\beta)$, we need to solve $\frac{d\bar{y}_b(\beta)}{d\beta} = 0$, where $\bar{y}_b(\beta)$ is given by (5.17). After some ordinary manipulations, it turns out that the problem reduces to solving the following equation

$$\sum_{i=0}^{N-1} \sigma_{\text{opt}}^2(l,\beta) + \sum_{i=0}^{N-1} |h_i|^2 + \frac{2(1-\beta)^2}{N_s} \sum_{i=0}^{N-1} |h_i|^2 \frac{d\sigma_{\text{opt}}^2(l,\beta)}{d\beta} + (1-\beta) \sum_{i=0}^{N-1} \frac{d\sigma_{\text{opt}}^2(l,\beta)}{d\beta} = 0$$

(C.1)

Let $(1-\alpha)|h_i|^2 = A_i$ and $\frac{E_s}{N_s N_0} = C$ in (5.16), then

$$\sigma_{\text{opt}}^2(l,\beta) = \left[ C\beta + \sqrt{C\beta \left( C\beta + \frac{1}{A_i} \right)} \right]^{-1} = \frac{A_i}{C\beta}$$

(C.2)

$$\frac{d\sigma_{\text{opt}}^2(l,\beta)}{d\beta} = \frac{-A_i^2 C \left( 2 \sqrt{A_i C \beta + 1} \right)}{2 \sqrt{A_i C \beta + 1} \left( A_i C \beta + \sqrt{A_i C \beta} \right)^2} = -\frac{\sqrt{A_i}}{2 \beta \sqrt{C \beta}}$$

(C.3)

Substituting (C.2) and (C.3) into (C.1), we have

$$\beta \sum_{i=0}^{N-1} \frac{A_i}{C} + \beta \sqrt{\beta \sum_{i=0}^{N-1} |h_i|^2} - (1-\beta)^2 \sum_{i=0}^{N-1} |h_i|^2 \sqrt{A_i C} - \frac{1-\beta}{2} \sum_{i=0}^{N-1} \frac{A_i}{C} = 0$$

(C.4)

Letting $\lambda = \sqrt{\beta}$, $U = \frac{1}{2} \sum_{i=0}^{N-1} \sqrt{\frac{A_i}{C}}$, $V = \sum_{i=0}^{N-1} |h_i|^2 \sqrt{A_i C}$ and $W = \sum_{i=0}^{N-1} |h_i|^2$, (C.4) is reduced to

$$V\lambda^4 - W\lambda^3 - (3U + 2V)\lambda^2 + U + V = 0$$

(C.5)

Since $V \ll 1$ for low SNR values and $\lambda < 1$, we can approximate that $V\lambda^4 = 0$ and (C.5) is further simplified to

$$\lambda^3 + a\lambda^2 - b = 0$$

(C.6)

in which $a = (3U+2V)/W$ and $b = (U+V)/W$.

Solving (C.6), we obtain that
$$\lambda = \left( \frac{1}{6} + \frac{2}{3} a^2 \right) \sqrt{108b - 8a^3 + 12\sqrt{81b^2 - 12a^3b}} - \frac{a}{3} = \sqrt[3]{b} - \frac{a}{3} \quad \text{(C.7)}$$

where the approximation is made by $a^2 = 0$ and $a^3 = 0$.

Substituting the above definitions of $a, b, U, V$ and $W$ into (C.7) and with $\lambda = \sqrt[3]{b}$, we obtain the expression for the optimal $\beta$ as follows

$$\beta_{\text{opt}} = \left[ \sqrt{\frac{(1-\alpha)N_s N_0}{4E_b} \sum_{k=1}^{L} |h_{k}|^2 + \frac{(1-\alpha)E_b}{N_s N_0} \sum_{k=1}^{L} |h_{k}|^3} - 3 \sqrt{\frac{(1-\alpha)N_s N_0}{4E_b} \sum_{k=1}^{L} |h_{k}| + 2 \frac{(1-\alpha)E_b}{N_s N_0} \sum_{k=1}^{L} |h_{k}|^3} + \frac{3}{2} \sum_{k=1}^{L} |h_{k}|^2 \right]^{-2} \quad \text{(C.8)}$$
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