Modeling of DC-OFDM UWB physical layer in Matlab and design of demapping module with high diversity gain

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Abstract

Due to the development of semiconductor and digital communication technology, amounts of wireless communication technology have been presented recently. Among those techniques, Ultra Wide Band (UWB) technology is an outstanding short distance wireless communication technology. With the characters of fast speed, low power consumption and high security, it is a perfect communication technology for Wireless Personal Area Network (WPAN).

According to the application environments of the UWB techniques, the International Standard Organization (ISO) has passed the European Computer Manufacturers Association (ECMA) 368 standard to be the UWB physical layer standard. Because of the wide bandwidth and frequency spectrum characters of UWB system, every country drafts their own UWB standards according to their own situations. For the system implementation, modeling is the first important part. The accuracy and practicality of model are in high accordance with the hardware implementation. The thesis main work is to set up the Matlab model of the Dual Carriers-Orthogonal Frequency Division Multiplexing (DC-OFDM) UWB physical layer system which is designed by China Electronics Standard Institute (CESI). In the modeling course, the thesis analyzes the differences between hardware implementation and algorithm simulation, and use fix-pointed to quantize the inputs and outputs of the modules, so that the model is more practical. In the system simulation, by changing the Sample Frequency Offset (SFO), Carrier Frequency Offset (CFO), Noise and Multi-path interferences, the model is verified and sets up the stable foundation for the hardware design. What’s more, in order to make full use of the wide band, this thesis uses diversity demapping method for the Dual Carrier Modulation. According to the biggest combine ratio diversity technique and estimated channel diversity technique, the biggest diversity gain can be got, and the performance is improved by 1.5dB. Considering the diversity demapping algorithm, in the hardware implementation, mapping and demapping modules are designed for hardware reuse. This can not only decrease the area and the power, but also be compatible with ECMA368 and DC-OFDM standard.
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<td>ADC</td>
<td>Analog to Digital Converters</td>
</tr>
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<td>ADSL</td>
<td>Asymmetric Digital Subscriber Loop</td>
</tr>
<tr>
<td>APK</td>
<td>Combination of ASK and PSK</td>
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<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<td>BPSK</td>
<td>Binary Phase shift keying</td>
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<td>CDMA</td>
<td>Code Division Multiple Access</td>
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<td>CESI</td>
<td>China Electronics Standard Institute</td>
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<tr>
<td>CFO</td>
<td>Carrier Frequency Offset</td>
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<td>CP</td>
<td>Cyclic Prefix</td>
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<tr>
<td>CPFSK</td>
<td>Continuous-Phase FSK</td>
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<tr>
<td>DAA</td>
<td>Detect and Avoid</td>
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<tr>
<td>DAB</td>
<td>Digital Audio Broadcasting</td>
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<td>DAC</td>
<td>Digital to Analog Converter</td>
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<td>DCM</td>
<td>Dual-Carrier Modulation</td>
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<td>DC-OFDM</td>
<td>Dual Carriers-OFDM</td>
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<td>DS</td>
<td>Direct Sequence</td>
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<tr>
<td>ECMA</td>
<td>European Computer Manufacturers Association</td>
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<tr>
<td>EIRP</td>
<td>Effective Isotropic Radiated Power</td>
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<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
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<td>FDM</td>
<td>Frequency Division Modulation</td>
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<tr>
<td>FFI</td>
<td>Fixed Frequency Interleaving</td>
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<td>FFT</td>
<td>Fast Fourier Transformation</td>
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<td>FSK</td>
<td>Frequency shift keying</td>
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<tr>
<td>HCS</td>
<td>Header Check Sequence</td>
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<tr>
<td>IF</td>
<td>Intermediate frequency</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse FFT</td>
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<tr>
<td>ISI</td>
<td>Inter-Symbol Interference</td>
</tr>
<tr>
<td>ISO</td>
<td>International Standard Organization</td>
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<tr>
<td>ITRS</td>
<td>International Technology Roadmap for Semiconductor</td>
</tr>
<tr>
<td>ITU</td>
<td>International Telecommunication Union</td>
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<tr>
<td>LTE</td>
<td>Long Term Evolution</td>
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<td>LNA</td>
<td>Low-Noise Amplifier</td>
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<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>MAC</td>
<td>Media Access Control</td>
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<tr>
<td>MB</td>
<td>Multi-Band</td>
</tr>
<tr>
<td>MBOA</td>
<td>Multi-band OFDM Alliance</td>
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<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
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<tr>
<td>PAPR</td>
<td>Peak and Average Power Ratio</td>
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<tr>
<td>PDF</td>
<td>Probability Density Function</td>
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<tr>
<td>PLCP</td>
<td>Physical Layer Convergence Protocol</td>
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<td>PRBS</td>
<td>Pseudo-Random Binary Sequence</td>
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<td>PSK</td>
<td>Phase shift keying</td>
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<td>QAM</td>
<td>Quadratic Amplitude Modulation</td>
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<td>QPSK</td>
<td>Quaternary Phase Shift Keying</td>
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<tr>
<td>RF</td>
<td>Radio Frequency</td>
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<tr>
<td>SoC</td>
<td>System on Chip</td>
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<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
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<tr>
<td>TFC</td>
<td>Time Frequency Code</td>
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<tr>
<td>UWB</td>
<td>Ultra Wide Band</td>
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<tr>
<td>VLSI</td>
<td>Very Large Scale Integrated</td>
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<tr>
<td>WLAN</td>
<td>Wireless Local Area Network</td>
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<td>WPAN</td>
<td>Wireless Personal Area Network</td>
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Chapter 1 Introduction

1.1 Research Background

1.1.1 Introduction of the UWB Communication Technology

The history of the UWB technology can be traced back to 1942 when De Rosa filed in a patent on random pulse system. In the 1960's, further research on time-domain electromagnetism promoted the development of the UWB technology. In 1973, Sperry was granted the first patent related to the UWB communication technology. Since then, the theory and technology on UWB have been developed rapidly and a number of related equipment has become available. In 1989 in particular, the U.S. Department of Defense coined the name "ultra-wideband", a term that has been widely accepted by the industry as well as the academia.

Owing to its high precision in timing, UWB has been applied to many radar systems. With the development of high-speed exchange technologies, it has gradually been found that UWB is actually of potential in other low-cost communication applications, especially for short distance wireless communication systems. On April 22\textsuperscript{nd}, 2002, the United States Federal Communications Commission (FCC) released an initial set of regulations on the UWB devices’ power and spectrum for the civilian use. According to FCC, a UWB device is defined as the device whose relative bandwidth is greater than 0.2 or the transmission bandwidth is greater than 500 MHz. Here, the relative bandwidth is defined as: 

\[ f_c = \frac{f_H - f_L}{(f_H + f_L)/2} \]

where \( f_H \) and \( f_L \) are, respectively, the highest and lowest frequencies of the system (at -10 dB level). According to FCC’s definition for UWB indoor communications, the actual use of the spectrum ranges from 3.1 to 10.6 GHz, and in this regulation, the Effective Isotropic Radiated Power (EIRP) couldn’t exceed -41.3 dBm/MHz. Figure 1-1 shows the UWB communication spectrum and power limitation defined by FCC.

Shannon’s information theory reveals that the channel capacity is directly proportional to the bandwidth: 

\[ C = B \times \log_2(1 + SNR) \]

Obviously, UWB technology can offer very high data transfer rate beyond the low power emission. In 2007, China
began to formulate the national standard of the UWB system. In order to be compatible with the FCC regulations and also satisfy with our countries characters, the standard is formulated according to ECMA368, but keeping our own characters.

Ultra-wideband technology has many advantages. Wireless, low-power, high-speed and high security are some of the greatest. It is positioned at the transfer rate of 100~1000 Mbps. In contrast, the existing short-range wireless data transmission technologies such as Bluetooth only offer 1 Mbps; Wireless Local Area Network (WLAN) 802.11a/g is somewhat better for its 10~20 Mbps data transmission capability. Although the WLAN transmission distance can be up to about 50 meters, the UWB transmission distance, i.e., 2 to 10 m, can completely satisfy the indoor and desktop applications. With UWB technology, consumers can easily achieve high-speed wireless connections among varieties of digital indoor devices (such as notebook computers, digital audio and video equipment, digital consoles and mobile phones, etc.).[1]

![Figure 1-1 Restrictions FCC provided for indoor and outdoor UWB power](image-url)

1.1.2 Comparison of Different Technical Solutions for UWB Communications

Ever since FCC opened the UWB frequency band, the battle around the ultra-wideband standard has not seen a break. The IEEE 802.15.3 Working Group has today two main proposals: Direct Sequence (DS)-based spread spectrum ultra-wideband program initiated by DS-UWB Forum [2], and Multi-Band (MB) Orthogonal Frequency Division Multiplexing (OFDM) ultra-wideband program advocated by the WiMedia Alliance [3]. These two major programs have been carried...
out with a four-year dispute in the ISO standard. In March 2007, the WiMedia Alliance's MB-OFDM standard was eventually adopted by ISO, and has formally become the first international UWB standard. The DS-UWB proposal has basically been abandoned.

The DS-UWB uses two frequency bands: 3.1 GHz~5.15 GHz and 5.825 GHz~10.6 GHz, based on the single-carrier direct sequence spread spectrum, and the bandwidth is 1.3 GHz. While the MB-OFDM proposal divides the frequency band 3.1 GHz-10.6 GHz into 14 sub-bands which is 528 MHz wide for each, through time-interleaving, sending information in sub-bands using narrow time OFDM symbols.

In what follows, the two programs are compared in their spectrum flexibility, bandwidth requirements, multi-path energy collection capability, Radio Frequency(RF) front-end complexity receiver and degree of difficulty [4][5][6]:

(1) The MB-OFDM system is better in the flexibility of the spectrum.

The MB-OFDM program divides the frequency band into 14 sub-bands with a bandwidth of 528 MHz for each sub-band. Each sub-band can work independently or in coordination with other operations, thus with high flexibility. In addition, in order to comply with the global spectrum planning, UWB is expected to use the Detect and Avoid (DAA) technology. When the UWB device detects other wireless devices such as WiMAX, WLAN at work in the band, it will automatically switch to another free band. This is relatively simple for the MB-OFDM technology when the transmitter doesn’t send data to the frequency points correlated to the OFDM symbols. For the DS-UWB, this is more difficult due to the filtering technology.

(2) The MB-OFDM system reduces the data conversion bandwidth between the analog and digital circuits.

The average bandwidths of the DS-UWB and MB-OFDM systems are identical. And the two systems operate at almost the same transmission power. The main advantage of the MB-OFDM program is that its instantaneous bandwidth could not exceed 528 MHz. This means that the requirements for the channel selection filters and variable gain amplifier bandwidths will be about three times lower than those for the DS-UWB program. In addition, the MB-OFDM system doesn’t impose more demands on Digital to Analog Converter (DAC) and Analog to Digital Converters (ADC), because for the MB-OFDM proposal, the request rate is lower than what the DS-UWB proposal would impose.
The OFDM system has better performance in the multi-path resistance and reduces the complexity of the receiving circuit.

When the OFDM system completes the high-speed data conversion from the serial to parallel, it will modulate the data onto N orthogonal sub-carriers and transmit them simultaneously. On each sub-carrier, the duration time of the symbol is increased by times of N, which will effectively reduce inter-symbol interferences and the complexity of the equalizer of the receiver circuit. Even without the equalizer, the inter-symbol interference can be eliminated by inserting a cyclic prefix or zero-fill prefix which is longer than the maximum delay spread before the OFDM symbol. By comparing the 16-Path rake-receiver with that of the MB-OFDM system, the performance of the 112/224/448 Mbps rate in the DS-UWB system is found to be inferior to that of the 110/200/480 Mbps rate in the MB-OFDM system, by 1.5, 4 and 6 dB, respectively[6].

The MB-OFDM system reduces the bandwidth of the RF front-end, which leads to reduction of the design complexity and power.

The DS-UWB RF front-end uses the single-channel structure, which is simpler than that of the MB-OFDM system. However, the bandwidth of the radio-frequency circuit is greater than 1.3 GHz, so it is very difficult to design. In addition, the power consumption is very high. This affects the future DS-UWB RF front-end System on Chip (SoC) embedded applications. The MB-OFDM RF front-end, whose bandwidth is 528 MHz, reduces the realization complexity and power of the RF circuit. Although the inter-band signal separation problem still persists, there are several mature circuits that can be utilized to overcome this problem.

It is therefore believed that the MB-OFDM solution is more suitable than the DS-UWB for high-speed UWB communication system. Because the CESI standard has all the merits of MB-OFDM UWB system, it’s also suitable foe the high-speed UWB system.

1.1.3 Development of OFDM Technology

OFDM is essentially a frequency division multiplexing technology. It was put forward as early as in the 1960’s. It divided the available bandwidth into several sub-bands separated from each other and sent low-speed signals (such as AT & T) at the same time. This traditional multi-carrier transmission was relative complex, because each sub-carrier needed its own analog front-end. In order to make the
receiver distinguish the different sub-bands, each sub-band must have enough space so as to avoid the spectrum aliasing after channel. Thus, it usually has very low spectral efficiency. However, in this parallel transmission mechanism, because of the lower data rate transmissions on the sub-carriers, the corresponding signals had longer cycles for symbols than the maximum channel delay spread. Such characteristics could effectively reduce the inter-symbol interference caused by the delay expansion of the channel units.

In order to increase the spectrum utilization of the Frequency Division Modulation (FDM) technology, G. A. Doelz put forward the Kineplex system in the 1950’s [7]. The design goal of the system was to send the data in a series of multi-path fading wireless channels. The system used 20 sub-carriers, with differential Quaternary Phase Shift Keying (QPSK) modulation. The spectrums of the sub-carriers overlap, but the sub-carriers are orthogonal with one another. So, the spectrum efficiency could be greatly improved. The system embodied the core idea of the OFDM technology, and could be seen as the embryonic form of the OFDM system. In order to limit the system spectrum, Chang analyzed how to make the sub-carriers hold orthogonal in the MB-OFDM communication system [8]. Later, Weinstein and Ebert put forward a solution to implementing the multi-carriers base-band modulation & demodulation modules with discrete Fast Fourier Transformation (FFT) transform [9]. With this solution, it would no longer have to use the analog front-end before each sub-carrier. As a result, it would greatly reduce the complexity of the multi-carrier system, and make great contribution to the OFDM evolution. In 1980, Peled and Ruiz proposed to insert the Cyclic Prefix (CP) in order to eliminate the symbol interference [10]. Their approach could hold the sub-carriers orthogonal after the multi-path channel. By then, the modern concept of the OFDM was confirmed. Later, Cimini used the OFDM concept to the cell mobile communication system, and found the base for the modern wireless OFDM system [11].

Now, the OFDM technology has been applied to many communication domains. For example, the European Digital Audio Broadcasting (DAB) standard uses the differential phase OFDM modulation [12]; Cable telephone line is based on the existing copper twisted pair of Asymmetric Digital Subscriber Loop (ADSL) [13]; 802.11a wireless LAN standard based on the 5-GHz band has also adopted the OFDM modulation [14]. The 5-GHz band of 802.11a wireless LAN standard has also adopted
the OFDM modulation with Quadratic Amplitude Modulation (QAM). In short, from WLAN to WiMAX, Flash-OFDM, from Long Term Evolution (LTE) to Beyond 3Generation (B3G), and then to ultra-wideband wireless communications technology, OFDM has become a technology logo of the new generation wireless communications.

1.1.4 Characteristics of the OFDM Technology

The OFDM technology is used for multi-carrier transmission, with N orthogonal sub-carriers dividing the whole channel into N sub-channels. The N sub-channels in turn transmit the information simultaneously. The technology implements the signal modulation and demodulation using FFT. The main advantages include:

(1) High spectrum efficiency

In order to separate the signal from different channels, the ordinary FDM system always sets up a certain guard period in the adjacent inter-channel (band) in order that the receiver can use the band-pass filter to isolate the signal from the corresponding signal channel. This approach results in a spectrum resource waste. But in the OFDM system, there doesn’t exist the guard period between the inter-channels. What’s more, the main lobe of each signal is overlapping, but these signals are orthogonal in the frequency domain. In the time domain, these sub-carriers are orthogonal, and the demodulation of the sub-carriers is completed by those characteristics. In addition, the sub-carriers of the OFDM system can use M-order modulation (such as high spectral efficiency of QAM), and further improve the spectrum efficiency of the OFDM systems.

(2) Simplification of the implementation by employing Fast Fourier Transform technique

The OFDM system uses FFT in the transmission to modulate the frequency-domain data to the time-domain data. In the receiver, it employs FFT to convert the time-domain signal into the frequency-domain signal. It then goes to the decision demodulation. This scheme avoids the use of multiple sets of modems before the analog front-ends, and will reduce the circuit complexity. According to International Technology Roadmap for Semiconductor (ITRS), along with the advances in technology, the computing ability of the digital signal processing chips will get faster, and the chip size can also be scaled down [15], which will further promote the OFDM technology application and development.
(3) Flexible spectrum

The OFDM system divides the frequency band into multiple sub-carriers, and the signal bandwidth depends on the number of sub-carriers used. After selecting the appropriate sub-carriers for transmission, it can easily avoid the interference of wireless communication equipment on the frequency bands. Hence, it is compatible with global spectrum norms. At the same time, spectrum diversity and multi-user diversity can take full advantage in order to get the best system performance by combining with the adaptive technology for dynamic allocation of the spectrum resources.

(4) High ability of anti-multipath interference and anti-fading

The OFDM system uses a number of parallel sub-carriers to transmit information. In this way, the high-speed data with bit rate (r) can be transferred into the sub-data stream at a transfer rate of r/n. With this approach, the duration of the modulated symbols becomes greater than the channel delay spread. This will reduce the system sensitivity for the delay spread. The inter-symbol interference (ISI) in the fading channel will come down as well.

In addition, the OFDM system uses a method with CP insertion in the time-domain. Then, when the duration of CP is greater than the maximum delay spread, the inter-symbol interference and the orthogonal destruction both caused by the multi-path spread can be completely eliminated. To conclude, the OFDM system has good ability of anti-multipath interference. In addition, sub-carriers of the OFDM system will divide the whole channel into many narrow channels. Although the whole channel is likely to be extremely uneven fading channel, the decline of each channel is similar to flat. This scheme makes the equalizer of the OFDM channel particularly simple, and often a tap of the equalizer could be acceptable.

But there also exist shortcomings with OFDM. For instance, it is sensitive to frequency offset and phase noise. To distinguish the various sub-channels, the OFDM technology uses strict orthogonal property between the various sub-carriers. Frequency offset and phase noise will destruct the orthogonal property between the sub-carriers. Only 1% frequency offset could lead to a decreased signal to noise ratio by 30 db. On the other hand, the Peak and Average Power Ratio (PAPR) is large, and the high mean peak will increase the RF amplifier requirements, resulting in a decreased RF power amplifier efficiency.
1.2 Thesis Motivation

Nowadays, wireless communications play an indispensable role in people's life. High-speed, reliable wireless communications will gradually replace the existing wired network, providing people with more convenient connections. From ITRS2007 system driver, we can see the trend of the development. And in the short distance wireless communication domain, the UWB technique has many advantages as described above, which is accordance with the ITRS. So we think it has great potential in the future.[16] The MB-OFDM proposal won at last the ISO certification among several contenders. It has then officially become the first international standard. The DC-OFDM UWB standard which has been drafted out by the Information standard Committee is based on the MB-OFDM technique, and accords with the Chinese wireless frequency division.

At present, the UWB market for the Chinese market has not really started. It provides us with a rare 2-3 year research and development "pre-application" period. It will be of great significance if the chip design is studied aiming at promoting a Chinese system and a Chinese standard.

The OFDM technology has been widely used in various communication systems, and has formed a relatively mature algorithm system. In UWB systems, there is more demand on the receiver algorithm performance as a result of the intensive multi-path channel characteristics. At the same time, the UWB systems couldn't put more demand on the hardware complexity for its low-cost. Thus, specific designs aiming at low-power and low-cost are required in order to optimize the circuit as well as the appropriate algorithm.

1.3 UWB Technology Development

1.3.1 Market Process

As mentioned earlier, UWB was first approved in 2002 by FCC for civilian use. After 2003, the working group of the International Telecommunication Union (ITU) has begun to test UWB. In October 2005, ITU confirmed the allocation of the UWB spectrum principles for countries and regions. In 2006, the United Kingdom, Japan and South Korea started, in accordance to the ITU recommendations, to announce the UWB regulatory norms, so as to gradually open its civilian UWB products. There are
currently more than 20 manufacturers developing the UWB chip, application development platforms and related equipment. One of the manufacturers in the United States comes in the forefront, followed by those in Israel, Japan, the United Kingdom, Europe and Chinese Taiwan. According to the prediction of the international authoritative institution, the high-speed UWB chip market will exceed a 400% annual growth rate.

Compared to those aforementioned countries, China's UWB market is making relatively slow progress. Worse, there is currently no domestic enterprise engaged in chip research and development. At present, a special working group has been set up to develop the standards. In September 2009, the working group has supported the Chinese Standard to the National standard Committee. Author has been one of the members in the working group, so can design the system earlier than the others.

1.3.2 Research

In May 2002, IEEE held the first international conference on UWB Science and Technology, i.e., UWBST'02. Since then, the IEEE has held its annual “International Conference on UWB”. By the end of 2002, IEEE authoritative academic journal “Journal of selected Area on Communications” published the first issue on UWB. Currently, several major IEEE international conferences, such as ICC, Globecom, VTC and PIMRC have special sessions on the UWB technology where a large number of research results have been presented.

Since 1999, Chinese researchers have begun to follow the development of the international UWB technology. In 2003 and 2007, the "863" program launched, respectively, projects on the demonstration of UWB systems and on related chip R & D projects. The National Natural Science Foundation has also supported a number of projects on UWB communication systems with a main focus on algorithm and system level research.

In August 2004, the "863" Program held the First Research and Development Seminar on UWB in Shanghai. In 2005, the National Academic Conferences on the UWB Wireless Communications Technology was held in Nanjing. In addition, journals in electronics and communication have published many R&D research results on UWB. Publications of Chinese scholars on UWB have grown rapidly both at home and abroad. But overall, the domestic research in this field still has a large gap from that in those advanced regions/countries. The large gap is mainly found in two aspects,
that’s, 1) low-cost, low-power chip development (integration of base-band and RF); 2) the objective of providing a higher rate of next generation UWB systems.

1.4 Main Focus of This Thesis

In this thesis, the main tasks include:

(1) System architecture design and verification platform. In order to implement the UWB system on chip, the system architecture should be designed in such a manner that the channel estimation, equalization and synchronization are all included. And the design course will reference the guide of the “ITRS2007 design”. [17] The system arithmetic simulation will be first realized using Matlab and C++. The results will show the correctness of the architecture. The algorithm will then be modified to adapt to Verilog HDL, the hardware description language used in this thesis. Through combining the concrete modules, the whole base band of the physical layer for the Chinese standard UWB system will be in place. The performance of the system should reach that of the software simulation. Finally, the back-end simulation based on FPGA will be performed.

(2) Develop the UWB technology to track the status and trends in market. As the UWB technology will finally reach the Chinese market, the CESI Standard has been examined already. Our research will play a great role in Chinese UWB market force.

In summary, the OFDM-based UWB system will be designed and optimized in this thesis through the specificity of UWB channels and bands. It is expected that the DCM demodulation algorithm will not only significantly improve the performance, but also help for the low-cost, low-power Very Large Scale Integrated (VLSI) design and verification. Moreover, the complete system functionality and performance verification will make great contribution to the China UWB standard, which will improve the China standard to the market.
Chapter 2 A Framework for Digital Communications

Modern society obviously depends on electronic communication for much of its functioning. Among the many possible ways of communicating, the class of techniques referred to as digital communications has become predominant in the latter part of the 20th century, and indications are that this trend will continue. There are a number of important conceptual reasons for this development, as well as some related to the advance of technology and to economics, and we will discuss these shortly. First, however, we should gain a working understanding of the digital communications process. [18]

Next, we shall address issues of single-source/single-destination digital communication. A generic model for such a point-to-point digital communication system is shown in Figure 2-1. The givens of the system are the information source, or message generator; the channel, or physical medium by which communication is to take place; and the user, or information sink. These system elements are emphasized in shaded boxes, and are presumed to be the parts of the system over which we have no control. We shall say shortly about the other elements of Figure 2-1, on which we can exert considerable design influence.

2.1 Sources, Channels, and Limits to Communication

The source may inherently be a discrete (or digital) source, such as an alphanumeric keyboard generating a message, or it may produce a sequence of real-valued samples as its message. In either case, elements of the source output sequence will be designated \( W_n \). A third possibility, often the case in practice, is that the source output is an electrical waveform \( W(t) \), continuous in amplitude and time, as in the example of a speech signal produced by a microphone. In any situation, however, the information source is modeled probabilistically, and we will view messages as outputs from some random experiment. [18]

The channel should be broadly understood as a physical mechanism that accepts as input signal, denoted \( S(t) \) in Figure 2-1, and produces output signal, \( R(t) \), which in general is an imperfect rendition of \( S(t) \). Our waveform-level view of the channel attempts to address the true process of the channel, although popular discrete-time, discrete-alphabet models for channels can be derived from the
waveform counterparts.

The corruption of the signal is typically of two forms.

1. The addition of noise by electronic equipment used to perform the communication process, or the influence of external noise processes such as cosmic and atmospheric electromagnetic noise or interfering signals.

2. Channel distortions due to physical channel limitations (e.g., the bandwidth limitations of the voice band telephone channel, or a magnetic tape recorder/player), or due to communication equipment again, such as filters or amplifiers.

In any case, we assume there is a well-defined mathematical model, which includes deterministic and stochastic aspects, for the action of the channel on the input signal.

There is usually some ambiguity over what constitutes the channel and what is
properly part of the other boxes of Figure 2-1, that is, the demodulator, decoder, and others. A perfectly reasonable operational understanding was once made by J.L. Kelly—the channel is that part of the transmission system that we can’t change, or don’t wish to change. For example, concerning channel specification involves optical signaling with a laser. Our best possibilities for efficient design remain when we process the electromagnetic signal directly; however, current technology typically allows that we observe the output of a photo detector, which converts optical photons into electrical current. Such detectors themselves are invariably noisy, exhibiting signal-dependent shot noise and dark currents and often have distorting effects upon the transmitted signals due to response-time limitations. Here then the channel definition could be limited to the electromagnetic medium (perhaps fiber-optic waveguide) or could incorporate a laser diode transmitter and a photo detector as well.

The user, or destination, is to the source we may attach a fidelity criterion that describes goodness of performance. In analog systems, the criterion might be mean-square error between source and destination waveforms, but in discrete alphabet communications, the performance is more traditionally measured by quantities such as symbol error probability or message error probability. All measured by quantities such as symbol error probability or message error probability. [18]

Communication system design gets played out in one of two ways:
(1) We are provided a channel with certain capabilities and wish to design a system that can provide communications at the largest aggregate rate, subject to tolerable distortion constraints.
(2) The traffic load and required fidelity are specified, and we must engineer an efficient channel to accomplish this task. This normally involves designing transmitters, antennas, and receivers to supply a certain signal-to-noise ratio (SNR) and bandwidth and, clearly, we would like to operate with minimal resources required to perform the job.

2.2 Operations in the Digital Transmission Pathway

We now examine the role of the remaining modules in Figure 2-1.

The source encoder’s task is commonly referred to as data compression. Although this description is open to misinterpretation, especially when the original signal is an analog waveform. Basically, the source encoder accepts the source outputs and produces a sequence of symbols, \( U_n \), usually a sequence of bits, that represents
the source output in the best possible way under the constraint of, say, allowing $R_b$ bits per source unit of time.

The source decoder performs a much simpler inverse process, essentially amounting to table lookup. It receives an identification string chosen by the source encoder (assuming no transmission errors) and outputs a message in appropriate form for the user (discrete alphabet characters, real numbers, or waveforms). Traditionally, this involves some form of digital-to-analog conversion and perhaps sample interpolation to produce waveforms.

In a formal sense, the source coding problem can be effectively decoupled from the other operations, notably channel modulation and coding. Fundamental results of information theory show that an efficient communication system, in the sense of Figure 2-1, can be realized as a cascade of the following:

1. An efficient source encoder/decoder, which associates the source output with a discrete message set of source approximations, typically labeled by binary strings.
2. An efficient channel modulation and coding system designed to convey these source coder labels.

Another topic we shall not address here in detail is cryptography, or secrecy coding. Encryption and decryption techniques were formerly relevant only to military and strategic governmental communications, but have lately gained importance in most aspects of telecommunications. Newer cryptographic techniques, called public-key systems, avoid the need for secure key distribution and can, in addition, provide an authentication or electronic signature feature, something digital transmission normally sacrifices.

### 2.3 Channel Coding

The channel encoder is a discrete-input, discrete-output device whose usual purpose is seen as providing some error-correction capability for the system. It does this by using a mapping from input sequences $\{U_n\}$ to code sequences $\{X_n\}$, which inserts redundancy and which utilizes memory. Whereas in N modulator time slots, an un-coded system could transmit $MN$ possible signals, the coded system will enforce constraints that allow a smaller number of coded signals. In this sense, each modulator symbol doesn’t carry as much information as it apparently could, and symbols are in some sense redundant. Memory is the other crucial aspect of good
encoding schemes. In essence, a given message bit at the encoder input influences several, perhaps many, output symbols, hence waveform intervals. This provides a noise-averaging feature, which makes the decoder less vulnerable to the effects of noise, distortion, fading, and the like, occurring in one signaling interval. We will find that this is nothing more than an exploitation of the law of large numbers associated with a random channel mechanism.

An additional role of the channel encoder, although one less commonly attributed to it, may be that of spectral shaping. The memory of the encoder can, if desired, produce an output symbol stream that ultimately shapes the power spectrum of the signal produces by the modulator. An example is the alternate mark inversion technique described previously; a very simple channel encoder remembers the polarity of the previous one symbol and uses the opposite upon receiving the next one. The resultant spectrum has very small power spectral content near zero frequency. Another important example is in coding for magnetic recording channels, where encoding the binary magnetization signal to satisfy run-length constraints helps to increase the information density per unit area of the medium. [19]

2.4 Modulation

A digital modulator is simply a device for converting a discrete-time sequence of symbols from a finite alphabet, whether precoded or not, into continuous-time signals suitable for transmission by the physical channel provided. Likewise, a demodulator is a device for processing a noisy, perhaps distorted, version of the transmitted signal and producing numerical outputs, one per modulator symbol. This series of actions involves both deterministic and stochastic effects.

There are clearly many possible types of modulation processes differing in the manner they manipulate an electromagnetic signal. Such manipulations include changing the amplitude, frequency, or phase angle of a sinusoidal signal, the polarization of the electromagnetic radiation, or the pulse position within a modulation interval. Some descriptions of modulation imply that modulation is only a conversion from a low-frequency (base-band) waveform to a high-frequency (carrier) signal, but we shall adopt a more unified view, that of a signal generator driven by a discrete-time, discrete sequence.

Most classical modulation schemes are memory-less; that is, the contribution to the transmitted waveform induced by a given modulation symbol is defined purely by
that symbol, and not the previous symbols. There are other important schemes, with “modulation” in their adopted name, that have memory. Normally, these are oriented toward spectrum control by introducing constraints on the signal over several intervals. One example is partial-response modulation, in which the transmitted pulse amplitude depends on several previous information bits. In continuous-phase modulation, the phase angle of the transmitted carrier is forced to be continuous at all points in time and often to have continuous derivatives, again effect to a compact power spectrum. [20]

Band-pass modulation can provide other important benefits in signal transmission. If more than one signal utilizes a signal channel, modulation may be used to separate the different signals. Such a technique is known as frequency-division multiplexing. Modulation can be used to minimize the effects of interference. A class of such modulation schemes, known as spread-spectrum modulation, requires a system bandwidth much larger than the minimum bandwidth that would be requires by the message. Modulation can also be used to place a signal in a frequency band where design requirements, such as filtering and amplification, can be easily met. This is the case when RF signals are converted to an intermediate frequency (IF) in a receiver.

2.4.1 Digital Band-pass Modulation Techniques

Band-pass modulation (either analog or digital) is the process by which an information signal is converted to a sinusoidal waveform; for digital modulation, such a sinusoid of duration $T$ is referred to as a digital symbol. The sinusoid has just three features that can be used to distinguish it from other sinusoids: amplitude, frequency, and phase. Thus band-pass modulation can be defined as the process whereby the amplitude, frequency, or phase of an RF carrier, or a combination of them, is carried in accordance with the information to be transmitted. The general form of the carrier wave is

$$s(t) = A(t)\cos \theta(t)$$  \hspace{1cm} (2. 1)

Where, $A(t)$ is the time-varying amplitude and $\theta(t)$ is the time-varying angle. It is convenient to write

$$\theta(t) = \omega_0 t + \phi(t)$$  \hspace{1cm} (2. 2)

So that
\[ s(t) = A(t) \cos[\omega_0 t + \phi(t)] \] (2.3)

Where, \( \omega_0 \) is the radian frequency of the carrier and \( \phi(t) \) is the phase. The terms \( f \) and \( \omega \) will each be used to denote frequency. When \( f \) is used, frequency in hertz is intended; when \( \omega \) is used, frequency in radians per second is intended. The two frequency parameters are related by \( \omega = 2\pi f \).

### 2.4.2 Phase Shift Keying

Phase shift keying (PSK) was developed during the early days of the deep-space program; PSK is now widely used in both military and commercial communication systems. The general analytic expression for PSK is

\[ s_i(t) = \sqrt{\frac{2E}{T}} \cos[\omega_0 t + \phi_i(t)] \quad 0 \leq t \leq T \quad i = 1, ..., M \] (2.4)

Where the phase term, \( \phi_i(t) \), will have \( M \) discrete values, typically given by

\[ \phi_i(t) = \frac{2\pi i}{M} \quad i = 1, ..., M \]
For the binary PSK (BPSK) example in Figure 2-2, M is 2. The parameter E is symbol energy, T is symbol time duration, and $0 \leq t \leq T$. In BPSK modulation, the modulating data signal shifts the phase of the waveform $s(t)$ to one of two states, either zero or $\pi$ (180°). The waveform sketch in Figure 2-2 shows a typical BPSK waveform with its abrupt phase changes at the symbol transitions, if the modulating date stream were to consist of alternating ones and zeros, there would be such an
abrupt change at each transition. The signal waveforms can be represented as vectors on a polar plot; the vector length corresponds to the signal amplitude, and the vector direction for the general M-ary case corresponds to the signal phase relative to the other M-1 signals in the set. For the BPSK example, the vector picture illustrates the two 180° opposing vectors. Signal sets that can be depicted with such opposing vectors are called antipodal signal sets. [20]

2.4.3 Frequency Shift Keying

The general analytic expression for Frequency shift keying (FSK) modulation is

\[
 s_i(t) = \sqrt{\frac{2E}{T}} \cos(\omega t + \phi) \quad 0 \leq t \leq T \quad i = 1, \ldots, M
\]

(2.5)

Where the frequency term \( \omega \) has M discrete values, and the phase term \( \phi \) is an arbitrary constant. The FSK waveform sketch in Figure 2-2 illustrates the typical frequency changes at the symbol transitions. At the symbol transitions, the figure depicts a gentle shift from one frequency (tone) to another. This behavior is only true for a special class of FSK called continuous-phase FSK (CPFSK). In the general MFSK case, the change to a different tone can be quite abrupt, because there is no requirement for the phase to be continuous. In this example, \( M \) has been chosen equal to 3, corresponding to the same number of waveform types (3-ary); note that this \( M=3 \) choice for FSK has been selected to emphasize the mutually perpendicular axes. In practice, \( M \) is usually a nonzero power of 2 (2, 4, 8, and 16). The signal set is characterized by Cartesian coordinates, such that each of the mutually perpendicular axes represents a sinusoid with a different frequency. As described earlier, signal sets that can be characterized with such mutually perpendicular vectors are called orthogonal signals. Not all FSK signaling is orthogonal. For any signal set to be orthogonal, it must meet the criterion

\[
 \int_0^T s_1(t)s_2(t)dt = 0
\]

(2.6)

For an FSK signal set, in the process of meeting this criterion, a condition arises on the spacing between the tones in the set. [20]
2.4.4 Amplitude Shift Keying

For the ASK example in Figure 2-2, the general analytic expression is

\[ s(t) = \sqrt{\frac{2E_i(t)}{T}} \cos(\omega_0 t + \phi) \quad 0 \leq t \leq T \quad i = 1, ..., M \] (2.7)

Where the amplitude term \( \sqrt{\frac{2E_i(t)}{T}} \) will have \( M \) discrete values, and the phase term \( \phi \) is an arbitrary constant. In Figure 2-2, \( M \) has been chosen equal to 2, corresponding to two waveform types. The ASK waveform sketch in the figure can be describe a radar transmission example, where the two signal amplitude states would be \( \sqrt{2E/T} \) and zero. The vector picture utilizes the same phase-amplitude polar coordinates as the PSK example. Here, we see a vector corresponding to the maximum-amplitude state, and a point at the origin corresponding to the zero-amplitude state. Binary ASK signaling (also called on–off keying) was one of the earliest forms of digital modulation used in radio telegraphy at the beginning of this century. Simple ASK is no longer widely used in digital communications systems, and thus it will not be treated in detail here.[20]

2.4.5 Amplitude Phase Keying

For the combination of ASK and PSK (APK) example, in Figure 2-2, the general analytic expression

\[ s(t) = \sqrt{\frac{2E_i(t)}{T}} \cos(\omega_0 t + \phi(t)) \quad 0 \leq t \leq T \quad i = 1, ..., M \] (2.8)

illustrates the indexing of both the signal amplitude term and the phase term. The APK waveform picture in Figure 2-2 illustrates some typical simultaneous phase and amplitude changes at the symbol transition times. For this example, \( M \) has been chosen equal to 8, corresponding to eight waveforms (8-ary). The figure illustrates a hypothetical eight-vector signal set on the phase-amplitude plane. Four of the vectors are at one kind of amplitudes, and the other four vectors are at the other different kind of amplitude. Each of the vectors is separated by \( 45^\circ \). When the set of \( M \) symbols in the two-dimensional signal space are arranged in a rectangular constellation, the
signaling is referred to as Quadrature Amplitude Modulation (QAM).

The vector picture for each of the modulation types described in Figure 2-2 (except the FSK case) is characterized on a plane whose polar coordinates represent signal amplitude and phase.

### 2.4.6 Quadrature Phase Shift Keying (QPSK)

Quadrature phase shift keying (QPSK) has twice the bandwidth efficiency of BPSK, since two bits are transmitted in a single modulation symbol. The phase of the carrier takes on one of four equally spaced values, such as $0, \pi/2, \pi$, and $3\pi/2$, where each value of phase corresponds to a unique pair of message bits. The QPSK signal for this set of symbol states may be defined as

$$s_{QPSK}(t) = \frac{2E_s}{T_s} \cos \left[ 2\pi f_s t + \left( i-1 \right) \frac{\pi}{2} \right] \quad 0 \leq t \leq T_s \quad i = 1, 2, 3, 4$$

(2.9)

In (2.9) $T_s$ is the symbol duration and is equal to twice the bit period.

Using trigonometric identities, the above equations can be rewritten for the interval $0 \leq t \leq T_s$ as

$$s_{QPSK}(t) = \frac{2E_s}{T_s} \cos \left( i-1 \right) \frac{\pi}{2} \cos (2\pi f_s t) - \frac{2E_s}{T_s} \sin \left( i-1 \right) \frac{\pi}{2} \sin (2\pi f_s t)$$

(2.10)

If the basis functions $\phi_1(t) = \sqrt{2/T_s} \cos(2\pi f_s t)$, $\phi_2(t) = \sqrt{2/T_s} \sin(2\pi f_s t)$ are defined over the interval $0 \leq t \leq T_s$ for QPSK signal set, then the four signals in the set can be expressed in terms of the basis signals as

$$s_{QPSK}(t) = \left\{ \sqrt{E_s} \cos \left( i-1 \right) \frac{\pi}{2} \phi_1(t) - \sqrt{E_s} \sin \left( i-1 \right) \frac{\pi}{2} \phi_2(t) \right\} \quad i = 1, 2, 3, 4$$

(2.11)

based on this representation, a QPSK signal can be depicted using a two-dimensional constellation diagram with four points as shown in Figure 2-3. It should be noted that different QPSK signal sets can be derived by simply rotating the constellation. As an example, Figure 2-3 shows another QPSK signal set where the phase values are $\pi/4, 3\pi/4, 5\pi/4, 7\pi/4$.

From the constellation diagram of a QPSK signal, it can be seen that the distance between adjacent points in the constellation is $\sqrt{2E_s}$. Since each symbol corresponds
to two bits, then $E_s = 2E_b$, thus the distance between two neighboring points in the QPSK constellation is equal to $2\sqrt{E_b}$. The average probability of bit error in the additive white Gaussian noise (AWGN) channel is obtained as

$$P_{e,QPSK} = Q\left(\frac{2E_b}{\sqrt{N_0}}\right)$$

(2.12)

A striking result is that the bit error probability of QPSK is identical to BPSK, but twice as much data can be sent in the same bandwidth. Thus when compared to BPSK, QPSK provides twice the spectral efficiency with exactly the same energy efficiency.

![QPSK constellation](image)

Figure 2-3 QPSK constellation
(a) the carrier phase are $0, \pi/2, \pi$ and $3\pi/2$; (b) the carrier phase are $\pi/4, 3\pi/4, 5\pi/4$ and $7\pi/4$.

2.4.7 M-ary Quadrature Amplitude Modulation (QAM)

In $M$-ary PSK modulation, the amplitude of the transmitted signal was constrained to remain constant, thereby yielding a circular constellation. By allowing the amplitude to also vary with the phase, a new modulation scheme called quadrature amplitude modulation (QAM) is obtained. Figure 2-4 shows the constellation diagram of 16-ary QAM. The constellation consists of a square lattice of signal points. The general form of an $M$-ary QAM signal can be defined as

$$s(t) = \sqrt{\frac{2E_{\text{min}}}{T_s}} a_i \cos(2\pi f_i t) + \sqrt{\frac{2E_{\text{min}}}{T_s}} b_i \sin(2\pi f_i t) \quad 0 \leq t \leq T, i = 1, 2, ..., M$$

(2.13)
In (2.13) $E_{\text{min}}$ is the energy of the signal with the lowest amplitude, $a_i$ and $b_i$ are a pair of independent integers chosen according to the location of the particular signal point. Note that $M$-ary QAM does not have constant energy per symbol, nor does it have constant distance between possible symbol states. It reasons that particular values of $s_i(t)$ will be detected with higher probability than others.

If rectangular pulse shapes are assumed, the signal $s(t)$ may be expanded in terms of a pair of basis functions defined as

$$
\phi_1(t) = \sqrt{2/T_s} \cos(2\pi f_s t) \quad 0 \leq t \leq T_s 
$$

$$
\phi_2(t) = \sqrt{2/T_s} \sin(2\pi f_s t) \quad 0 \leq t \leq T_s
$$

![Figure 2-4 Constellation diagram of an M-ary QAM (M=16) signal set](image)

The coordinates of the $i$th message point are $a_i \sqrt{E_{\text{min}}}$ and $b_i \sqrt{E_{\text{min}}}$, where $(a_i, b_i)$ is an element of the $L \times L$ matrix given by

$$
\{a_i, b_i\} = \begin{bmatrix}
(L-1, -L-1) & (L-1, L-1) & \cdots & (L-1-1, L-1) \\
(L-1-1, L-3) & (L-1-1, L-3) & \cdots & (L-1-3, L-3) \\
\vdots & \vdots & \ddots & \vdots \\
(L-1-1, L-1) & (L-1-1, L-1) & \cdots & (L-1-1, L-1)
\end{bmatrix}
$$

In (2.16), $L = \sqrt{M}$. For example, for a 16-QAM with signal constellation as shown in Figure 2-4, the $L \times L$ matrix is
2. A framework for digital communications

\[
\{a_i, b_i\} = \begin{bmatrix}
(-3, 3) & (-1, 3) & (1, 3) & (3, 3) \\
(-3, 1) & (-1, 1) & (1, 1) & (3, 1) \\
(-3, -1) & (-1, -1) & (1, -1) & (3, -1) \\
(-3, -3) & (-1, -3) & (1, -3) & (3, -3)
\end{bmatrix}
\] (2.17)

It can be shown that the average probability of error in an Additive White Gaussian Noise (AWGN) channel for \(M\)-ary QAM, using coherent detection, can be approximated by

\[
P_e \approx 4 \left(1 - \frac{1}{\sqrt{M}}\right) Q\left(\sqrt{\frac{2E_{\text{min}}}{N_0}}\right)
\] (2.18)

In terms of the average signal energy \(E_{\text{av}}\), this can be expressed as

\[
P_e \approx 4 \left(1 - \frac{1}{\sqrt{M}}\right) Q\left(\sqrt{\frac{3E_{\text{av}}}{(M - 1)N_0}}\right)
\] (2.19)

The power spectrum and bandwidth efficiency of QAM modulation is identical to \(M\)-ary PSK modulation. In terms of power efficiency, QAM is superior to \(M\)-ary PSK. Table 2.1 lists the bandwidth and power efficiencies of a QAM signal for various kinds of \(M\), assuming optimum raised cosine roll off filtering in AWGN. As with \(M\)-PSK, the table is optimistic, and actual bit error probabilities for wireless systems must be determined by simulating the various impairments of the channel and the particular receiver implementation. Pilot tones or equalization must be used for QAM in mobile systems. [21]

<table>
<thead>
<tr>
<th>(M)</th>
<th>4</th>
<th>16</th>
<th>64</th>
<th>256</th>
<th>1024</th>
<th>4096</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\eta_B)</td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>4</td>
<td>5</td>
<td>6</td>
</tr>
<tr>
<td>(E_b/N_0) for BER=10^{-6}</td>
<td>10.5</td>
<td>15</td>
<td>18.5</td>
<td>24</td>
<td>28</td>
<td>33.5</td>
</tr>
</tbody>
</table>

### 2.5 Diversity Techniques

Diversity is a powerful communication receiver technique that provides wireless link improvement at relatively low cost. Unlike equalization, diversity requires no training overhead since a training sequence is not required by the transmitter. Furthermore, there are a wide range of diversity implementations, many which are very practical and provide significant link improvement with little added cost.

Diversity exploits the random nature of radio propagation by finding independent
(or at least highly uncorrelated) signal paths for communication. In virtually all applications, diversity decisions are made by the receiver, and are known to the transmitter.

The diversity concept can be explained simply. If one radio path undergoes a deep fade, another independent path may have a strong signal. By having more than one path to select from, both the instantaneous and average SNR at the receiver may be improved, often by as much as 20dB to 30dB.

In the channels, there are two types of fading—small-scale and large-scale fading. Small-scale fades are characterized by deep and rapid amplitude fluctuations which occur as the mobile moves over distances of just a few wavelengths. These fades are caused by multiple reflections from the surroundings in the vicinity of the mobile. For narrowband signals, small-scale fading typically results in a Rayleigh fading distribution of signal strength over small distances. In order to prevent deep fades from occurring, microscopic diversity techniques can exploit the rapidly changing signal. By selecting the best signal at all times, a receiver can mitigate small-scale fading effects (this is called antenna diversity or space diversity).

Large-scale fading is caused by shadowing due to variations in both the terrain profile and the nature of the surroundings. In deeply shadowed conditions, the received signal strength at a mobile can drop well below that of free space. Large-scale fading is always shown to be log-normally distributed with a standard deviation of about 10dB in urban environments. By selecting a base station which is not shadowed when others are, the mobile can improve substantially the average signal-to-noise ratio on the forward link. This is called macroscopic diversity, since the mobile is taking advantage of large separations (the macro-system differences) between the serving base stations.

The reason that we use diversity techniques is that the average SNR in the branch which is selected using diversity naturally increases, since it is always guaranteed to be above the specified threshold. It offers an average improvement in the link margin without requiring additional transmitter power or sophisticated receiver circuitry. The diversity improvement can be directly related to the average bit error rate for various modulations.

Diversity techniques contain selection diversity, maximal ratio combining, equal-gain combining, polarization diversity, frequency diversity, time diversity and so on. Next we will illustrate some of it in detail.
Frequency diversity is implemented by transmitting information on more than one carrier frequency. The rationale behind this technique is that frequencies separated by more than the coherence bandwidth of the channel will be uncorrelated and will thus not experience the same fades. Theoretically, if the channels are uncorrelated, the probability of simultaneous fading will be the product of the individual fading probabilities.

Frequency diversity is often employed in microwave line-of-sight links which carry several channels in a frequency division multiplex mode (FDM). Due to propagation and resulting refraction, deep fading sometimes occurs. In practice, 1: N protection switching is provided by a radio license, wherein one frequency is nominally idle but is available on a stand-by basis to provide frequency diversity switching for any one of the other N carriers (frequencies) being used on the same link, each carrying independent traffic. When diversity is needed, the appropriate traffic is simply switched to the backup frequency. This technique has the disadvantage that is not only requires spare bandwidth but also requires that there be as many receivers as there are channels used for the frequency diversity. However, for critical traffic, the expense may be justified. [21]

New OFDM modulation and access techniques exploit frequency diversity by providing simultaneous modulation signals with error control coding across a large bandwidth, so that if a particular frequency undergoes a fade, the composite signal will still be demodulated.

Time diversity repeatedly transmits information at time spaces that exceed the coherence time of the channel, so that multiple repetitions of the signal will be received with independent fading conditions, thereby providing for diversity. One modern implementation of time diversity involves the use of RAKE receiver for spread spectrum Code Division Multiple Access (CDMA), where the multi-path channel provides redundancy in the transmitted message. By demodulating several replicas of the transmitted CDMA signal, where each replica experiences a particular multi-path delay, the RAKE receiver is able to align the replicas in time so that a better estimate of the original signal may be formed at the receiver.
Chapter 3 The DC-OFDM UWB System

This chapter first illustrates the OFDM technique and analyzes the key point of it, then describes the Dual Carriers (DC)-OFDM system in detail.

3.1 OFDM Technology

OFDM technology is used for the wireless high-speed transmission. As we know, in frequency field, the spectrum of the wireless channel is not flat, while the main idea of the OFDM technique is to divide the given channel into many orthogonal sub-channels, then modulate the sub-channels respectively and transmit the different sub-carries in parallel. In this way, although the general channel is not flat and selective, the sub-channel is relatively flat. The signal transmission on each sub-channel is narrow-band transmission, and the bandwidth of the signal is narrower than the relative bandwidth of the channel, so the signal interference can be eliminated greatly. On the other side, the sub-carries are orthogonal with each other, so their frequency spectrums are overlapped. This can not only eliminate the mutual interference but also enhance the spectrum efficiency. One OFDM symbol is composed of many modulated sub-carriers. Each sub-carrier is modulated by PSK or QAM and so on. And the OFDM symbol can be expressed by formula 3.1

\[ s(t) = \sum_{i=0}^{N-1} d_i \cdot \text{rect} \left( t - t_s - \frac{T}{2} \right) \exp \left[ j 2\pi \left( f_c + \frac{i}{T} \right)(t - t_s) \right] \quad t_s \leq t \leq t_s + T \]

\[ \text{else} \]

In which, \(N\) is number of the channels, \(d_i\) expresses the sub-channel symbol (the symbol modulated by PSK or QAM), \(T\) is the width of the OFDM symbol, \(t_s\) is the start of the symbol and \(f_c\) expresses the first sub-carrier frequency.

\(\text{rect}(t) = 1, |t| \leq T/2\).

The equivalent base-band signal of the OFDM signal is expressed by the following formula 3.2, in which we use \(f_c = 0\).
3. The UWB system based on DC-OFDM techniques

\[ s(t) = \begin{cases} \sum_{i=0}^{N-1} d_i \text{rect}(t - t_i - \frac{T}{2}) \exp \left[ j 2\pi \frac{i}{T}(t - t_i) \right] & t_i \leq t \leq t_i + T \\ 0 & \text{else} \end{cases} \]  

(3.2)

In which, the real and image parts of \( s(t) \) respect to the in-phase and orthogonal parts of the OFDM symbol. When refer to the implementation, we can multiply \( d_i \) with the correspond sub-carriers’ sine and cosine parts. [22]

3.2 DC-OFDM UWB System

Dual Carriers-OFDM (DC-OFDM) UWB system has many similar points with the traditional MB-OFDM UWB system. In this system, the frequency band can be divided into several sub-bands. Each frequency band is less than 264MHz and form one OFDM symbol. It is composed by many orthogonal sub-carriers.

Figure 3-1 is the spectrum of the DC-OFDM UWB system. The 4.2-4.8GHz and 6.0-9.0GHz can be divided into 12 sub-bands. Bands 1-2 belong to the first group while bands 3-12 belong to the second group. DC-OFDM UWB system is based on the OFDM technique, using Time Frequency Code (TFC) technology, and modulates the signals into different frequency carriers at different times, and then to transmit the signals in different sub-bands. At the same time, the instant bandwidth is the dual-carriers bands, that’s dual 264MHz. Table 3.1 lists some parameters of the system.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( N ): Sub-carriers number</td>
<td>128</td>
</tr>
<tr>
<td>( B ): sub-band bandwidth</td>
<td>264MHz</td>
</tr>
<tr>
<td>( \Delta_F ): sub-carrier period</td>
<td>2.0625MHz(= B/N )</td>
</tr>
<tr>
<td>( T_{FFT} ): IFFT/FFT period</td>
<td>484.84ns(=1/\Delta_F )</td>
</tr>
</tbody>
</table>

Figure 3-1 DC-OFDM UWB system frequency spectrum division

Table 3.1 DC-OFDM UWB system main parameters
### 3. The UWB system based on DC-OFDM techniques

<table>
<thead>
<tr>
<th><strong>symbol</strong></th>
<th><strong>description</strong></th>
<th><strong>value</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_{ZP}$</td>
<td>zero padding time</td>
<td>$140.15\text{ns} (=37/B)$</td>
</tr>
<tr>
<td>$T_{G}$</td>
<td>Guard period time</td>
<td>$18.9394\text{ns} (=5/B)$</td>
</tr>
<tr>
<td>$T_{SYM}$</td>
<td>symbol period</td>
<td>$625\text{ns} (=T_{FFT} + T_{ZP})$</td>
</tr>
</tbody>
</table>

Figure 3-2 shows the framework of the physical layer for DC-OFDM UWB system. In the transmitter, the information from the Media Access Control (MAC) layer is processed orderly by scrambler, channel encoder, interleave, QPSK mapping and so on, then go to the IFFT module to generate the OFDM symbol. After that, zero padding and guard period are added into the symbol. The digital signals then turn into analog signals after DAC module and become base-band analog signals. In the RF part, the signal is mixed into high frequency. When in the receiver, the signal is demodulated into low frequency in RF part after passing band-pass filter and Low-Noise Amplifier (LNA), then is sent to ADC to get base-band digital signal. The receiver will process the frame detection and synchronization and then go to FFT module to get the frequency signal. In the frequency field, we can process channel estimation and equalization. After that the signals go to QPSK demodulation, de-interleave module, Viterbi decoder, descrambler, and then output data and return to MAC layer.

**Figure 3-2 DC-OFDM UWB system framework**

### 3.2.1 Physical Layer Frame Configuration

Data frame configuration is shown in Figure 3-3. Each frame contains three parts, firstly goes with the preambles, shown in

**Figure 3-4**, composed of 12 package synchronization symbols, 4 frame
synchronization symbols and 4 channel estimation symbols. Package synchronization symbol is used for package detection, symbol synchronization and coarse frequency offset estimation. Frame synchronize symbol is used to indicate the end of the channel estimation.

After preamble come the header of the frame, which contains physical layer header, Media Access Control layer (MAC) header and Header Check Sequence (HCS). Physical layer header contains information of speed, frame length shown in Table 3.2. The PHY layer first pre-appends the PHY header to the MAC header and then calculates the HCS over the combined headers. Tail bits are then inserted after the PHY header in order to return the convolution encoder to the “zero state”. The resulting HCS is appended to the end of the MAC header along with an additional set of tail bits. Pad bits are finally added to the end of the tail bits, in order to align the data stream on the OFDM symbol interleave boundaries.

Tail bits are also added to the MAC frame body (i.e., the frame payload plus FCS) in order to return the convolution encoder to the “zero state”. Pad bits are added to the end of the tail bits in order to align the data stream on the OFDM symbol interleave boundary. [1]

The Physical Layer Convergence Protocol (PLCP) preamble is sent first, followed by the PLCP header, the frame payload, the FCS, the tail bits, and finally the pad bits. As shown in Figure 3-3, the PLCP header is always sent at the speed of 53.2Mbps. The remainders of the PLCP frame (frame payload, FCS, tail bits, and pad bits) are sent at the desired speed of 53.2, 80, 106.4, 200, 320, 400 or 480Mbps.

![Figure 3-3 DC-OFDM UWB system PLCP frame format](image-url)
3. The UWB system based on DC-OFDM techniques

Figure 3-4 DC-OFDM UWB system standard preamble format

Table 3.2 PHY Header parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Bit length</th>
<th>Value</th>
<th>Signification</th>
</tr>
</thead>
<tbody>
<tr>
<td>RATE</td>
<td>5</td>
<td>00000</td>
<td>The speed of Payload is 53.2Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00001</td>
<td>The speed of Payload is 80Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00010</td>
<td>The speed of Payload is 106.4Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00011</td>
<td>The speed of Payload is 160Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00100</td>
<td>The speed of Payload is 200Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00101</td>
<td>The speed of Payload is 320Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00110</td>
<td>The speed of Payload is 400Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>00111</td>
<td>The speed of Payload is 480Mbps</td>
</tr>
<tr>
<td></td>
<td></td>
<td>01000-11111</td>
<td>Reserved</td>
</tr>
<tr>
<td>LENGTH</td>
<td>12</td>
<td>0~4095</td>
<td>12bit unsigned integer, meaning payload length</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>(not contains FCS), output LSB first.</td>
</tr>
<tr>
<td>Scrambler</td>
<td>2</td>
<td>00,01,10,11</td>
<td>Define initial state of scrambler</td>
</tr>
<tr>
<td>Init</td>
<td></td>
<td></td>
<td>{S1,S2,13\’b1}</td>
</tr>
<tr>
<td>Burst Mode</td>
<td>1</td>
<td>0</td>
<td>Next packet not part of burst</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1</td>
<td>Next packet is part of burst</td>
</tr>
<tr>
<td>Preamble</td>
<td>1</td>
<td>0</td>
<td>Standard preamble</td>
</tr>
<tr>
<td>Type</td>
<td></td>
<td>1</td>
<td>Shortened preamble</td>
</tr>
</tbody>
</table>

3.2.2 Data Scrambler and Descrambler

A side-stream scrambler shall be used for the PLCP header and entire PLCP frame body. In the PLCP header, only the combination of the MAC header, HCS, the tail bits following it and the pad bits shall be scrambled; the PLCP preamble, the PHY header and the tail bit field following the PLCP header shall not be scrambled (the tail
The UWB system based on DC-OFDM techniques

bits following the HCS shall be reset to zero after scrambling). The polynomial generator, \( g(D) \), for the Pseudo-Random Binary Sequence (PRBS) generator shall be \( g(D) = 1 + D^{14} + D^{15} \), where \( D \) is a single bit delay element. The polynomial not only forms a maximal length sequence, but is also a primitive polynomial. Using this generator polynomial, the corresponding PRBS, \( x_m \), is generated as

\[
x_n = x_{n-14} \oplus x_{n-15}, \quad n = 0, 1, 2, \ldots
\]

(3.3)

Where “\( \oplus \)” denotes modulo-2 addition. The following sequence defines the initialization vector, \( x_{\text{init}} \), which is specified by the parameter “seed value” in table 3.3.

The scrambled data bits, \( s_m \), are obtained as follows:

\[
s_m = d_m \oplus x_m, \quad m = 0, 1, 2, \ldots
\]

(3.4)

In (3.4), \( d_m \) represents the unscrambled data bits. The side-stream descrambler at the receiver shall be initialized with the same initialization vector, \( x_{\text{init}} \), used in the transmitter scrambler. The initialization vector is determined from the seed identifier contained in the PLCP header of the received frame.

The 15-bit initialization vector or seed value shall correspond to the seed identifier as shown in Table 3.3. The MAC shall set the seed identifier value to 00 when the PHY is initialized and this value shall be incremented in a 2 bit rollover counter for each frame sent by the PHY. The value of the seed identifier sent in the PLCP header shall be used to initialize the scrambler for both the PLCP header and the frame body.[1]

<table>
<thead>
<tr>
<th>Seed identifier (S1,S2)</th>
<th>Seed value ( x_{\text{init}} = x_1[-1]x_1[-2]...x_1[-15] )</th>
<th>Scrambler output first 16 bits ( x[0],x[1]...x[15] )</th>
</tr>
</thead>
<tbody>
<tr>
<td>00</td>
<td>00111 1111 1111 111</td>
<td>0000 0000 0000 1000</td>
</tr>
<tr>
<td>01</td>
<td>01111 1111 1111 111</td>
<td>0000 0000 0000 0100</td>
</tr>
<tr>
<td>10</td>
<td>10111 1111 1111 111</td>
<td>0000 0000 0000 1110</td>
</tr>
<tr>
<td>11</td>
<td>11111 1111 1111 111</td>
<td>0000 0000 0000 0010</td>
</tr>
</tbody>
</table>

### 3.2.3 Channel Encoder and Viterbi Decoder
The convolution encoder shall use the rate $R=1/3$ code with generator polynomials $g_0 = 133_8$, $g_1 = 165_8$, $g_2 = 171_8$, as shown in Figure 3-5. The various coding rates are derived from the rate $R=1/3$ convolution code by employing “puncturing”. Puncturing is a procedure for omitting some of the encoded bits in the transmitter and inserting a dummy “zero” metric into the convolution decoder on the receive side in place of the omitted bits. The puncturing patterns are illustrated in Figure 3-6 through Figure 3-8. In each of these cases, the tables shall be filled in with encoder output bits from the left to the right. For the last block of bits, the process shall be stopped at the point at which encoder output bits are exhausted, and the puncturing pattern applied to the partially filled block. The rate-dependent parameters are shown in Table 3.4. [1]
3. The UWB system based on DC-OFDM techniques

Figure 3.7 An example of the bit-stealing and bit-insertion procedure (R=5/8)

Figure 3.8 An example of the bit-stealing and bit-insertion procedure (R=3/4)
3. The UWB system based on DC-OFDM techniques

<table>
<thead>
<tr>
<th>Data rate (Mbps)</th>
<th>Modulation</th>
<th>Coding rate (R)</th>
<th>Time Spreading Factor (TSF)</th>
<th>Coded bits per OFDM symbol (N_{CBPS})</th>
<th>Interleave block length ((= 4 \times TSF / N_{CBPS}))</th>
</tr>
</thead>
<tbody>
<tr>
<td>53.2</td>
<td>QPSK</td>
<td>1/3</td>
<td>2</td>
<td>200</td>
<td>400</td>
</tr>
<tr>
<td>80</td>
<td>QPSK</td>
<td>1/2</td>
<td>2</td>
<td>200</td>
<td>400</td>
</tr>
<tr>
<td>106.4</td>
<td>QPSK</td>
<td>1/3</td>
<td>2</td>
<td>400</td>
<td>800</td>
</tr>
<tr>
<td>160</td>
<td>QPSK</td>
<td>1/2</td>
<td>2</td>
<td>400</td>
<td>800</td>
</tr>
<tr>
<td>200</td>
<td>QPSK</td>
<td>5/8</td>
<td>2</td>
<td>400</td>
<td>800</td>
</tr>
<tr>
<td>320</td>
<td>DCM</td>
<td>1/2</td>
<td>1</td>
<td>400</td>
<td>1600</td>
</tr>
<tr>
<td>400</td>
<td>DCM</td>
<td>5/8</td>
<td>1</td>
<td>400</td>
<td>1600</td>
</tr>
<tr>
<td>480</td>
<td>DCM</td>
<td>3/4</td>
<td>1</td>
<td>400</td>
<td>1600</td>
</tr>
</tbody>
</table>

Decoding by the Viterbi algorithm should first insert the dummy “zero” for the transmitter puncture and then process the Viterbi decoder, the dummy is not calculated when measure the depth.

3.2.4 Interleave and De-interleave

The coded and padded bit stream shall be interleaved prior to modulation to provide robustness against burst errors. The bit interleaving operation is performed in three distinct stages, as shown in Figure 3-9:

1. Symbol interleaving, which permutes the bits across 4 consecutive OFDM symbols, enables the PHY to exploit frequency diversity within a band group.

2. Intra-symbol tone interleaving, which permutes the bits across the data sub-carriers within an OFDM symbol, exploits frequency diversity across sub-carriers and provides robustness against narrow-band interferer.

3. Intra-symbol cyclic shifts, which cyclically shift the bits in successive OFDM symbols by deterministic amounts, enables modes that employ time-domain spreading and the Fixed Frequency Interleaving (FFI) modes to better exploit frequency diversity.

The additional parameters needed are listed in Table 3.5 as a function of the data rate.
3. The UWB system based on DC-OFDM techniques

The additional parameters needed by the interleave module are listed in Table 3.5 as a function of the data rate.

**Table 3.5 Parameters for the Interleave module**

<table>
<thead>
<tr>
<th>Data rate (Mb/s)</th>
<th>Time Domain Spreading factor ($N_{TDS}$)</th>
<th>Coded bits/OFDM symbol ($N_{CBPS}$)</th>
<th>Intra-symbol tone interleaving ($N_{int}$)</th>
<th>Intra-symbol cyclic shifts ($N_{cyc}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>53.2</td>
<td>2</td>
<td>200</td>
<td>20</td>
<td>100</td>
</tr>
<tr>
<td>80</td>
<td>2</td>
<td>200</td>
<td>20</td>
<td>100</td>
</tr>
<tr>
<td>106.4</td>
<td>2</td>
<td>400</td>
<td>40</td>
<td>200</td>
</tr>
<tr>
<td>160</td>
<td>2</td>
<td>400</td>
<td>40</td>
<td>200</td>
</tr>
<tr>
<td>200</td>
<td>2</td>
<td>400</td>
<td>40</td>
<td>200</td>
</tr>
<tr>
<td>320</td>
<td>1</td>
<td>400</td>
<td>40</td>
<td>100</td>
</tr>
<tr>
<td>400</td>
<td>1</td>
<td>400</td>
<td>40</td>
<td>100</td>
</tr>
<tr>
<td>480</td>
<td>1</td>
<td>400</td>
<td>40</td>
<td>100</td>
</tr>
</tbody>
</table>

**3.2.5 Sub-Carriers Constellation Mapping**

In the DC-OFDM UWB system, the coded and interleaved binary data sequence will be mapped onto a complex constellation. For data rates of 200 Mb/s and lower, the binary data shall be mapped onto a QPSK constellation. For data rates of 320 Mb/s and higher speed, the binary data shall be mapped onto the multi-dimensional constellation, such as Dual-Carrier Modulation (DCM).

**QPSK**

The coded and interleaved binary serial input data, $b[i]$ where $i = 0, 1, 2, ...$, shall be divided into groups of two bits and converted into a complex number representing one of the four QPSK constellation points. The conversion shall be performed according to the Gray-coded constellation mapping, illustrated in Figure
3-10, with the input bit, \( b[2k] \) where \( k = 0, 1, 2, \ldots \), being the earliest of the two in the stream. The output values, \( d[k] \) where \( k = 0, 1, 2, \ldots \), are formed by multiplying \((2 \times b[2k] - 1) + j(2 \times b[2k+1] - 1)\) value by a normalization factor of \( K_{MOD} \), as described in the following equation:

\[
d[k] = K_{MOD} \times \left[ (2 \times b[2k] - 1) + j(2 \times b[2k+1] - 1) \right], \quad k = 0, 1, 2, \ldots
\]

The normalization factor is \( K_{MOD} = 1/\sqrt{2} \) for a QPSK constellation. An approximate value of the normalization factor may be used, as long as the device conforms to the modulation accuracy requirements. For QPSK, \( b[2k] \) determines the \( I \) value, and \( b[2k+1] \) determines the \( Q \) value, as illustrated in Table 3.6.[1]

![QPSK constellation bit encoding](image)

**Figure 3-10 QPSK constellation bit encoding**

<table>
<thead>
<tr>
<th>Input bits ((b[2k], b[2k+1]))</th>
<th>I-output</th>
<th>Q-output</th>
</tr>
</thead>
<tbody>
<tr>
<td>00</td>
<td>-1</td>
<td>-1</td>
</tr>
<tr>
<td>01</td>
<td>-1</td>
<td>1</td>
</tr>
<tr>
<td>10</td>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td>11</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

**Table 3.6 QPSK Encoding Table**

**DCM**

The coded and interleaved binary serial input data, \( b[i] \) where \( i = 0, 1, 2, \ldots \), shall be divided into groups of 200 bits and converted into 100 complex numbers using a technique called dual-carrier modulation. The conversion shall be performed as follows.

1. The 200 coded bits are grouped into 50 groups of 4 bits. Each group is represented as \((b[g(k)], b[g(k)+1], b[g(k) + 50]), b[g(k) + 51])\), where \( k \in [0, 49] \) and
3. The UWB system based on DC-OFDM techniques

\[ g(k) = \begin{cases} 
2k & k \in [0, 24] \\
2k + 50 & k \in [25, 49]
\end{cases} \quad (3.6) \]

(2) Each group of 4 bits \( (b[g(k)], b[g(k)+1], b[g(k) + 50], b[g(k) + 51]) \) shall be mapped onto a four-dimensional constellation, as shown in Figure 3-11, and converted into two complex numbers \( (d[k], d[k + 50]) \). The mapping between bits and constellation is enumerated in Table 3.7.

(3) The complex numbers shall be normalized using a normalization factor \( K_{MOD} \). The normalization factor \( K_{MOD} = 1/\sqrt{10} \) is used for the dual-carrier modulation. An approximate value of the normalization factor may be used, as long as the device conforms to the modulation accuracy requirements.

![Diagram](a)
3. The UWB system based on DC-OFDM techniques

![DCM Encoding Diagram](image)

**Figure 3-11 DCM encoding.** (a) mapping for $d[k]$. (b) mapping for $d[k+50]$

**Table 3.7 Dual-carrier Modulation Encoding Table**

<table>
<thead>
<tr>
<th>Input bit (b[(g(k)]), b[(g(k)+1)], b[(g(k)+50)], b[(g(k)+51)])</th>
<th>I-output $d[k]$</th>
<th>Q-output $d[k]$</th>
<th>I-output $d[k+50]$</th>
<th>Q-output $d[k+50]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0000</td>
<td>-3</td>
<td>-3</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>0001</td>
<td>-3</td>
<td>-1</td>
<td>1</td>
<td>-3</td>
</tr>
<tr>
<td>0010</td>
<td>-3</td>
<td>1</td>
<td>1</td>
<td>3</td>
</tr>
<tr>
<td>0011</td>
<td>-3</td>
<td>3</td>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td>0100</td>
<td>-1</td>
<td>-3</td>
<td>-3</td>
<td>1</td>
</tr>
<tr>
<td>0101</td>
<td>-1</td>
<td>-1</td>
<td>-3</td>
<td>-3</td>
</tr>
<tr>
<td>0110</td>
<td>-1</td>
<td>1</td>
<td>-3</td>
<td>3</td>
</tr>
<tr>
<td>0111</td>
<td>-1</td>
<td>3</td>
<td>-3</td>
<td>-1</td>
</tr>
<tr>
<td>1000</td>
<td>1</td>
<td>-3</td>
<td>3</td>
<td>-1</td>
</tr>
<tr>
<td>1001</td>
<td>1</td>
<td>-1</td>
<td>3</td>
<td>-3</td>
</tr>
<tr>
<td>1010</td>
<td>1</td>
<td>1</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>1011</td>
<td>1</td>
<td>3</td>
<td>3</td>
<td>-1</td>
</tr>
<tr>
<td>1100</td>
<td>3</td>
<td>-3</td>
<td>-1</td>
<td>1</td>
</tr>
<tr>
<td>1101</td>
<td>3</td>
<td>-1</td>
<td>-1</td>
<td>-3</td>
</tr>
<tr>
<td>1110</td>
<td>3</td>
<td>1</td>
<td>-1</td>
<td>3</td>
</tr>
<tr>
<td>1111</td>
<td>3</td>
<td>3</td>
<td>-1</td>
<td>-1</td>
</tr>
</tbody>
</table>
3. The UWB system based on DC-OFDM techniques

3.2.6 OFDM Modulation and Demodulation

The DC-OFDM UWB system demands the 128-point IFFT to implement the OFDM modulation. Figure 3-12 and Figure 3-13 show IFFT inputs’ relationship with sub-carriers. The inputs of the Figure 3-12 are corresponding to the abscissa indexes of the Figure 3-13. The modulated symbols on the sub-carriers are divided as

(1) NULL symbol

We don’t modulate at the numbers of 0, 62, 64, 65, 66 sub-carriers, of which the 0 sub-carrier is DC, and the 62, 64, 65, 66 sub-carriers are the boundaries of the frequency. [1]

(2) Data symbol

The number of modulated data is 100, and their positions are shown in Table 3.8.

<table>
<thead>
<tr>
<th>n</th>
<th>0</th>
<th>[1,9]</th>
<th>[10,18]</th>
<th>[19,27]</th>
<th>[28,36]</th>
<th>[37,45]</th>
<th>[46,49]</th>
</tr>
</thead>
<tbody>
<tr>
<td>M(n)</td>
<td>n-56</td>
<td>n-55</td>
<td>n-54</td>
<td>n-53</td>
<td>n-52</td>
<td>n-51</td>
<td>n-50</td>
</tr>
<tr>
<td>n</td>
<td>50.53</td>
<td>54.62</td>
<td>63.71</td>
<td>72.80</td>
<td>81.89</td>
<td>90.98</td>
<td>99</td>
</tr>
<tr>
<td>M(n)</td>
<td>n-49</td>
<td>n-48</td>
<td>n-47</td>
<td>n-46</td>
<td>n-45</td>
<td>n-44</td>
<td>n-43</td>
</tr>
</tbody>
</table>

According to different data rate, there are two mapping methods:

- 53.2, 80Mbps

When the payload is coded at rate 53.2Mbps and 80Mbps, the standard demands it with frequency spreading and time spreading together. So the complex data

\[ d_{\text{frame}}[k], \text{ in which } k = 0, 1, 2, \ldots \text{ should be divided into many groups, each group contains } N_D = 100 \text{ complex numbers, and the group should be mapped onto the } n\text{th OFDM symbol group, the } i\text{th ( } i = 1 \text{ or } 2 \text{ ) OFDM symbol and the } l\text{th data carrier, that’s } C_{D,n,i}[l], \text{ and } C_{n,k} \text{ shows that the sub-carrier is in the position of } M(n) \text{ among the } k\text{th OFDM symbol. And the following shows this.}

\[
C_{D,2n,1}[l] = d_{\text{frame}} \left[ \frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + l \right]
\]

(3.7)

\[
C_{D,2n,1}[l + \frac{N_D}{2}] = d_{\text{frame}}^* \left[ \frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + \left( \frac{N_D}{2} - 1 - l \right) \right]
\]

(3.8)

\[
C_{D,2n,2}[l] = p_{\text{spread}}[n] \times d_{\text{frame}} \left[ \frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + \left( \frac{N_D}{2} + l \right) \right]
\]

(3.9)
3. The UWB system based on DC-OFDM techniques

\[ C_{D,2n,2}\left[l + \frac{N_D}{2}\right] = p_{\text{spread}}[n] \times d_{\text{frame}}^{*} \left[\frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + (N_D - 1 - l)\right] \] (3.10)

\[ C_{D,2n+1,1}[l] = -p_{\text{spread}}[n] \times d_{\text{frame}} \left[\frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + \left\{\frac{N_D}{2} + l\right\}\right] \] (3.11)

\[ C_{D,2n+1,1}\left[l + \frac{N_D}{2}\right] = -p_{\text{spread}}[n] \times d_{\text{frame}}^{*} \left[\frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + (N_D - 1 - l)\right] \] (3.12)

\[ C_{D,2n+1,2}[l] = d_{\text{frame}} \left[\frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + l\right] \] (3.13)

\[ C_{D,2n+2,1}\left[l + \frac{N_D}{2}\right] = d_{\text{frame}}^{*} \left[\frac{N_D}{2} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + \left\{\frac{N_D}{2} - 1 - l\right\}\right] \] (3.14)

In which \( p_{\text{spread}}[n] \) is defined in the formula

\[ p_{\text{spread}}[n] = p \mod \left(n - \frac{N_{\text{sync}}}{2} + 4, N_{\text{FFT}} - 1\right) \] (3.15)

\( p[n] \) is a pseudorandom sequence with the length of 127, \( l \in \left[0, \frac{N_D}{2} - 1\right] \), \( n \in \left[\frac{N_{\text{sync}} + N_{\text{hdr}}}{2}, \frac{N_{\text{packet}}}{2} - 1\right] \), \( N_{\text{D}} \) is the number of the data sub-carriers, \( N_{\text{sync}} \) is the symbol number of PLCP preamble, \( N_{\text{hdr}} \) is the symbol number of PLCP header, \( N_{\text{packet}} \) is the symbol number of the packet.

- 106.4, 160, 200Mbps

When the data rate is 106.4Mb/s, 160Mb/s or 200Mb/s, only the time spreading technique is needed. So the complex data \( d_{\text{frame}}[k] (k=0, 1, 2, \ldots) \) should be divided into many groups, and each group contains \( 2 \times N_{\text{D}} = 200 \) complex data. The \( C_{D,n,i}[l] \) means the complex should be mapped onto the \( i \)th sub-carrier of the \( n \)th OFDM symbol groups’ \( i \)th (i=1 or 2)OFDM symbol. The details are as following.

\[ C_{D,2n,1}[l] = d_{\text{frame}} \left[N_{\text{D}} \times (2n - N_{\text{sync}} - N_{\text{hdr}}) + l\right] \] (3.16)

\[ C_{D,2n,2}[l] = d_{\text{frame}} \left[N_{\text{D}} \times (2n + 1 - N_{\text{sync}} - N_{\text{hdr}}) + l\right] \] (3.17)
3. The UWB system based on DC-OFDM techniques

\[ C_{D,2n+1,1}[l] = -d_{frame}^{*} \left[ N_D \times (2n+1-N_{sync} - N_{hdr}) + (N_D - 1 - l) \right] \]  

(3.18)

\[ C_{D,2n+1,2}[l] = d_{frame}^{*} \left[ N_D \times (2n-N_{sync} - N_{hdr}) + (N_D - 1 - l) \right] \]  

(3.19)

\[ l \in [0, N_D - 1], \quad n \in \left[ \frac{N_{sync} + N_{hdr}}{2}, \frac{N_{packet} - 1}{2} \right], \quad N_D \text{ is the number of sub-carriers}, \quad N_{sync} \text{ is the symbol number of PLCP preamble}, \quad N_{hdr} \text{ is the symbol number of the PLCP header}, \quad N_{packet} \text{ is the symbol number of the packet}. \]

- 320, 400, 480Mbps

When the data rate is 320Mb/s, 400Mb/s or 480Mb/s, we wouldn’t use any spreading technique. So, the complex data \( d_{frame}[k] \) \( (k = 0, 1, 2, \ldots) \) should be divided into many groups, and each group contain \( 2 \times N_D = 200 \) complex data. The \( C_{D,n,i}[l] \) means the complex should be mapped onto the \( i \)th sub-carrier of the \( n \)th OFDM symbol groups’ \( i \)th (i=1 or 2) OFDM symbol. The details are as following.

\[ C_{D,n,i}[l] = \begin{cases} 
  d_{frame}[2N_D \times (n-N_{sync} - N_{hdr}) + l] & l \in [0, \frac{N_D}{2} - 1] \\
  d_{frame}[2N_D \times (n-N_{sync} - N_{hdr}) + l - 50] & l \in [\frac{N_D}{2}, N_D - 1] 
\end{cases} \]  

(3.20)

\[ C_{D,n,2}[l] = \begin{cases} 
  d_{frame}[2N_D \times (n-N_{sync} - N_{hdr}) + l + 50] & l \in [0, \frac{N_D}{2} - 1] \\
  d_{frame}[2N_D \times (n-N_{sync} - N_{hdr}) + l + 100] & l \in [\frac{N_D}{2}, N_D - 1] 
\end{cases} \]  

(3.21)

In which, \( l \in [0, N_D - 1], \quad n \in [N_{sync} + N_{hdr}, N_{packet} - 1], \quad N_D \text{ is the number of the data sub-carriers}, \quad N_{sync} \text{ is the symbol number of the PLCP preamble}, \quad N_{hdr} \text{ is the symbol number of the PLCP header}, \quad N_{packet} \text{ is the symbol number of the packet}.

(3) Guard period

Each OFDM symbol begins with the channel estimation sequence. For the first 10 sub-carriers, 5 of them are in the boundary of the frequency band, used for the guard period. When to implementation, the power of the guard period is correlated to the power of the data sub-carriers. And this relationship wouldn’t change in one packet. On the other side, the power level of the guard period should satisfy the bandwidth demands regulated by the national radio administration committee.
10 guard period is positioned in the boundary of the OFDM symbol. That’s the logical frequency subcarriers -61,-60,…,-57 and 57,58,…,61. The data on the guard carriers should be copied from the boundary of the OFDM symbols.

\[
C_{G,n,i}[l] = \begin{cases} 
C_{D,n,i}[l] & l \in \left[0, \frac{N_G}{2} - 1\right] \\
C_{D,n,i}[l + 90] & l \in \left[\frac{N_G}{2}, N_G - 1\right]
\end{cases}
\]  

(3.22)

In which \(C_{G,n,i}[l]\) is the \(l\)th guard period of the \(n\)th OFDM groups’ \(i\)th (\(i=1\) or \(2\)) OFDM symbol. \(n \in [N_{sync}, N_{packet} - 1]\), \(N_{sync}\) is the symbol number of the PLCP preamble, \(N_{packet}\) is the number of the packet.

Note: The guard period can be used to relax the filter demands in analog transmitter and receiver, or to improve the performance.

(4) Pilot

In all the OFDM symbols after the PLCP preamble, 12 sub-carriers are defined as pilots for the correlative detection to overcome the frequency offset and phase noise. These pilots’ positions are the logic frequency-55, -45, -35, -25, -15, 5, 15, 25, 35, 45, 55. The actual relationship between pilot sequences and subcarriers depends on the payload and data rates of the PPDU. In conclusion, the pilot has high relation with the data. And the mapping methods change together with the data rate. [1]

![Figure 3-12 Frequency input and time output of the IFFT](Image)
3. The UWB system based on DC-OFDM techniques

After IFFT, zero padding and guard period are inserted, as figure shown.

![Sub-carriers allocations](image)

Figure 3-13 Sub-carriers allocations

![Time-domain OFDM symbol format](image)

Figure 3-14 Time-domain OFDM symbol format

3.2.7 Synchronization and Channel Estimation

OFDM system demands high for the synchronization. The inaccurate symbol timing, the asynchronous between the carrier frequency and sample frequency will directly affect the performance of the OFDM system.

The synchronization involves 3 aspects.

1. Symbol synchronization

   For the OFDM system takes symbol as unit to transmit, the receiver should find the valid start point before the IFFT, and this is symbol synchronization.

2. Carrier frequency synchronization

   Because of the crystalloid mismatch and the Doppler frequency offset between the transmitter and receiver, the carrier frequency offset and phase offset exists. The effect of the carrier frequency synchronization is to estimate the error, and compensate it.

3. Sample timing synchronization

   Because of the crystalloid mismatch between the transmitter and the receiver, the sample timing synchronization is to estimate the offset and adjust the receiver’s sample clock.

In the high-speed OFDM system, the correlated demodulation is used usually. Because the correlated demodulation depends on channel information, the receiver needs channel estimation. In the single-carrier system, the adapted equalization is generally accepted to resist the multi-path effect, but in the OFDM system, two methods are used more. One is to insert the pilot to estimate the channel so that the received pilot can dynamic track the channel, equalize with the feedback of the
channel information, and eliminate the channel bad effect. The other is to add the training sequence to the data frame, using the training sequence to estimate the channel and equalize. For the UWB system is always used indoor, and the channel is relatively invariable. We needn’t follow the channel diversification. So we choose the estimation method based on the training sequence.

Channel is the indispensable part in the whole wireless communication system. And its character in time and frequency domain will directly affect the structure and performance of the wireless communication system. Using a proper channel model to describe the character is very important for the UWB system design. Currently, there are many channel models to describe the UWB system, and we use the IEEE802.15.3a workgroup’s model. It is modified based on the S-V model. S-V model is the dual-exponential model using two Poisson processes. The processes describe the character of the arriving multi-path, but it depicts the multi-path signal with the Rayleigh distribution using in normal narrow system. Based on the testing, improved S-V model uses logarithm normal distribution to describe the multi-path signal scope. The indoor model of the IEEE802.15.3a work group can not only describe the key characters of the practical channel, but also very sample and easy to use. The model contains the path loss and the multi-path in many kinds of environment, such as the multi-path delay expression, attenuation and so on. These characters are based on the testing data. The UWB indoor channel model of the IEEE802.15.3a work group defines four kinds of typical channel environments.

1. CM1, Direct Ocular Path (LOS), the distance between transmitter and receiver is less than 4m.
2. CM2, None Direct Ocular Path (NLOS), the distance between transmitter and receiver is less than 4m.
3. CM3, None Direct Ocular Path (NLOS), the distance between transmitter and receiver is 4~10m.
4. CM4, The situation with serious delay and dispersive signal, the delay expression is 25ns.

Figure 3-15 is the example based on the four channel environments described in the front, and the result is composed by 100 channels. [22]
3. The UWB system based on DC-OFDM techniques

3.3 System Simulation

After building the system, we should simulate the system. We have described the transmitters in detail, and the most important thing is the receiver algorithm. In order to analyze the practical results, we must consider the hardware conditions and the different channels. In our system simulation, we consider to quantize the signal in each sub modules which is good for the hardware verification. We simulate the system in 6 bit quantization and also simulate the system in different channels and different rates. For example, the different SFO, CFO and the ADC fixed points would bring us new challenge to design the system algorithm. Figure 3-16 shows some results of the simulations. That’s under CM1-CM3 environments with the data rate changing from 480Mbps to 53.2Mbps, the figure shows that the performance changes with the SNR and the SFO, CFO.
3. The UWB system based on DC-OFDM techniques

![System performance in different channel environments and rates](image)

Figure 3-16 System performance in different channel environments and rates
Chapter 4 DCM Demodulation Algorithm

4.1 DCM Character Analysis in UWB System

The essence of the DCM modulation method, as described in DC-OFDM UWB standard, is to generate two different 16-QAM symbols from a set of four (coded and interleaved) bits to be transmitted. This is done in two steps which we would describe below. Denote the bits transmitted by \(b_0, b_1, b_2, b_3\). The first step involves generating a set of bipolar symbols \(X_0, X_1, X_2, X_3\), corresponding to the bits \(b_0, b_1, b_2, b_3\) respectively. The mapping between the bits and the bipolar symbols is given by \(x_i = -(\text{\(b_i\)})\) for \(i=0, 1, 2, 3\). Next, two 16-QAM symbols \(y_0\) and \(y_1\) are generated from the bipolar symbols \(X_0, X_1, X_2, X_3\) as follows:

\[
\begin{bmatrix}
  y_0 \\
  y_1 \\
\end{bmatrix} = \begin{bmatrix}
  2 & 1 \\
  1 & -2 \\
\end{bmatrix} \begin{bmatrix}
  x_0 + jx_2 \\
  x_1 + jx_3 \\
\end{bmatrix}
\]

(4.1)

The mapping between \(b_0b_1b_2b_3\) and \(y_0, y_1\) is shown in Table 4.1 where \(R()\) and \(I()\) refer to the real and imaginary parts respectively. The Multi-band OFDM Alliance (MBOA) standard regulates the normalization of the constellation symbols by factor of \(1/\sqrt{10}\). Since that is not crucial to our discussion, we ignore the normalization. In the DC-OFDM standard, there are 100 data sub-carriers. The two 16-QAM symbols, generated as in (4.1), are placed on sub-carriers whose indices are separated by 50. For instance, \(y_0\) could be used to modulate sub-carrier 0 and \(y_1\) to modulate sub-carrier 50, and so on.

<table>
<thead>
<tr>
<th>(b_0b_1b_2b_3)</th>
<th>(R(y_0))</th>
<th>(I(y_0))</th>
<th>(R(y_1))</th>
<th>(I(y_1))</th>
</tr>
</thead>
<tbody>
<tr>
<td>0000</td>
<td>-3</td>
<td>-3</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>0101</td>
<td>-1</td>
<td>-1</td>
<td>-3</td>
<td>-3</td>
</tr>
<tr>
<td>1010</td>
<td>1</td>
<td>1</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>1111</td>
<td>3</td>
<td>3</td>
<td>-1</td>
<td>-1</td>
</tr>
</tbody>
</table>

Table 4.1 The mapping between the bits and the complex symbols.
4. DCM Demodulation Algorithm

4.2 The General DCM Demodulation Algorithm

4.2.1 Practical Algorithms for Soft-demapping of DCM Symbols

(1) Signal Model

The symbols $y_0$ and $y_1$, generated as shown in (4.1), are transmitted over independent Rayleigh flat fading channels, as mentioned above. Assume the received symbols $r_0$ and $r_1$ are corresponded to $y_0$ and $y_1$, respectively. And we have

$$r_i = h_i y_i + n_i \quad i = 0, 1$$  \hspace{1cm} (4.2)

In (4.2), $h_0$ and $h_1$ are the multiplicative channel coefficients; $n_0$ and $n_1$ represent the additive noise. We assume that $h_0$ and $h_1$ are independent zero mean complex Gaussian random variables. And $n_0$, $n_1$ are also zero mean complex Gaussian random variables with variance $\sigma^2$. Note that in the context of an OFDM system, the symbols $r_0$ and $r_1$ represent the received symbols in the frequency domain, assuming no inter-symbol and inter-carrier interference. At the receiver, the following decision variables are formed.

$$z_i = \hat{h}_i r_i = \hat{h}_i h_i y_i + \hat{h}_i n_i \quad i = 0, 1$$  \hspace{1cm} (4.3)

Where $\hat{h}_i$ represents the channel estimates. The decision variables $z_0$ and $z_1$ represent equalized frequency domain received values in a typical OFDM receiver chain.

For the purpose of presenting the expression for the LLR values, we assume that the channel estimation is perfect, and hence, $\hat{h}_i = h_i$. Substituting into (4.3), we obtain

$$z_i = h_i r_i = |h_i| y_i + h_i n_i = |h_i| y_i + n_i' \quad i = 0, 1$$  \hspace{1cm} (4.4)

Where $n_i'$ is a complex Gaussian random variable with variance $|h_i|^2 \sigma^2$. From this point onwards, we shall use the notation $H_i$ for $i=0,1$ to represent $|h_i|^2$. [23]

(2) Exploiting the structure of the DCM mapping

Before proceeding on to solving the demapping problem stated above, we make two simplifying observations about the structure of the DCM mapping.

First of all, observe that $R(y_0)$ and $R(y_1)$ depend only on bits $b_0$ and $b_1$, where $I(y_0)$ and $I(y_1)$ depend only on $b_2$ and $b_3$. Since the real and the imaginary parts represent orthogonal dimensions (I and Q channels), the noise on these two parts is
independent. As such, we make the following observation.

Observation (a) The computation of \( \hat{b}_0 \) and \( \hat{b}_1 \) depends only on \( R(z_0) \) and \( R(z_1) \) and the computation of \( \hat{b}_2 \) and \( \hat{b}_3 \) depends only on \( I(z_0) \) and \( I(z_1) \).

Next, notice that the relation between bits \( b_0, b_1 \), and \( R(y_0) \) is exactly the same as the relation between bits \( b_2, b_3 \) and \( I(y_0) \). Likewise, the relation between bits \( b_0, b_1 \), and \( R(y_1) \) is exactly the same as the relation between bits \( b_2, b_3 \) and \( I(y_1) \) (see Table 4.1). This leads us to the following observation.

Observation (b) The processing required to obtain \( \hat{b}_0 \) and \( \hat{b}_1 \) from \( R(z_0) \) and \( R(z_1) \) is exactly the same as that required to obtain \( \hat{b}_2 \) and \( \hat{b}_3 \) from \( I(z_0) \) and \( I(z_1) \).

Define the log-likelihood ratios

To proceed with the demapping, we define log-likelihood ratios (LLR) for the different bits in terms of the joint probability density functions of \( R(z_0), R(z_1) \) and \( I(z_0), I(z_1) \). Specifically, in light of Observation 1, we have

\[
\Lambda_i(z_0, z_1, H_0, H_1) = \begin{cases} 
\log \frac{p(R(z_0), R(z_i)|b_i = 1)}{p(R(z_0), R(z_i)|b_i = 0)} & i = 0,1 \\
\log \frac{p(I(z_0), I(z_i)|b_i = 1)}{p(I(z_0), I(z_i)|b_i = 0)} & i = 2,3
\end{cases}
\]  

(4.5)

Where \( \Lambda_i(z_0, z_1, H_0, H_1) \) the LLR for bit \( i \), is \( p(R(z_0), R(z_i)|b_i = 1) \), for example, is the joint Probability Density Function (PDF) of \( R(z_0), R(z_1) \) conditioned on the bit \( b_i \) being 1.

Furthermore, from Observation 2 we note that the algebraic expression \( \Lambda_2(z_0, z_1, H_0, H_1) \) AND \( \Lambda_3(z_0, z_1, H_0, H_1) \) can be obtained from the algebraic expressions for \( \Lambda_0(z_0, z_1, H_0, H_1) \) and \( \Lambda_1(z_0, z_1, H_0, H_1) \) respectively, by changing \( R(z_0) \) to \( I(z_0) \) and \( R(z_1) \) to \( I(z_1) \). [23]

The decision rule

After computing the \( \Lambda_i(z_0, z_1, H_0, H_1) \) values, we apply the following decision rule
\[ \hat{b}_i = \begin{cases} 1 & \Lambda_i(z_0, z_1, H_0, H_1) > 0 \\ 0 & \Lambda_i(z_0, z_1, H_0, H_1) \leq 0 \end{cases} \] \quad (4.6)

The \( \Lambda_i(z_0, z_1, H_0, H_1) \) values can also be used as the inputs to the channel code decoder for soft-decision decoding. [23]

(5) **A pictorial representation**

For exposition sake, let us look at the aforementioned demapping problem pictorially. In Figure 4-1, we plot the transmitted data points on a constellation symbol. These points (every point is a couple \( R(y0), R(y1) \)) are denoted by \( X \) and the bits corresponding to them (\( b0, b1 \)) are marked next to the constellation points. The demapping problem is that of determining the LLR values, given that an arbitrary point \( R(z0), R(z1) \) (denoted as \( R \) in Figure 4-1) has been received. [23]

![Figure 4-1 A pictorial representation of DCM constellation.](image)

Note that the \( x \) and \( y \) axes are \( R(z0) \) and \( R(z1) \) respectively. The points marked with \( X \) are the transmitted constellation points. The point \( R \) is the received constellation point.

4.2.2 **CSI-aided Demapping of DCM**

With the DCM principle described in front, we can easily get the input and output relation as the following equation shown.
DCM Demodulation Algorithm

\[
\begin{align*}
\left( \begin{array}{c}
s[k] \\
s[k+50]
\end{array} \right) &= \left( \begin{array}{c}
2 \\
1
\end{array} \right) \left( \begin{array}{c}
b[2k] + j \cdot b[2k + 50] \\
b[2k + 1] + j \cdot b[2k + 51]
\end{array} \right) \\
\left( \begin{array}{c}
s[k+25] \\
s[k+75]
\end{array} \right) &= \left( \begin{array}{c}
2 \\
1
\end{array} \right) \left( \begin{array}{c}
b[2k+100] + j \cdot b[2k + 150] \\
b[2k + 101] + j \cdot b[2k + 151]
\end{array} \right)
\end{align*}
\]

(4.7)

The second mapping rule for \(s[k+25], s[k+75]\) is identical to the first one for \(s[k], s[k+50]\) and thus we only focus on \(s[k], s[k+50]\) from now on. From (4.7), it is easily found that every coded and interleaved bit is transmitted over two different sub-carriers, not a single sub-carrier (QPSK). This is capable of providing frequency diversity, as will be explained more clearly later.

Assuming that OFDM perfectly converts frequency selective channels into parallel frequency flat channels, the received signal \(y[k]\) is given as

\[
y[k] = H[k] \cdot s[k] + w[k], \quad k = 0, \ldots, 99
\]

(4.8)

In (4.8), \(H[k]\) and \(w[k]\) represent the channel gain and the noise signal at the \(k\)-th sub-carrier, respectively. Both are modeled as zero-mean complex Gaussian random variables with \(E\{ |H[k]|^2 \} = 1\) and \(E\{ |w[k]|^2 \} = N_0\). The problem is how to calculate soft decision values for \(b[i]\) with given \(y[k]\) and \(H[k]\).

From (4.7), it is easily found that the real parts of \(s[k]\) and \(s[k+50]\) depend on \(b[2k]\) and \(b[2k+1]\), whereas the imaginary parts depend on \(b[2k+50]\) and \(b[2k+51]\). Therefore, it is not surprising that the LLR values for \(b[2k]\) and \(b[2k+1]\) depend on only the real parts of \(\tilde{s}[k]\) and \(\tilde{s}[k+50]\) as

\[
\hat{d}[2k] = \log \sum_{i \in \{1, -1\}} p(\tilde{s}_a[k], \tilde{s}_a[k+50] | s_a[k], s_a[k+50]) - \log \sum_{i \in \{1, -1\}} p(\tilde{s}_a[k], \tilde{s}_a[k+50] | s_a[k], s_a[k+50])
\]

(4.9)

\[
\hat{d}[2k+1] = \log \sum_{i \in \{1, -1\}} p(\tilde{s}_a[k], \tilde{s}_a[k+50] | s_a[k], s_a[k+50]) - \log \sum_{i \in \{1, -1\}} p(\tilde{s}_a[k], \tilde{s}_a[k+50] | s_a[k], s_a[k+50])
\]

(4.10)

Where \(\tilde{s}[k]\) is defined as the conventional ZF estimate. The summation in (4.9) and (4.10) are performed with respect to all possible values of \((b[2k], b[2k+1])\) that satisfy a specific condition: the first summation is performed over \(((1,0), (1,1))\), while the second one is over \(((0,0), (0,1))\). Here the conditional probability density function (PDF) in (4.9) and (4.10) is given as

\[
p(\tilde{s}_a[k], \tilde{s}_a[k+50] | s_a[k], s_a[k+50]) = (\pi N_0)^{-1/2} \exp \left( -\left| H[k] \cdot (\tilde{s}_a[k] - s_a[k]) \right|^2 / N_0 \right) \times \exp \left( -\left| H[k+50] \cdot (\tilde{s}_a[k+50] - s_a[k+50]) \right|^2 / N_0 \right)
\]

(4.11)

Thus, the exact calculation of LLR requires the calculation of log function. By applying the log-max approximation to (4.9) and (4.10) and simply omitting the
common factors, the approximate LLR can be expressed as
\[ \tilde{b}[2k] = \min \left( d^2(0,0), d^2(0,1) \right) - \min \left( d^2(1,0), d^2(1,1) \right) \] (4.12)
\[ \tilde{b}[2k+1] = \min \left( d^2(0,0), d^2(1,0) \right) - \min \left( d^2(0,1), d^2(1,1) \right) \] (4.13)
In (4.13) \( d^2(0,0), d^2(0,1), d^2(1,0), d^2(1,1) \) are defined as
\[ d^2(0,0) = f(\tilde{s}_0[k] \tilde{s}_0[k+50], H[k], H[k+50]) |_{s_0[k], s_0[k+50] = (-3,1)} \] (4.14)
\[ d^2(0,1) = f(\tilde{s}_0[k] \tilde{s}_0[k+50], H[k], H[k+50]) |_{s_0[k], s_0[k+50] = (-1,-3)} \] (4.15)
\[ d^2(1,0) = f(\tilde{s}_0[k] \tilde{s}_0[k+50], H[k], H[k+50]) |_{s_0[k], s_0[k+50] = (1,3)} \] (4.16)
\[ d^2(1,1) = f(\tilde{s}_0[k] \tilde{s}_0[k+50], H[k], H[k+50]) |_{s_0[k], s_0[k+50] = (3,-1)} \] (4.17)

And \( f(\tilde{s}_0, \tilde{s}_1, H_0, H_1) \) is defined as
\[ f(\tilde{s}_0, \tilde{s}_1, H_0, H_1) = \left( |H_0| \cdot \tilde{s}_0 - |H_0| \cdot s_0[k] \right)^2 + \left( |H_1| \cdot \tilde{s}_1 - |H_1| \cdot s_0[k+50] \right)^2 \] (4.18)

From (4.12) and (4.13), it is easily found that the ZF-based linear demapping followed by weighting by squared channel gains no more calculates the approximate LLR, since every soft decision value is related to two different sub-carriers, not a single sub-carriers (PSK or QAM). Recalling the bit-to-symbol mapping in (4.7), it is readily understood that \( d^2(0,0), d^2(0,1), d^2(1,0), d^2(1,1) \) are the squared distances of \( (|H[k]|, \tilde{s}_0[k], |H[k+50]|, \tilde{s}_0[k+50]) \) from the four points, assuming that \(|H[k]| = 2, |H[k+50]| = 1\). Consequently, the soft decision values in (4.12) and (4.13) are given as the differences of two squared distances and they can be calculated based on \( d^2(0,0), d^2(0,1), d^2(1,0), d^2(1,1) \). Although the straightforward implementation is possible, a more efficient way of calculating (4.12) and (4.13) is proposed in the following. [24]
Figure 4-2 Calculation and selection of soft decision values

Figure 4-2 shows that the two-dimensional space of \((|H[k]|\cdot \bar{s}_k[k],|H[k+50]|\cdot \bar{s}_{k+[50]}\) can be divided into several regions (9 regions in the figure) where the soft decision values are represented by an identical closed-form expression. Note that \(\bar{b}[2k]\) is selected based on two boundaries represented by \(d^2(0,0) - d^2(0,1) = 0\) and \(d^2(1,0) - d^2(1,1) = 0\), while \(\bar{b}[2k+1]\) is on two boundaries represented by \(d^2(0,0) - d^2(1,0) = 0\) and \(d^2(0,1) - d^2(1,1) = 0\).

The selection of soft decision values is summarized in Table 4.2. It is important to note that the soft decision values depend on only these 6 differences of squared distances, \(d^2(0,0) - d^2(0,1), d^2(1,0) - d^2(1,1), d^2(0,0) - d^2(1,0), d^2(0,1) - d^2(1,0), d^2(0,1) - d^2(1,1), d^2(0,0) - d^2(1,1)\), not the squared distances themselves. Consequently, the problem is how to calculate these 6 values, which immediately determines the hardware complexity of the proposed demapping method. [24]

<table>
<thead>
<tr>
<th>(d^2(0,0) - d^2(0,1))</th>
<th>(d^2(1,0) - d^2(1,1))</th>
<th>(\bar{b}[2k])</th>
</tr>
</thead>
<tbody>
<tr>
<td>-</td>
<td>-</td>
<td>(d^2(0,0) - d^2(1,0))</td>
</tr>
<tr>
<td>-</td>
<td>+</td>
<td>(d^2(0,0) - d^2(1,1))</td>
</tr>
</tbody>
</table>
### 4. DCM Demodulation Algorithm

<table>
<thead>
<tr>
<th>$+$</th>
<th>$-$</th>
<th>$d^2(0,1) - d^2(1,0)$</th>
</tr>
</thead>
</table>

| $+$ | $+$ | $d^2(0,1) - d^2(1,1)$ |

<table>
<thead>
<tr>
<th>$d^2(0,0) - d^2(1,0)$</th>
<th>$d^2(0,1) - d^2(1,1)$</th>
<th>$\tilde{b}[2k + 1]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$-$</td>
<td>$-$</td>
<td>$d^2(0,0) - d^2(0,1)$</td>
</tr>
<tr>
<td>$-$</td>
<td>$+$</td>
<td>$d^2(0,0) - d^2(1,1)$</td>
</tr>
<tr>
<td>$+$</td>
<td>$-$</td>
<td>$d^2(1,0) - d^2(0,1)$</td>
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<tr>
<td>$+$</td>
<td>$+$</td>
<td>$d^2(1,0) - d^2(1,1)$</td>
</tr>
</tbody>
</table>

Fortunately, the differences of squared distances can be calculated easily. To be specific, they can be calculated from three common factors (referred to as elementary signals in this paper) as

\[
d^2(0,0) - d^2(1,0) = x_k[k] - 2x_k[k + 50] + p[k]
\]

\[
d^2(1,0) - d^2(1,1) = x_k[k] - 2x_k[k + 50] - p[k]
\]

\[
d^2(0,0) - d^2(1,0) = 2x_k[k] + x_k[k + 50] + p[k]
\]

\[
d^2(0,1) - d^2(1,0) = x_k[k] + 3x_k[k + 50]
\]

\[
d^2(0,1) - d^2(1,1) = 2x_k[k] + x_k[k + 50] - p[k]
\]

\[
d^2(0,0) - d^2(1,1) = 3x_k[k] - x_k[k + 50]
\]

where $x_k[k]$ and $x_j[k]$ denotes the real and imaginary parts of $H^*[k]y[k]$, respectively, and $p[k]$ is defined as $p[k] = 2\left(\|H[k]\|^2 - \|H[k + 50]\|^2\right)$.

Notice that the differences of distances can be calculated by multiplications of $x_k[k]$ (or $x_j[k]$) by pre-determined integers followed by addition to (or subtraction by) $p[k]$.

Taking a closer look at (8) ~ (14), the conventional (ZF-based) linear demapping is nothing but a special case of the proposed demapping where $|H[k]| = |H[k + 50]|$.

The calculation of LLR values for $b[2k+50]$ and $b[2k+51]$ is quite similar, thus detailed explanation is not presented here, because of space limitation. [24]
4. DCM Demodulation Algorithm

4.2.3 A Technique for Demapping DCM Signals with Improved Performance

As shown in the chapter 2, to provide the intra-OFDM-symbol frequency diversity, the DCM modulation technique divides the coded and interleaved binary serial input data, \( b[i] \) where \( i = 0,1,2, \ldots \), into groups of \( 2N \) bits and converts them into \( N \) complex symbol values in a unique method (\( N = 100 \)). The conversion consists of two procedures. First, the \( 2N \) coded and interleaved bits are grouped into \( \frac{N}{2} \) groups of 4 bits. The group is represented as \( \{ b[g(k)], b[g(k)+1], b[g(k)+N/2], b[g(k)+N/2+1] \} \). Next, each group of 4 bits \( \{ b[g(k)], b[g(k)+1], b[g(k)+N/2], b[g(k)+N/2+1] \} \) shall be mapped onto a four-dimensional constellation, and converted into two complex numbers \( \{ Y_T(k), Y_T(k+N/2) \} \). After applying a normalization factor, \( KMOD = \frac{1}{\sqrt{10}} \), the block of complex symbols \( \{ Y_T(k) \} \) shall then form the input to the OFDM modulation block. Actually, the DCM constellation mapping can be obtained by mapping the bipolar symbols \( X_T(i) \) to complex symbols \( Y_T(k) \) as

\[
\begin{bmatrix}
Y_T\left(n + \frac{mN}{4}\right) \\
Y_T\left(n + \frac{mN}{4} + \frac{N}{2}\right)
\end{bmatrix}
= \frac{1}{\sqrt{10}} \begin{bmatrix} 2 & 1 \\ 1 & -1 \end{bmatrix}
\begin{bmatrix}
X_T(2n + mN) + jX_T(2n + mN + \frac{N}{2}) \\
X_T(2n + mN + 1) + jX_T(2n + mN + \frac{N}{2} + 1)
\end{bmatrix}
\]

In (4.26) \( n = 0 \) to \( N/4-1 \) and \( m = 0 \) to 1. Correspondingly, the DCM demapping at the receiver end shall be performed as

\[
\begin{bmatrix}
X_R(2n + mN) + jX_R(2n + mN + \frac{N}{2}) \\
X_R(2n + mN + 1) + jX_R(2n + mN + \frac{N}{2} + 1)
\end{bmatrix}
= \sqrt{10} \begin{bmatrix} \text{Re}\{U\} + j\text{Im}\{U\} \\ \text{Re}\{V\} + j\text{Im}\{V\} \end{bmatrix}
\]

In (4.27), \( U = 2\tilde{Y}_R\left(n + mN/4\right) + \tilde{Y}_R\left(n + mN/4 + N/2\right) \) and \( V = \tilde{Y}_R\left(n + mN/4\right) - 2\tilde{Y}_R\left(n + mN/4 + N/2\right) \), \( n = 0 \) to \( N/4-1 \) and \( m = 0 \) to 1.

It can be seen from (4.26) that the real and imaginary parts of \( \{ Y_T(k) \} \) are valued at \( \frac{1}{\sqrt{10}}, -\frac{1}{\sqrt{10}}, \frac{3}{\sqrt{10}} \) or \( -\frac{3}{\sqrt{10}} \). As a result, the DCM constellation demapping at the receiver end is sensitive to the magnitude variation of the real or imaginary part of \( Y_R(k) \). This is different from QPSK. Moreover, although the constellation mapping shown in standard looks similar to that of a 16-QAM modulation, it should be pointed out that they are also different as, in the case of
DCM, each demapped value, \( X_R(i) \), is found to be related to not a single sub-carrier (which is true in the case of 16-QAM) but two different sub-carriers, i.e., as a function of \( Y_R(k) \) and \( Y_R(k+N/2) \). Obviously, here, one can not directly follow 
\[
\hat{Y}_R(k) = Y_R(k) \left| \hat{H}(k) \right|^2
\]
to weight \( Y_R(k) \) by \( \left| \hat{H}(k) \right| \) and \( Y_R(k+N/2) \) by \( \left| \hat{H}(k+N/2) \right| \) independently. If one of \( \left| \hat{H}(k) \right| \) and \( \left| \hat{H}(k+N/2) \right| \) is large and the other is very small (deep fading on that sub-carrier), for instance, the direct use of 
\[
\hat{Y}_R(k) = Y_R(k) \left| \hat{H}(k) \right|^2
\]
may not only change the magnitude of the demapped value, \( X_R(i) \), but also mistakenly reverse its sign. This, of course, is not desirable for the following soft decision decoding in this case. [25]

Based on the above observations, a new method for weighting the input of DCM demapped is proposed as follows,
\[
\hat{Y}_R(k) = Y_R(k) \cdot \rho(k) \cdot \lambda(k) \quad k = 0, \ldots, N-1 \tag{4.28}
\]

Here, \( \rho(k) = \min(\hat{H}(k), \sigma) \) (\( \sigma \) is an empirical constant) and \( \lambda(k) = \lambda(i+N/2) = \min(\hat{H}(i), \left| \hat{H}(i+N/2) \right|) \) for \( i = 0 \to N/2-1 \). The idea is to first apply the “Y-domain” weighting to a limited extent and then apply the “X-domain” weighting to counteract the high degree of noise on deeply faded sub-carriers. By using this two-dimensional tuning, better soft-decision decoding performance can be achieved. To further explain and examine the effectiveness of the proposed method, we consider the following three typical scenarios.

Case (1) describes that both \( \left| \hat{H}(k) \right| \) and \( \left| \hat{H}(k+N/2) \right| \) are very small. Without loss of generality, \( \left| \hat{H}(k) \right| < \left| \hat{H}(k+N/2) \right| \) is assumed. Following (4.28), in this case, the weighting factors on \( Y_R(k) \) and \( Y_R(k+N/2) \) become \( \left| \hat{H}(k) \right|^2 \) and \( \left| \hat{H}(k) \right| \cdot \left| \hat{H}(k+N/2) \right| \), respectively. Since the difference between these two weighting factors is within a small range, i.e., \( \left| \hat{H}(k) \right|^2 = \left| \hat{H}(k) \right| \cdot \left| \hat{H}(k+N/2) \right| \), the two-dimensional tuning converges to one-dimensional weighting.

Case (2) describes that both \( \left| \hat{H}(k) \right| \) and \( \left| \hat{H}(k+N/2) \right| \) are large. Without loss of generality, \( \left| \hat{H}(k) \right| < \left| \hat{H}(k+N/2) \right| \) is assumed. In this case, the weighting factors on \( Y_R(k) \) and \( Y_R(k+N/2) \) is found to be same with \( \sigma \left| \hat{H}(k) \right| \). Again, the two-dimensional tuning converges to one-dimensional weighting and improved CSI-aided soft decision demapping can also be expected in this scenario.

Case (3) describes that one of \( \left| \hat{H}(k) \right| \) and \( \left| \hat{H}(k+N/2) \right| \) is large and the other is very small. Without loss of generality, we assume \( \left| \hat{H}(k) \right| \) is very small and \( \left| \hat{H}(k+N/2) \right| \) is large. Following (5), in this case, the weighting factors on \( Y_R(k) \) and \( Y_R(k+N/2) \) become \( \left| \hat{H}(k) \right|^2 \) and \( \sigma \left| \hat{H}(k) \right| \), respectively. Here, we have
\[ |\tilde{H}(k)|^2 < \sigma |\tilde{H}(k)| < |\tilde{H}(k+N/2)| < |\tilde{H}(k+N/4)|^2 \] (4.29)

The selection of weighting factor on \( Y_k(k+N/2) \) greater than \( |\tilde{H}(k)|^2 \) has taken advantage of the fact that the sub-carrier \( k+N/2 \) is of higher reliability in this case. Meanwhile, the selection of it to be less than \( |\tilde{H}(k+N/2)|^2 \) and even less than \( |\tilde{H}(k)| \cdot |\tilde{H}(k+N/2)| \) has limited the difference between the two weighting factors to an acceptable range whereby the afore-mentioned undesirable sign reversion of the demapped value can be prevented. As a result, a balanced weighting on two sub-carriers has been achieved with a good compromise between maintaining the signal integrity and suppressing the noise. This, of course, is helpful for a soft decision channel decoder to achieve good decoding performance. [25]

### 4.2.4 The DCM Demodulation Algorithm Based on the Diversity Technique

**1. The problem**

In the front, the formula of Dual Carrier Modulation has been give. And commonly, when the receiver processed the data, it always adversely changes with the transmitter, and gets the general expression.

Now, assuming the channel is ideal, then \( Y_T = X_T \), \( X_R = X_R \), so the expression (4.26) changes to (4.30).

\[
\begin{bmatrix}
Y_R(n + \frac{mN}{4}) \\
Y_R(n + \frac{mN}{4} - \frac{N}{2})
\end{bmatrix} = \frac{1}{\sqrt{10}} \begin{bmatrix} 2 & 1 \\ 1 & -2 \end{bmatrix} \begin{bmatrix} X_R(2n+mN)+jX_R(2n+mN+\frac{N}{2}) \\ X_R(2n+mN+1)+jX_R(2n+mN+\frac{N}{2}+1) \end{bmatrix}
\] (4.30)

Where \( m = 0, 1; N = 100; 0 \leq n \leq N/4-1 \)

In the receiver, the formula (4.30) processes with the matrix in reverse change, and can get (4.31).

\[
\begin{bmatrix} X_R(2n+mN)+jX_R(2n+mN+\frac{N}{2}) \\ X_R(2n+mN+1)+jX_R(2n+mN+\frac{N}{2}+1) \end{bmatrix} = \frac{\sqrt{10}}{5} \begin{bmatrix} 2Y_R(n+mN/4)+Y_R(n+mN/4+N/2) \\ Y_R(n+mN/4)-2Y_R(n+mN/4+N/2) \end{bmatrix}
\] (4.31)

Where \( m = 0, 1; N = 100; 0 \leq n \leq N/4-1 \)

But actually, the channel for \( Y_T \) transmission is not ideal. It will be affected by
many kinds of the factors shown in (4.2), according to (4.31), $X_R$ also contains noise. In order to get the pure $X_T$, we need to analysis the relationship between $X_R$ and $X_T$. In order to implement this course, we adopt two steps. First, insert (4.2) to (4.31) and can get the relationship between $X_T$ and $Y_T$. Then, we insert (4.26) to the front $Y_T$, and get the relationship between $X_R$ and $X_T$ shown in (4.32). In (4.32), the "Noise" is AWGN. Analysis (4.32), we can find that the channel $H$ is the complex channel. After demodulating, the mapped information $X_R$ compared with $X_T$, is changed from 1 bit information to 4 bits information. The result comes from two factors. First, the channel has multiply interfere and it is a complex number, i.e., $H(i)$ in (4.32): Second, the two different groups are modulated from two different kinds of carriers, and the channel gains interfere the data. This can be seen in (4.32), the second factor $[H(n+mN/4)-H(n+N/2+mN/4)]$ couldn’t counteract with each other. But these are not all faults. If we can make good use of the channel diversity, we can not only counteract the information interference but also get the additional gain. So the two kinds of DCM demodulation methods based on diversity technique are grown in the following (4.32).

\[
X_R(2n+mN) + jX_R(2n+mN+N/4) = J\left[4H(n+mN/4)+H(n+mN+N/4)\right]X_T(2n+mN) + jX_T(2n+mN+N/4) + \text{Noise}
\]

\[
X_R(2n+mN+1) + jX_R(2n+mN+N/4+1) = J\left[H(n+mN/4)+4H(n+mN+N/4)\right]X_T(2n+mN+1) + jX_T(2n+mN+N/4+1) + \text{Noise}
\]

\[
X_T(2n+mN) + jX_T(2n+mN+N/4) = \text{Noise}
\]

\[
X_T(2n+mN+1) + jX_T(2n+mN+N/4+1) = \text{Noise}
\]

### (2) The DCM demodulation method

In order to make fully use of the diversity information to demodulate the data, we can process the data with two methods according to the interference. One is to use the biggest combine ratio diversity technique which is to add the channel information to the received signal, and change the complex multiply interference to the real gain factor. This can be implemented by multiplying the channel estimation information’s conjugate so that the two group data is still orthogonal with each other in the receiver. Second we can modulate the two group data received to the same channel gain factor, and the data can have two channels’ information. This can be implemented by multiplying the complementary channel estimate information. Then we will describe
the two methods in detail.

Method 1) the biggest combine ratio diversity technique

According to the biggest combine ratio diversity technique, the received \( Y_R(k) \) should be processed with formula (4.33) firstly.

\[
\hat{Y}_R(k) = Y_R(k) \ast \hat{H}(k) \quad k \in [0, 99]
\]

(4.34)

where, \( \hat{H}(k) \) is the estimation of channel \( H(k) \), \( \hat{H}(k) \ast \) is conjugate of \( \hat{H}(k) \). We assume the channel estimation is totally right, which is \( H(k) = \hat{H}(k) \). Then we make \( \hat{Y}_R(k) \) as the received data to instead of the \( Y_R(k) \) in formula (4.31), and get formula (4.34).

\[
\begin{bmatrix}
X_R(2n+mN) + jX_R(2n+mN+N/2) \\
X_R(2n+mN+1) + jX_R(2n+mN+N+1)
\end{bmatrix}
= \frac{\sqrt{M}}{3} \begin{bmatrix}
2\hat{Y}_R(n+mN/4) + \hat{Y}_R(n+mN/4+N/2) \\
\hat{Y}_R(n+mN/4) - 2\hat{Y}_R(n+mN/4+N/2)
\end{bmatrix}
\]

(4.35)

Then we insert (4.33) into (4.34), and insert (4.2) into \( Y_R(k) \), and we can get the relationship of \( X_R \) and \( X_T \) shown in formula (4.35).

\[
X_T(2n) = \frac{H(2n)^2 + 4H(2n+1)^2}{3||H(2n)||H(2n+1)} X_R(2n) - \frac{2H(2n)^2 - 2H(2n+1)^2}{3||H(2n)||H(2n+1)} X_R(2n+1)
\]

(4.36)

\[
X_T(2n+1) = \frac{H(2n)^2 + 4H(2n+1)^2}{3||H(2n)||H(2n+1)} X_R(2n+1) - \frac{2H(2n)^2 - 2H(2n+1)^2}{3||H(2n)||H(2n+1)} X_R(2n)
\]

in formula (4.35), \( n \in [0, 99] \).

We can see the diversity phenomenon from formula (4.35). According to (4.32), 1bit received data contain 4 bits related transmitted information, so the two signals in the diversity receiver can get the transmitted signal.

Analysis the formula (4.35), we can also see that, if the two modulated signals are multiplied with the same channel gain, which means that the \( H(2n) \) and \( H(2n+1) \) can be counteracted after transforming, and then the affection from the (4.32) back item will disappear. So we can get method 2.

Method 2) Estimated channel diversity technique

Analysis formula (4.32) we can find if the received information is affected by the same channel gain factor, the border item interference will disappear. In order to implement this, we need to process the received data with (4.36).
\[ \hat{Y}_R(k) = Y_R(k) \hat{H}(k + 50) \]
\[ \hat{Y}_R(k + 50) = Y_R(k + 50) \hat{H}(k) \quad (4.37) \]

Where \( k \in [0,49] \). Like method 1, we assume the channel estimation is ideal \( \hat{H}(k) = H(k) \), then use \( \hat{Y}_R(k) \) as the received information to instead the \( Y_R(k) \) in formula (4.31), and get formula (4.34). After that, we insert (4.36) into (4.34), and insert (4.2) into \( Y_R(k) \), then we can get the relationship between \( X_R \) and \( X_T \) as shown in formula (4.37).

\[ X_T(k) = \frac{aX_R(k) + bX_R(k+50)}{a^2 + b^2}, \quad X_T(k) = \frac{aX_R(k+50) - bX_R(k)}{a^2 + b^2} \quad (4.38) \]

where, \( a = \text{Re}[H(k)H(k+50)], b = \text{Im}[H(k+50)], a^2 + b^2 = |H(k)H(k+50)|^2 \).

The diversity can be seen from (4.37) that \( X_R(k) \) and \( X_R(k + 50) \) are got from the same transmitted data but different channel. After the process of formula (4.36), each modulated data goes with the same channel gain factor; the interference from the border received bit can be eliminated. And we can get the demodulation information exactly.

In conclusion, the DCM modulation is to process the information with two constellation modulation using 4 bits as the unit. We can understand to divide the data into IQ way, then modulate respectively with the QPSK constellation. The diversity demodulation methods are completed after analyzing the DCM modulation characters. Method 1 (the biggest combine ratio diversity technique) first eliminates the effect from the complex channel, then use the non-conherence channel frequency diversity to estimate the right data. Method 2 (estimated channel diversity technique) first eliminated the effect from the dual carriers mismatch, then demodulate the DCM using the QPSK associated carriers. But analysis from the theory, these two methods all use the non-conherence channel information with the diversity technique to get the demodulation results, and they should get the same performance. In the forth part we’ll show the simulation result to prove the methods.[27]

### 4.3 Performance Compare

#### 4.3.1 Simulation Compare

In order to prove the improvement of the system performance brought by the diversity demodulation method, we code the algorithm into the UWB system and
simulate under the MATLAB environment. The system framework is shown in Figure 3-2. The system parameters are shown in the following Table 4.3 and Table 4.4. In order to comparation, we also simulate the other demodulation methods to see the results.

<table>
<thead>
<tr>
<th>Table 4.3 System parameters-1</th>
</tr>
</thead>
<tbody>
<tr>
<td>datarate</td>
</tr>
<tr>
<td>Frame length</td>
</tr>
<tr>
<td>Simulation repetition (packets)</td>
</tr>
<tr>
<td>Channel condition</td>
</tr>
<tr>
<td>Carrier offset</td>
</tr>
<tr>
<td>Sample offset</td>
</tr>
<tr>
<td>Channel estimation</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Table 4.4 System parameters-2(according toDC-OFDM standard)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bit rate</td>
</tr>
<tr>
<td>Modulation mode</td>
</tr>
<tr>
<td>Time spread factor(TSF)</td>
</tr>
<tr>
<td>Frequency spread factor</td>
</tr>
<tr>
<td>Information bits per symbol(N_CBPS)</td>
</tr>
<tr>
<td>Interleave block length</td>
</tr>
</tbody>
</table>

When the signal to noise ratio(SNR) changes from 5dB to 16dB, the packet error rate changes with the SNR, which is shown in Figure 4-3.
Figure 4-3 The packet error rate comparison with different demapping methods

From the Figure 4-3 we can see that the diversity demodulation method can make the packet error rate less than 8% with the SNR 8.5dB. Compared with the other methods, the demand for the SNR is depresses 1.5dB, which prove our method. For the two diversity methods both use the non-conherence channel information, the performance is accordant which is shown in Figure 4-3.

4.3.2 Hardware Implementation Compare

The merits for diversity demodulation algorithm are obvious. It can not only estimitate the information much exactly, but also can decrease the hardware implementation complexity. First, this method has merged the equalization technology, and its results can be send to the Hanming-distence Vitibi decoder directly, as we know, Hanming-distance Vitibi decoder is easy to implement than the Euclidean-distance Vitibi decoder. What’s more, this algorithm has low complexity, the signal processing need complex multiplier, real multiplier and real adder to be implemented. Figure 4-4 describes it.
4. DCM Demodulation Algorithm

From the Figure 4-4 we can see that one point information need only two complex multipliers, two complex adders and two real adders. Compared with the traditional LLR estimation logarithm arithmetic and the distance measure method which need several square arithmetic, the complexity is greatly decreased.

4.4 The Hardware Implementation of Mapping & Demapping Module in the System

4.4.1 Hardware Implementation of Mapping

(1) Function Description

The UWB system based on the OFDM technique transmits data with symbol as the unit. The purpose of this hardware module is to map the binary data onto the constellation points (QPSK/DCM), and then map the pilots and data onto the fixed frequency. Simply, that means to arrange the pilots and data in sequence, and organize the 128 data to process IFFT. Another important thing for hardware implementation is to combine the two path signals into one which is convenient for the signal to experience the time spread and zero inserted.

(2) Module Structure

First input the data into the register file “reg_in”. When all data needed for one symbol is saved, we choose the mapping mode (QPSK/DCM) using the parallel path information according to the DC-OFDM standard, change the input data into the defined symbol, and save the data with pilots into the register file “reg_out”, which need only one cycle, and then output the data from the “reg_out” in sequence.
4. DCM Demodulation Algorithm

Figure 4-5 Mapping module structure

Figure 4-6 shows the interference of the mapping module.

(3) Design in Detail

According to the different data rate, there are three mapping modes:

1) Rate=53.3Mbps/80Mpbs. In this mode, 100-bit data is mapped onto 128 points according to the QPSK constellation. Then conjugate the data and finish the frequency spread.

First, save 200-bit continuous data into the register file “reg_in”, when it’s full, map the data onto the constellation points according to the QPSK in one cycle, and insert the pilots and the data into “reg_out_i” and “reg_out_q” in sequence. In the same cycle, the data begin to be outputted from the “reg_out_i” and “reg_out_q”. The timing is shown in Figure 4-7. Assuming the rate=53.2Mbps with two parallel path, then after 100 cycles, 200-bit data would all be saved into “reg_in”. If the output parallel path is four (real part is four and imagine part is also four), then after 32 cycles, the 128 complex data would all be outputted. Then the timing demands to wait
4. DCM Demodulation Algorithm

9 clocks for the zero inserted, and these 128 points should be spread in frequency field according to DC-OFDM standard. Then wait 92 clocks, the next group data would be saved in the reg_in, and it would work repeatedly.

Figure 4-7 Mapping input/output timing (2 path)

2) Rate=106.4Mbps/160Mbps/200Mbps. In this mode, 400-bit data would be mapped into 2 groups of 128 points according to QPSK, without frequency spread.

First, save the 400-bit continuous data into the register file “reg_in”, when it’s full, map the data onto the constellation points according to the QPSK in one cycle, and save the pilots and the data into “reg_out_i” and “reg_out_q” in sequence. In the same cycle, the data begin to be outputted from the “reg_out_i” and “reg_out_q”. The timing is shown in Figure 4-8. Assuming the rate=106.4Mbps, with four parallel path, then after 100 cycles, 400-bit data would be all saved into “reg_in”. If the output parallel path is four (real part is four and imagine part is also four), then after 32 cycles, the 128 complex data would all be outputted. Then the timing demands to wait 9 clocks for the zero inserted according to DC-OFDM standard. After waiting 92 clocks, the next group data would be saved in the reg_in, and it would work repeatedly.

Figure 4-8 Mapping input/output timing (4 path)

3) Rate=320Mbps/400Mbps/480Mbps. In this mode, 400-bit data would be mapped into 2 groups of 128 points according to DCM, without frequency spread.

First, save the 400-bit continuous data into the register file “reg_in”, when it’s full, map the data onto the constellation points according to the QPSK in one cycle, and save the pilots and the data into “reg_out_i” and “reg_out_q” in sequence. In the
same cycle, the data begin to be outputted from the “reg_out_i” and “reg_out_q”. The timing is shown in Figure 4-9. Assuming the rate=320Mbps, with eight parallel path, then after 50 cycles, 400-bit data would be saved into “reg_in”. If the output parallel path is four (real part is four and imagine part is also four), then after 32 cycles, the 128 complex data all be outputted. Then the timing demands to wait 9 clocks for the zero inserted according to DC-OFDM standard. And after that, the next group data would be saved in the “reg_in”. But because of the interleave timing (the front module of the mapping) is 50 clocks valid, 32 clocks invalid, 50 clocks valid and then 33 clocks invalid. So after mapping module process 4 symbols, there will be 10 clocks invalid. And these four symbols will integrate into one group for the back module to insert the zero periods.

![Diagram](image_url)

Figure 4-9 Mapping input/output timing (8 paths)

(4) Discussion

In the front structure, the output data don’t need to wait the 200/400-bit data all coming in and then to change, so we can consider to decrease the register depth. The input 2/4/8-bit data can be changed immediately when they come, and in the next clock, the data in “reg_in” is saved in the “reg_out”. After 100/50 clocks, the data can be outputted. In this case, the depth of “reg_in” is 8, and the depth of “reg_out” is 400+4*28=512.

Otherwise, we can use memory to instead of register file. But memory couldn’t be read or write several addressed data simultaneously, so it couldn’t complete the data converse from “reg_in” to “reg_out” in one clock. But we can use two or more memory to finish this process.

But in the course of the simulation, we find when the high speed payload follows with the header, the timing will have problem showing in Figure 4-10.
Figure 4-10 Timing with the payload in rate=320Mbps/400Mbps/480Mbps after header

According to Figure 4-10, o_txd_valid should be valid when the first snap line appears in
Figure 4-11. But

Figure 4-11 shows us at this time the header data hasn’t been output totally. If we wait for it, the output of the payload will delay 50 clocks, so we demand specially for the interleave module. This is when the rate is 320/400/480 Mbps, the payload should delay 115 clocks to come while not 65 clocks.

Figure 4-12 and

Figure 4-13 show the sequence when the payload rate is lower than 320Mbps, and we can see from the figures that this situation couldn’t happen.
4. DCM Demodulation Algorithm

Figure 4-13 Timing with the payload in rate=106.4Mbps/160Mbps/200Mbps after header

We implement the module in Xilinx Virtex 5 LX330; Table 4.5 shows the hardware utilization:

<table>
<thead>
<tr>
<th>Slice Logic Utilization</th>
<th>Used</th>
<th>Available</th>
<th>Utilization</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of Slice Registers</td>
<td>1,215</td>
<td>207,360</td>
<td>1%</td>
</tr>
<tr>
<td>Number of Slice LUTs</td>
<td>3,466</td>
<td>207,360</td>
<td>1%</td>
</tr>
</tbody>
</table>

4.4.2 Hardware Implementation of Demapping

(1) Function description

This module demodulates the constellation points onto binary data according to QPSK/DCM, and then chooses the 128-point data from FFT module. Combine the dual carrier signals. For the low and middle speeds, spread reverse.

(2) Module structure

![Demapping module structure](image-url)
## 4. DCM Demodulation Algorithm

![Diagram](image_url)

**Figure 4-15 Interface description**

(3) **Design in detail**

According to the different rate, there are three kinds of demapping modes.

1) Rate = 53.3Mbps/80Mbps. In this mode, four group data should be first processed with the spread in reverse, including time spread in reverse and frequency spread in reverse. Then we would get 200-point data (each point data has three bits).

First, send the 2 continuous symbols into “reg_out” register. Real and imaginary parts should be saved into two “reg_out”. When the third data comes, time and frequency spread in reverse should be processed simultaneously and then be outputted. The timing is shown in Figure 4-16. Assuming rate=53.3Mbps, and input with four parallel path, including real and imaginary part. Then after 32+9+32+9=82 clocks, the front two groups’ data are all saved in the “reg_out”, then they should be demapped and outputted in sequence. Output has 2 parallel paths, and each path has 3 bits. After 100 clocks, 200-bit data are outputted. Waiting for 65 clocks, the next group data would go into “reg_out” and process in sequence.

![Diagram](image_url)

**Figure 4-16 Demapping timing in low speed mode**

2) Rate = 106.7Mbps/160Mbps/200Mbps. In this mode, four group data should be first processed with the time spread in reverse. Then we would get 200-point data (each point data has three bits).
First, send the 2 continuous symbols into “reg_out” register. Real and imaginary parts should be saved into two “reg_out”. When the third data comes, time spread in reverse should be processed and then be outputted. The timing is shown in Figure 4-17. Assuming rate=53.3Mbps, and input with four parallel path, including real and imaginary part. Then after 32+9+32+9=82 clocks, the front two groups’ data are all saved in the “reg_out”, then they should be demapped and outputted in sequence. Output has 4 parallel paths, and each path has 3 bits. After 100 clocks, 400-bit data are outputted. Waiting for 65 clocks, the next group data would go into “reg_out” and process in sequence.

3) Rate=320Mbps/400Mbps/480Mbps. In this mode, after demapping, two group data would change to 400-point data. (Each data is 3 bits).

First, send the 2 continuous symbols into “reg_out” register. Real and imaginary parts should be saved into two “reg_out”. When the third data comes, output it. The timing is shown in Figure 4-18. Assuming rate=320Mbps, and input with four parallel path, including real and imaginary part. Then after 32+9+32+9=82 clocks, the front two groups data are all saved in the “reg_out”, and they should be demapped and outputted in sequence. Output has 8 parallel paths, and each path has 3 bits. After 50 clocks, 400-bit data are outputted. After waiting for 32 clocks, next group data would all be saved into “reg_out” and then processed in sequence. But in order to constitute 165 clocks, the next group data should wait for 33 clocks.
(4) **Discussion**

1) The output of the module can be connected with hard input Vitibi decoder. That means the output is the quantized from -3 to 3.

2) After the header, the payload continues with different rates. The timing is shown in the following.

When the rate is fixed in 53.2Mbps, the timing is shown in Figure 4-16.

When the rate is changed from low speed mode to middle speed mode, the timing is shown in Figure 4-19.

When the rate is changed from low speed mode to high speed mode, the timing is shown in Figure 4-20.

3) In Figure 4-20, the output valid signal waits for 15 clocks to output the
payload. And it decides 15 clocks for the process speed of the de-interleave module. It also affects the feedback time of the whole system. If the system waits more than 15 clocks, the depth of the registers would increase, but it can loose the sequence of the whole system. And that’s the tradeoff.

The FPGA results for demapping module are shown in table 4.6.

<table>
<thead>
<tr>
<th>Slice Logic Utilization</th>
<th>Used</th>
<th>Available</th>
<th>Utilization</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of Slice Registers</td>
<td>6,772</td>
<td>207,360</td>
<td>3%</td>
</tr>
<tr>
<td>Number of Slice LUTs</td>
<td>32,740</td>
<td>207,360</td>
<td>15%</td>
</tr>
</tbody>
</table>

### 4.4.3 Compatible Design of Mapping and Demapping for Different Standards

For the mapping/demapping module, the differences between MB-OFDM and DC-OFDM standard are the depth of the data block and the spread modes. The mapping of pilots, guard periods and zero periods are same in these two systems. The direct implementation is to map the data carriers, guard periods and pilots to the fixed positions according to the DC-OFDM standard shown in chapter 3. In this way, we can decide the relationship between the input registers and output registers. Then, the data is outputted according to the address of the registers. But this implementation will increase the complexity and decrease the hardware agility. When the speed is less than 200Mbps, the two standards all use QPSK modulation, while when the speed is more than 200Mbps, the two standards all use DCM modulation. Because there will be many input and output registers, the complex connects will introduce the crosstalk and so on to decrease the system performance. On the other side, in order to implement the module compatible, we design the address generator. This module can generate the input/output address according to the different standards and constellations to control the read/write process. As shown in Figure 4-21, the input data goes into the registers with different paths (2/4/8) at the different speeds. The mapping mode controls the address generator to save the input data, while the address generator controls the data stream to save data in the output registers. When the speed
is less than 80Mbps, 100-point output data can generate a symbol. When the speed is more than 200Mbps, 400-point output data can generate a symbol. While when the speed is between them, 200-point output data can generate a symbol. The output data would be sent to a MUX unit; the control signal controls the output and then finishes the mapping output. The last function for this module is the time/frequency spread according to the different data rate. After this, the final data of this module outputs in sequence.

![Block diagram of mapping](image)

Demapping is the process in reverse of the mapping. According to the ECMA-368/DC-OFDM UWB standard, OFDM UWB system is the simplex system. So the mapping/demapping module can share the same hardware by time division multiplex application (TDMA). But when the speed is less than 80Mbps, the inter symbol frequency spread exists. In order to use the frequency diversity to get the integrated signal, shown in Figure 4-22, we design the operation module after the output registers. Note that the output of demapping is the soft information used for VitiBi decoder. And in our system, it’s 3 bits. So we need more registers/memories to save the data than the mapping module.
4. DCM Demodulation Algorithm

Figure 4-22 Block diagram of demapping
Chapter 5 System Hardware Implementation

5.1 Transmitter Structure

The transmitter processes the data from MAC layer, generates the frame defined in DC-OFDM standard, and outputs to digital analog conversion (DAC).

Transmitter contains such modules as FIFO (tx_fifo), Scrambler, Reed-Solomon encoder, Convolution encoder, Puncture, Interleave, QPSK/DCM Mapping, IFFT, Zero inserted, Dual carriers generated (insert the preambles).

Note: In order to be compatible with MB-OFDM system, transmitter generates the dual carrier signals in the last module. The timing in front is compatible with the MB-OFDM timing (132MHz×4).

![Figure 5-1 UWB transmitter](image)

Clock design:

When the data comes through different modules, the data stream changes. But they change without any multiple relations. So we couldn’t have the clock in accordance with the data. For if we do this, there will be many different clock sources, and their relations are complexity so that they are not easy to be created from the same clock source. If there are several clock sources, the multi-clock fields design will increase the difficulty. Considering this, we use one clock and one data valid signal to control the stream. There are nine data rates from 39.4Mb/s to 480Mb/s, so we choose one tradeoff frequency to be the system clock, that’s 132MHz. The data stream demands can be satisfied through different parallel paths. The Table 5.1 lists the parallel factor in different modules at different speeds. Obviously, if the biggest data throughput is more than the output data throughput, the data wouldn’t lose. The data throughput is controlled by the data valid signal. Actually, the data valid signal is related to the throughput. Table 5.1 only lists the transmitter parameters of the
modules, and the receiver parameters can be got easily from the table, only change the output data to the input data.

Table 5.1 The data throughput and the parallel factor for each module

<table>
<thead>
<tr>
<th>module</th>
<th>Data rate (Mbps)</th>
<th>Output data throughput</th>
<th>Parallel factor</th>
<th>The biggest output data throughput</th>
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5. System hardware structure and implementation

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<td>2</td>
<td>264x2</td>
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5.2 Transmitter Sub-module Design

m_tx_fifo

This module is the interface between the MAC layer and PHY layer. The data from MAC layer has nine data rates, that's 39.4~480Mbps, the interface clock is 66MHz, data bandwidth is 8bit. M_tx_fifo changes the MAC data (66MHz) to the PHY data (132MHz). Different data rates are controlled by the output parallel factor and the data valid signal timing.

m_scrm

This module is to avoid the continuous “0” or “1”. It is implemented by a pseudo random sequence to scramble the continuous “0” or “1”, and make “0” “1” to have the same the probability.

m_encoder

This module contains RS encoder and Convolution encoder. RS encoder only used for the Header. Generally, Convolution encoder is described by 3 parameters (n, k, K). Here k/n denote the code efficient (information contained in 1 coded bit), K is the restriction length, denotes the levels of the coded registers. An important character that Convolution encoder different from group encoder is that the Convolution has memory function. The output is not only the function of the Kth group data, but also the function of the front K-1th group data.

m_punc

This module is the puncture. It is used to create the high rate codes. The principle is first we design a low rate encoder, and then we delete several bits to make it a high rate encoder. This can avoid the fixed complexity of the high rate Convolution encoder.

m_intlv

This module used before modulation module to interleave the coded data. In the memory channel, the errors are not the absolute bits. They are coherent with each other. If we interleave the coded information before transmission and deinterleave the
information in the receiver, the abrupt error happened in the channel would be spread, and the decoder can deal the errors as the random errors. The key is to divide codes in the time domain. And the time between them can be padded by the other coded bits. In the system, the timing of \texttt{m_intlv} is the key point. The output of this module uses OFDM group as the unit. The timing of the valid signal is to ensure the final transmitted data continuously.

\textbf{m\_mapping}
This module has been described in front, so here omitted.

\textbf{m\_ifft}
This module processes the IFFT operation, and the input/output data is arranged in sequence.

\textbf{m\_insert\_zp}
This module finishes two functions. 1) Insert 32 zero prefixes before the 128 points, and insert 5 zero postfixes behind the 128 points. The zero prefixes have the same function with cyclic prefixes, used for the interference elimination. The zero postfixes supply the time period for the frequency-hopping. 2). When the data rate is less than 320Mbps, use the time spread technique.

\textbf{m\_dc\_gen}
This module is used to form the dual carriers. The \texttt{m\_insert\_zp} module can build an OFDM group using the continuous two OFDM symbols. Change the clock domain from 4×132MHz to 2×264MHz.

\section*{5.3 Receiver Structure}
Receiver processes the sampled signals from ADC with signal processing techniques in order to decrease the noise and channel attenuation, and then deals the signal with the transform in reverse and returns them to MAC layer.

The main modules contain synchronization (time synchronization, frequency synchronization), remove the zero prefix, parallel to serial, FFT, Channel Estimation and equalization, QPSK/DCM Demapping, deinterleave Viterbi decoder, Reed-Solomon decoder, Descrambler, received FIFO.

Note: In order to be compatible with the MB-OFDM system, the receiver uses parallel to serial module after synchronization and removing zero prefixes.
5. System hardware structure and implementation

Figure 5-2 UWB receiver

5.4 Receiver Sub-module Design

**m_s2p**
This module changes the DC-OFDM received signal from the 264MHz (2*132MHz) clock domain to the 132MHz (4*132MHz) clock domain.

**m_sync**
This module contains the TFC detection, symbol synchronization and Carrier Frequency Offset (CFO) cursory estimation.

**m_rx_prmtr_gen**
This module calculates the OFDM symbols and the pad bits, and then creates the data valid signal used for the Viterbi decoder and other modules.

**m_cfo**
This module uses the synchronization information from the m_syn module to estimate and compensates the time-domain phase used for CFO in different hopping frequency. Note that for the DC-OFDM, we should compensate the different carrier phases and different frequency groups respectively.

**m_rmvzp_dc**
This module uses zero prefixes to eliminate the symbol interference. Being different from the cyclic prefixes, the zero prefixes should add the front 32 points of the \( K + 1 \)th symbol to the tail 32 points of the \( K \)th symbol. The module uses this operation and outputs 128-point valid data. According to the valid signal of the dual carriers, this module combines them (from 2*2 to 4*1).

It doesn’t deal with the time spread which should be deal after equalization and synchronization.

**m_fft**
This module processes the FFT operation, and arranges the input/output data in
sequence. The structure is same with m_iift, and just chooses fft operation mode.

**m_equ**
This module estimates and compensates the channels in frequency.

**m_track**
This module tracks the sample offset, carrier offset estimation and channel compensation according to the pilot in OFDM symbol.

**m_demapping**
This module has been described in front, so here omitted.

**m_deintlv**
The structure is same with interleave module. Just reverse the course.

**m_decoder**
The system uses Viterbi algorithm to decode, containing de-puncture and Viterbi decoder. It decodes the PLCP Header with RS-Decoder and gets the frame information of the PLCP Header. Note that pad bit won’t be outputted.

Note: The time of header decoding should be early than the payload decoding or the m_demap module wouldn’t deal with the data normally. In the DC-OFDM system, the symbol number of header is \((266*3+2)*2/100=16\) before demapping, which needs \(16*(32+9)+4=660\) clocks. That means the time from m_demap module to m_rs_decoder should be less than 660 clocks.

**m_descrm**
This module won’t work when PHY header going through. Tail bits needn’t be deal but the states of the registers should change.

**m_rx_interface**
This module is the interface between PHY layer and MAC layer. It transmits the outputs of the m_descrm to MAC layer, and changes clock domain from 132MHz to 66MHz.
Chapter 6 Conclusion and Future Work

In this thesis, we implement the digital base band of the PHY layer in DC-OFDM UWB system. When we do this work, we first design the system in Matlab. In this way, we simulate the base band in different channel environments in order to verify the correctness of the algorithm. And through this way, we can get the validation data of each module for the hardware implementation. When we do the hardware implementation, we code the module according to the algorithm design in the Matlab, after simulate them, we add the device delay, wire delay and also the other influence with the ISE using FPGA Virtex 5. The transmission data can be looped back through the wire. This verifies that the base band implementation is OK.

When we use the UWB technique, we couldn’t only use the base band. We must combine the radio frequency module through ADC and DAC. The data stream is coming from MAC layer, then through the PHY base band, DAC, transferred to the RF modules. Through the RF module, the data is modulated to the high frequency. Receiver works when the RF first detects the data. Then the data is transferred to the ADC module and goes to the PHY base band. So then, we should work with RF module to verify the system working in different environment demanded.

In the future, UWB system will be very popular. But in order to expand the market, I think the DC-OFDM UWB system should be implemented combined with MB-OFDM. Considering the compatibility with different standards in different environments, the different modes should be switched to the other easily. In this thesis, the mapping/demapping module has been implemented considering the compatibility. And the other base band modules can also be designed similarly. In this way, the multi-standard UWB system will have more customers, more markets and more application environments.
Bibliography

[4] Anuj Batra. Why MB-OFDM DS technique is more appropriate for high speed UWB communication[R].
Acknowledgements

As time goes by, the three years will be over, and I will leave school. When I was an undergraduate student, I dreamed that one day I could studied in Fudan University. But in a few days, I’ll leave. The three years’ time is short. But I’ll remember the days in Fudan forever. In these three years, I have grown up from a freshman in microelectronics to an ASIC engineer. And I have learned to be a person with high responsibility.

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Last my thanks would go to my beloved family for their loving considerations and great confidence in me all through these years. They have done their best for my grown. Whenever and whatever I am, you are always my spirit source. Thank everyone in my family. Thank you forever. And I also would like to thank my boyfriend Mingshuo Wang. Thank you for your support and encouragement. Thank you for your understanding and endurance. Thank you for coming along with me in my difficult time. Thank you for your lovely care. Meeting you is my destiny.

Wish everyone healthy and happy.