Study on Ultra-Wideband Wearable Body Area Communication for Medical Applications

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学位授与年月日 | 2010-03-23
URL | http://id.nii.ac.jp/1476/00002928/
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NAGOYA INSTITUTE OF TECHNOLOGY
Nagoya, Japan
January 2010
Doctoral Dissertation

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Submitted to
Department of Computer Science and Engineering,
Graduate School of Engineering,
Nagoya Institute of Technology
Nagoya, Japan

January 2010
Summary

Ultra wideband (UWB) wearable body area communication is of high importance for promising new biomedical applications. An adequate propagation channel model is essential for the design of a UWB body area communication system. However, there are scattering and absorption phenomena in the human body due to its frequency-dependent dielectric properties, as well as diffractions and creeping waves along the body surface. In particular, since the human body might take various postures, or one or more body parts might move during the communication period, this results in a complicated multi-path signal transmission. There are currently few measurements or models describing on-body area propagation channels which sufficiently imitate practical body postures and movements. Using the frequency-dependent finite difference time domain (FDTD) method and a realistic adult male body model, this work simulates various body postures for modeling on-body channels. Based on the FDTD numerical results, we derive an on-body propagation model and determine the model parameters for some representative transmission links on the human body for medical applications. The derived channel model is a discrete impulse response model which is convenient for characterizing the multi-path effect. By virtue of a high time resolution in the numerical simulation and concluded low correlation between multi-path rays, this impulse response model is implemented based on distinguishable multi-path rays without resorting to a uniformly-spaced tapped delay line model. A good match is obtained between the data derived from FDTD method and the statistically implemented models in terms of key communication metrics. Furthermore, an experiment is performed in order to verify the chest-to-right-waist transmission model, and it is found that all parameters obtained from the experiment are in good agreement with the statistical parameters, which provides additional verification of the statistical model.

The channel modeling enables the analysis and evaluation of on-body communication performances. The body postures and movements induce multi-paths in on-body propagation channel, so that a received signal ends up being the superimposition of several attenuated,
delayed, and eventually distorted replicas of a transmitted waveform. System performance is therefore in fact significantly degraded by the distortion of pulses due to propagation over such a real channel. On the other hand, the performance degradation can be mitigated if a detailed characterization of the multi-path channel is available at the receiver. The maximum desirable data rates for upper body area communication by conventional correlation receiver are concluded as 70 Mbits/s and 90 Mbits/s which correspond to the bit error rate (BER) threshold of 0.001 and 0.01 respectively. Conventional correlation receiver seems not competent in multi-path-affected channel concluded from the performance of probability of error, while RAKE receiver outperforms the correlation receiver at a cost of structure complexity. Under the Federal Communications Commission (FCC) emission spectral mask, the effective maximum communication distance for pulse position modulation time hopping UWB (PPM-TH-UWB) with 2 and 4 fingers RAKE receiver is at least around 0.6-0.8 m on the human body. The 2-finger RAKE receiver can compromise the poor communication performance of conventional correlation receiver and the complicated structure of 4-finger RAKE receiver. The gratifying communication performances based on practical scheme consideration make the medical body area networking (MBAN) feasible. The UWB wearable body area communication systems can be expected to perform adequately for MBAN applications.

Last but not the least, UWB wearable body area communications inevitably brings forward the human body safety and bio-electromagnetic compatibility (Bio-EMC) issues since it operates on human body. Electromagnetic emission from body worn communication devices may give rise to body tissue energy absorption as well as possible interference to implanted medical device such as a cardiac pacemaker. These issues have to be considered in the design of the BAN communication system, but almost no study has been done until now. For the human safety issue, although the signal from one UWB device is very low, however, it is unclear that the energy absorption will increase to what extent when multiple UWB devices are adorned simultaneously to a human body, which is the actual situation for a body area network. From the viewpoint of biological safety evaluation, this work proposes two approaches in time-domain and frequency-domain respectively to calculate the specific
energy absorption (SA) and specific absorption rate (SAR) for multiple UWB device exposure. It is shown that the two approaches have the same accuracy but the time-domain approach is more straightforward in the numerical analysis. We have also demonstrated the SA/SAR levels under the FCC UWB emission limit for various antenna locations on an anatomical human body model. The results have shown that the SA and SAR levels are much smaller than the IEEE safety limits and moreover the multiple exposure with several UWB devices adorn simultaneously does almost not obviously increase the SA and SAR as long as two UWB devices have a separation as large as 30 cm. For the potential electromagnetic interference (EMI) to the cardiac pacemaker, this work presents a two-step hybrid approach to model the induced EMI voltage at the cardiac pacemaker by on-body communication signals. The hybrid approach consists of an electromagnetic (EM) field analysis and a nonlinear electric circuit analysis. The EM field analysis employs the FDTD method to calculate the interference input voltage of the pacemaker circuit by considering the pacemaker as a receiving antenna. The electric circuit analysis employs a nonlinear operational amplifier (opamp) model with Volterra series representation to predict the output voltage of the analogue sensing circuit of pacemaker for evaluating the EMI effect. The predicted interference output voltage has shown a good agreement with the measured result from 10 to 100 MHz in literature, which confirms the validity of the proposed approach. Furthermore, applying this approach to a UWB wearable body sensor communication scenario, we have obtained the interference output voltage of a pacemaker under the FCC UWB emission limit, and demonstrated it is much smaller than the actual sensing threshold of pacemaker, i.e., with a safety margin of around 35 dB.
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<th>Description</th>
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<tbody>
<tr>
<td>AICc</td>
<td>Akaike Information Criterion</td>
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<td>APDP</td>
<td>Average Power Delay Profiles</td>
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<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<td>BAN</td>
<td>Body Area Network</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Bit</td>
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<tr>
<td>CDF</td>
<td>Cumulative Density Function</td>
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<tr>
<td>ECG</td>
<td>Electrocardiogram</td>
</tr>
<tr>
<td>EEG</td>
<td>Electroencephalogram</td>
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<tr>
<td>EGC</td>
<td>Equal Gain Combining</td>
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<tr>
<td>EIRP</td>
<td>Equivalent Isotropically Radiated Power</td>
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<tr>
<td>EMC</td>
<td>Electromagnetic Compatibility</td>
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<tr>
<td>EMI</td>
<td>Electromagnetic Interference</td>
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<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
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<tr>
<td>FDTD</td>
<td>Finite-Difference Time-Domain</td>
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<tr>
<td>ICNIRP</td>
<td>International Commission on Non-Ionizing Radiation Protection</td>
</tr>
<tr>
<td>IR</td>
<td>Impulse Radio</td>
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<tr>
<td>ISI</td>
<td>Inter-Symbol Interference</td>
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<tr>
<td>LAN</td>
<td>Local Area Network</td>
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<tr>
<td>MBAN</td>
<td>Medical Body Area Network</td>
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<tr>
<td>MRC</td>
<td>Maximal Ratio Combining</td>
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<tr>
<td>opamp</td>
<td>Operational Amplifier</td>
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<tr>
<td>PAM</td>
<td>Pulse Amplitude Modulation</td>
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<tr>
<td>PDF</td>
<td>Probability Density Function</td>
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<td>PDP</td>
<td>Power Delay Profile</td>
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XI
<table>
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<tr>
<th>Abbreviation</th>
<th>Full Form</th>
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<tr>
<td>PML</td>
<td>Perfectly Matched Layer</td>
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<tr>
<td>PN</td>
<td>Pseudo Noise</td>
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<tr>
<td>PPM</td>
<td>Pulse Position Modulation</td>
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<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
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<tr>
<td>RF</td>
<td>Radio Frequencies</td>
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<tr>
<td>RMS</td>
<td>Root Mean Square</td>
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<tr>
<td>SA</td>
<td>Specific Energy Absorption</td>
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<tr>
<td>SAR</td>
<td>Specific Absorption Rate</td>
</tr>
<tr>
<td>SD</td>
<td>Selection Diversity</td>
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<tr>
<td>SNR</td>
<td>Signal Noise Ratio</td>
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<tr>
<td>TH</td>
<td>Time-Hopping</td>
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<tr>
<td>UWB</td>
<td>Ultra Wideband</td>
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<tr>
<td>WPAN</td>
<td>Wireless Personal Area Networks</td>
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Chapter 1 Introduction

Today's aging population is deriving wide-scale demand for more advanced and efficient healthcare treatment with the aid of information and communication techniques. The information and communication techniques have played more and more important roles in supporting medical and healthcare services. The ever-advancing miniaturization of electronic devices, combined with recent developments in wearable computer technology, is leading to the creation of a wide range of personal information and communication appliances which can be attached to the bodies or clothes of users (Zimmerman 1996). Such devices include cellular phones, PDAs, pocket games, biotic sensors, and so on. Such wireless network devices located on or near a human body can share data, reduce functional redundancies, and allow for new conveniences and services. For example, the human body communication allows high security and convenience by transmitting a password signal through the human body to a user certification system, because the signal radiation towards the outside of the human body can be minimized. Also, it may provide new possibility of high-quality service from hospitals by linking various biotic sensors to establish a body area network of personal health information. Using wireless sensors and appliances around the body to monitor body information is a promising new biomedical application for medical and health care service.

As an emerging technology, body area networks (BAN) have caught significant attention in recent years (Astrin et al. 2009). As one of the working groups under IEEE 802 standard of wireless communication, IEEE 802.15 concentrates on wireless personal area networks (WPAN). A working group for medical body area network (MBAN) IEEE 802.15.6 group was established under IEEE 802.15 in November 2007 and the standardization is now under the way.
1.1 Motivation

Body area communication is of high importance for promising new biomedical applications. However, there are scattering and absorption phenomena in the human body due to its frequency-dependent dielectric properties, as well as diffractions and creeping waves along the body surface. In particular, since the human body might take various postures, or one or more body parts might move during the communication period, this results in a complicated multi-path signal transmission.

The ultra wideband (UWB) system is a promising candidate for use in such situations due to its efficiency with respect to multi-path fading, low power consumption, high transmission speed, and simple structure (Shmidt et al. 2002; Gyselinckx et al. 2005). The IEEE 802.15.4a standard has been developed as a low-rate wireless personal area network UWB standard for various application scenarios.

An adequate propagation channel model is essential for the design of a UWB body area communication system. Only few papers are actually focusing on UWB BAN propagation channel modeling for wireless biomedical communications. (Zhao et al. 2006) is based on finite difference time domain (FDTD) numerical approach with simplified human models. Attempts also focus on FDTD simulations for narrowband systems (Latre et al. 2004; Ryckaert et al. 2004) and UWB systems in the 3–6-GHz band (Fort et al. 2005). Narrowband measurements near the body in the 2.4-GHz band (Hall et al. 2002.), as well as UWB measurements in the 3–6-GHz band (Welch et al. 2002; Zasowski et al. 2003; Kovacs et al. 2004a; Kovacs et al. 2004b; Fort et al. 2006a) have also been reported.

However, these derived channel models do not include the statistical characteristics of body postures and movements. Fort et al. first developed an on-body UWB channel model based on measurements in the 3 – 6 GHz frequency range (Fort et al. 2006b). They fixed the locations of the transmitters on the chest and moved the receiver around the body. The derived statistical characteristics were therefore mainly based on the different locations of the receiver, although the case of arm movement was studied. For a BAN application, however,
the transmitter and the receiver might be fixed in most cases. Therefore, models based on various body postures or body movements should be more general. Tan et al. made an attempt to develop a statistical model in order to account for different links (Tang et al. 2006). The model parameters were derived experimentally, and no numerical phantoms were employed, which limits the variety of the postures.

Moreover, the ultimate performance limits of BAN communications systems, as well as the performance of practical systems, are determined by the channel it operates in. BAN communication differentiates with other wireless communications with the human body as the transmission channel. The multi-path-affected channel characteristics on human body greatly depend on body postures and movements. Besides, transmit power in BAN communication is restricted by human safety guidelines. Based on the BAN transmission characteristics and the human safety consideration, it is feasible to clarify the communication performance and depict communication scheme. These related issues cover the maximum possible data rate, the possible communication distance on body, the modulation and demodulation schemes, and the effective demodulator structure and so on.

On the other hand, since BAN communicates around the human body, it inevitably brings forward the human body safety and bio- electromagnetic compatibility (Bio-EMC) issues. Safety to human body has a higher priority than the other wireless communications. The transmit power should be limited as low as possible in order to assure the safety to humans. This is also one of the major issues to be addressed by IEEE 802.15.6 standard (Astrin et al. 2009). If the electromagnetic emission from body worn communication devices is over high, the specific energy absorption (SA, defined as the energy absorbed per unit mass of biological body) and the specific absorption rate (SAR, defined as the power absorbed per unit mass of biological body) may violate the human safety guideline which is used to ensure the human safety under electromagnetic exposure. This issue has to be considered in the design of the BAN communication system, but almost no study has been done until now. On the other hand, for UWB wearable body area communications, although the signal from one UWB device is very low, however, it is unclear that the energy absorption will increase to what extent when
many UWB devices are adored simultaneously to a human body, which is the actual situation for a body area network. An analysis method is therefore required from the point of view of biological safety evaluation.

Besides the human safety issue, electromagnetic interference (EMI) from body communications to other implanted device like cardiac pacemaker cannot be ignored. External electromagnetic (EM) fields can couple into the pacemaker to cause an interference voltage at the input of the internal sensing circuit. The induced interference voltage at the input of the sensing circuit of pacemaker will be amplified and low-pass filtered. When the output voltage of the amplifier and low-pass filter exceeds a threshold, the pulse voltage to simulate the heart beat may be triggered and a malfunction of cardiac pacemaker occurs. A hybrid modeling method, which can implement both electromagnetic modeling and electric modeling, is needed in order to predict the potential interference voltage at pacemaker.

1.2 Objectives

The main objective of this thesis is to develop an effective UWB wearable body area communication system for medical and healthcare applications. Three major issues are covered: (1) channel modeling, (2) communication performance, and (3) human safety and EMC.

The first objective of this thesis is to derive an on-body UWB channel model with an emphasis on the statistical variations of body postures and movements. The present studies have been limited to the path loss modeling for a standing human body. However, in most cases, the transmitter and the receiver are fixed in some body locations, and the human may have various body postures or moves. A channel model, including the body postures and movement effect, is therefore necessary for a wearable body area communication system design.

The approach is outlined as follows:
• Obtain the on-body propagation characteristics by using human body models and full wave FDTD simulation
• Extract the statistical channel characteristic from power delay profile
• Implement the channel by re-producing a discrete impulse response
• Confirm the validation via experimental measurement

The second objective is to make clear the communication performance for UWB communications in the multi-path-affected on-body channel based on the derived channel model and the human safety consideration. The possible data rate and the possible communication distance, as well as the communication bit error rate (BER) performance and the effective demodulator structure, are clarified. It mainly contains 3 parts: (1) derive the maximum data rate according to error correction technique; (2) investigate the improvement effect such as RAKE receiver structure for on-body multipath fading; (3) calculate the effective communication distance at given data rates and receiver structures.

The third objective is to clarify the SA/SAR and EMI to implanted pacemaker. For SA/SAR evaluation, we will propose two approaches in the time-domain and frequency-domain to calculate the SA and SAR for UWB pulse exposure. Then calculate the SA for a single UWB pulse exposure and derive the SAR based on the UWB pulse radio technique. At Federal Communications Commission (FCC) maximum allowed transmit power we will also calculate the 10-gram averaged SA and SAR under single exposure and multiple exposures with multiple UWB devices adorned on body simultaneously in order to clarify that the energy absorption will increase to what extent when many UWB devices are adorned simultaneously to a human body. Besides, the variation relationship of SA/SAR and the distance between the antenna and the human body will be also calculated to show the influence of the antenna location on the SA/SAR results. For EMI to implanted pacemaker, we will present a two-step hybrid approach to model the induced EMI voltage at the cardiac pacemaker by on-body communication signals. The hybrid approach consists of an EM field analysis and a nonlinear electric circuit analysis. The EM field analysis employs the FDTD method to calculate the interference input voltage of the pacemaker circuit by considering the
pacemaker as a receiving antenna. The electric circuit analysis employs a nonlinear operational amplifier (opamp) model with Volterra series representation to predict the output voltage of the analogue sensing circuit of pacemaker for evaluating the EMI effect. Safety margin has been concluded for cardiac pacemaker under the FCC UWB emission limit.

1.3 Overview

This research produced several journal and conference papers and this work is a compilation of those papers' propositions and results. The papers' material was rearranged in a more comprehensive order and a more extensive review of the literature was made.

Chapter 2 presents a review about the terminology and definitions covered in the UWB communication of body area network. Basic concepts are given to clarify the BAN application background and BAN classifications, as well as to further introduce the subjective of this thesis research work. Wearable BAN for medical and healthcare services is described in detail to show the practical application value of body area communication research. UWB emission mask regulations as well as the ongoing standardization work are also introduced.

Chapter 3 establishes a comprehensive dynamic UWB wearable body area channel model with an emphasis on various body postures and movement. A frequency-dependent FDTD numerical method and a human body model capable of various body postures are developed. This chapter describes channel characterization and parameterization based on the FDTD calculation results. Amplitude and time information is statistically derived to implement the discrete impulse response function. Experimental measurement is also carried out to validate the statistical derived channel model.

Based on the derived channel model in chapter 3, chapter 4 discusses and evaluates the communication performances for given pulse position modulation time hopping UWB (PPM-TH-UWB) system with and without the RAKE receiver in the multi-path-affected chest-to-right-waist channel. The possible data rate and the possible communication distance, as well as the communication BER performance and the effective demodulator structure, are clarified.
Different RAKE receiver structures are compared to show the improvement of the BER performance.

Chapter 5 calculates SA and SAR levels for various antenna locations on an anatomical human body model and also the interference safety margin for pacemaker under the FCC UWB emission limit. The results have shown that the SA and SAR levels are much smaller than the International Commission on Non-Ionizing Radiation Protection (ICNIRP) safety limits and the UWB multiple exposures do almost not obviously increase the SA and SAR. Two approaches in the time-domain and frequency-domain to calculate the SA and SAR for UWB pulse exposure are presented. The time-domain approach is more straightforward to the SA and SAR analysis for UWB systems. The relationship of SA/SAR and the distance between the antenna and the human body are also concluded. Moreover, the interference output voltage of a pacemaker under the FCC UWB emission limit is much smaller than the sensing threshold for UWB wearable communication applications. A two-step hybrid approach, which consists of an EM field analysis and a nonlinear electric circuit analysis, is presented to model the induced EMI voltage at the cardiac pacemaker by on-body communication signals.

Chapter 6 makes a summarization of the main conclusions of this research and also discusses the remaining research for future work.
Chapter 2 Terminology and Definitions

This chapter aims to elaborate the basic concepts and terminology covered in the UWB wearable BAN communications in this research. Starting from basic BAN categorizations and wearable BAN application scenarios for medical and healthcare services, it covers UWB definition and characteristics for BAN communications, closing with related international standardizations.

2.1 BAN

2.1.1 BAN definition

According to IEEE 802.15, the BAN is a technology, that to provide interconnection of wearable lightweight sensors and/or actuators and/or devices attached to or in the vicinity of the body (< 2 – 5 meters), depending on a person’s need. The BAN can take a continuous measure and transmit a vital sign or body physiological data to facilitate the remote monitoring for the purpose of health care service, assistance to people with disabilities and, body interaction and entertainment.

2.1.2 BAN applications

There are different categorizations for BAN application and usage models. Figure 2-1 shows a categorization given by (Astrin et al. 2009). They are (I) medical healthcare services, (II) assistance to people with disabilities, and (III) body interaction and entertainment.
There is a wide range of applications for BAN in supporting medical and healthcare services. In general manner, a BAN device is a BAN transceiver pairing with a life sign sensor or a set of life sign sensors. These sensors can collect various vital and healthcare data for medical or healthcare purposes. Examples of collectable sensor data include blood pressure, SpO2, electroencephalogram (EEG), electrocardiogram (ECG), carotid pulse, glucose rate, body temperature, etc. A typical application scenario of using these sensor data is real time monitoring of patient state in a hospital. By attaching BAN devices to patients, vital or healthcare data can be automatically collected, which is then forwarded to a nurse centre for patient state monitoring. The benefit of this scenario is that it can reduce the working load of nurses and result in increased efficiency on patient management. Other usage models of using sensor data include at home healthcare, aging people support, physical rehabilitation assistance, etc.

Advanced medical applications can be extended to medical closed loop control. In this case, the sensor data are sent to a control unit. At the control unit proper medical measure is decided and sent to an operation unit. The operation unit carries out medical treatment obeying the command from the control unit. One example is an automatically controlled pacemaker. A pacemaker is an electronic device that helps people with irregular heart beat problem. First, a pacemaker controller needs to collect sympathetic nerve signal using sensors.
Then, the pacemaker controller calculates the correct heart beat rate and instructs the pacemaker. Finally, the pacemaker helps to adjust the heart beat to a correct beating rhythm. Another example is automatic insulin injection for diabetes patients. Using the data from glucose sensor, an injection controller can decide the correct amount of insulin to be injected. Then, an insulin pump carries out the injection according to the instruction from the controller.

In the second category of applications, there are also plenty of examples. One of the typical examples is the application for assisting people with visual disabilities. In this application, BAN sensors are attached to the belongings of a person. A reasonable range between these sensors with a receiver carried by the person is set in advance. When the person forgets his belongings and leaves them over the pre-set range, warning signal is generated automatically. Furthermore, in advanced applications, cameras can be attached to a people with visual disability. Pictures taken by cameras are sent to a receiver carried by the people, where they are converted to audio signal to provide guidance to the people. The similar principle can be used for assisting people with speech disability. Here, sensors to catch finger and hand movements are used. The obtained information caught by sensors is converted into speech.

Examples of the third category of applications include user interface, wireless headphone, simultaneous audio and video, audio or video streaming, video game controller, entertainment data and sensor, etc.. Using BAN in these applications can not only increase convenience by deleting wires but also provide a method of source sharing. For example, two or more persons can share a same music player by using wireless headphone.

2.1.3 Wearable BAN and implant BAN

BAN can be divided into wearable BAN and implant BAN according to its location in or on the body where it operates.

Figure 2-2 shows the wearable BAN links and implant BAN links. Wearable BAN links have all of the devices on body, while implant BAN links have some devices inside body which communicate with on-body or outside-body devices.
There are different characteristics between wearable and implant BAN. Wearable BAN may suffer from multipath channel and shadowing, while implant BAN mainly undergoes severe signal decay during transmission. According to (Astrin et al. 2009), the following differences exist between a wearable BAN and an implant BAN.

1. Different requirements on frequencies due to different operating environment on and in body or air channels. Wearable BAN covers the whole UWB band from 3.1 GHz to 10.6 GHz, while implant BAN covers mainly the UWB low band from 3.4 GHz to 4.8 GHz, which is determined by the different propagation environments.

2. Battery powered implant BAN devices are generally more power limited and sometimes requires smaller or specific shape form factor due to their location in a body (e.g. hearing aid or a pacemaker).

3. Both need to consider tissue protection (e.g., SAR or transmit power restriction), while wearable BAN has the freedom of choosing an antenna pattern which is pointed away from sensitive parts of the body.

This thesis mainly focuses on the wearable BAN research for medical and healthcare services. Implant BAN is out of the scope of this thesis, we will not go into further detail.

2.1.4 Wearable BAN for medical and healthcare services

As a promising new biomedical application for medical and health care service, BAN may provide new possibility of high-quality service from hospitals by linking various biotal sensors to establish an on-body area network of personal health information. Wearable BAN
for medical and healthcare services is one of the major BAN branches. There is a wide range of applications for BAN in supporting medical and healthcare services. As shown in Figure 2-1, medical and healthcare services mainly cover three parts: medical check-up, elder people assistance and physical rehabilitation, and physiological monitoring. The medical and healthcare services are outlined as follows:

(1) Medical check-up
   - EEG sensors for monitoring brain electrical activity
   - ECG sensors for monitoring heart activity
   - A breathing sensor for monitoring respiration
   - A blood pressure sensor
   - Heart rate
   - Body temperature

(2) Elder people assistance and physical rehabilitation
   - Tilt sensors for monitoring accident fall-down
   - Foot sensors for monitoring steps
   - Movement sensors for monitoring activities
   - A breathing sensor for monitoring respiration
   - A blood pressure sensor
   - Heart rate
   - Body temperature

(3) Physiological monitoring
   - An accelerate sensor for monitoring instant
   - Foot sensors for monitoring steps
   - A breathing sensor for monitoring respiration
   - A blood pressure sensor
   - Heart rate
   - Body temperature

Figure 2-3 shows a wearable BAN using wireless sensors and appliances around the body to monitor personal health information. Various on-body sensors monitor body health
information like EEG, ECG, blood pressure, body temperature, vision information etc. for medical or healthcare purposes. The main sensor located at the left chest will collect the information from all of the sensors and then send the collected data to remote hospital via cellular or local area network (LAN). The scenario is no doubt beneficial since it can reduce the working load of hospital and result in increased efficiency on medical management.

![Figure 2-3 Wearable BAN for medical and healthcare services](image)

### 2.2 UWB

The last two years have witnessed an increased interest in standardization bodies in UWB. Appealing features such as flexibility and robustness, as well as high-precision ranging capability, have polarized attention and made UWB an excellent candidate for a variety of applications. UWB is emerging as a particularly appealing transmission technique for applications requiring either high bit rates over short ranges or low bit rates over medium-to-long ranges. IEEE 802.15.4 standard (IEEE 802.15.4-2003, 2003) has been brought forward as low rate wireless personal area network UWB standard for various applications.
The UWB signal is defined with the fractional bandwidth greater than about 0.20-0.25 or the whole occupied bandwidth is greater than 500 MHz. Traditionally, UWB signals have been obtained by transmitting very short pulses, rather than continuous waveforms, with typically no radio frequencies (RF) modulation. This technique has been extensively used and goes under the name of impulse radio (IR). In April 2002, the FCC approved the first guidelines allowing the intentional emission of UWB signals contained within specified emission mask between 3.1–10.6 GHz (FCC, 2002). Figure 2-4 shows the FCC UWB emission mask. From 3.1 GHz to 10.6 GHz, the maximum emission power density is not allowed to exceed -41.3 dBm/MHz.

UWB transmission is a powerful candidate for wireless BAN. As for the feature of UWB, the power spectrum density is extremely low and the effects against another system are low. From these, it can be thought that the influence against medical equipment by the radio wave is less than other telecommunications systems including Bluetooth, wireless LAN, cell phones, cordless phones etc. Simultaneously, the wideband nature of the UWB technology permits a fine time resolution. It is particularly beneficial to biomedical applications, e.g., health monitoring, human body probing, real-time diagnosis, etc. (Staderini 2002; Tan and Chia 2004), which basically require low transmit power together with fine time-resolution. In addition, the miniaturization and energy-saving of the telecommunication equipments can be given. All of these features make UWB an ideal candidate for wireless BAN applications.

Figure 2-4 FCC UWB emission mask
2.3 Standardizations

IEEE 802 standardization committee is an international organization that develops international standards on wireless communication. As one of the working groups under IEEE 802, IEEE 802.15 (WG15) concentrates on wireless personal area network (WPAN). WG15 had created a number of wireless standards. Examples include IEEE 802.15.1 which is also known as Bluetooth, IEEE 802.15.4 (Gutierrez et al. 2004) which defines the physical layer (PHY) for low rate WPAN and is applied for Zigbee, IEEE 802.15.4a which defines an alternative PHY for IEEE 802.15.4 using UWB technology (Astrin et al. 2007; Li et al. 2006), etc.. There are also several ongoing standardizations including IEEE 802.15.3c of high-rate WPAN using millimetre wave, IEEE 802.15.4c for Chinese WPAN, and IEEE 802.15.4d for Japanese WPAN.

A standardization committee referred to as IEEE 802.15.6 was formally set up in December 2007. The objective of 15.6 is to define new physical and media access control layers for wireless BAN (WBAN). Main focus of IEEE802.15.6 is laid on medical healthcare applications. (Kohno et al. 2008) introduces a progress of research and development of BAN standardization in IEEE 802.15.6 in a field of medical information communication technology. The progress of the IEEE 802.15.6 standardization covers a number of issues like frequency regulations, human safety, SAR, low power radio requirement, high quality of service and low latency as well as PHY layer discussion for wearable BAN and implant BAN. Moreover, a simple prototype of wearable BAN has been given to address the issues of modulation scheme and possible data rate. A number of works remain left and currently the IEEE 802.15.6 group are working at several documents including (1) application matrix, (2) technical requirements, (3) regulation report, and (4) channel models. The IEEE 802.15.6 standard is expected to form in 2010.
Chapter 3 Channel Modeling

3.1 Introduction

Channel modeling is the first step to develop an effective wearable UWB BAN communication system for medical and healthcare applications. The objective of channel modeling is to derive a reasonable UWB wearable channel model for wireless BAN communications. The channel model in this research put an emphasis on the statistical variations of body postures and movements.

A realistic human body model and a numerical EM field analysis technique are employed for the derivation of the model. The human body model was developed on the basis of average statistical values of the body parameters of Asian adults (RIHEL 1997) and using data from the database for human body sizes established by the Research Institute of Human Engineering for Quality Life, Japan. The derivation of generalized channel characteristics can benefit from the construction of a model of a statistically average human body.

The numerical analysis technique is known as the frequency-dependent FDTD method (Kunz and Luebbers 1993). The merits of this numerical approach are as follows: (1) It is more flexible with respect to modeling various situations in on-body area communication since a numerical human model can easily simulate various body postures; (2) it can use a statistical average of human body models, which produces better results and is more convenient to work with than the experimental methods, which require a large number of people in order produce a generalized experiment; (3) it is easy to extend the frequency to the entire UWB band (3.1-10.6 GHz) and achieve high time resolution for the multi-paths; and (4)
the influence of the structure of the antennas on the channel modeling can be eliminated by employing an ideal point source. However, the numerical approach also has some shortcomings in comparison with experimental measurements. For example, the computation time is usually very long, and it is also not easy to include the effects of the surrounding environment.

In this chapter, we will first describe the frequency-dependent FDTD numerical method and the human body model employed in this thesis. And then based on the numerical approach we will characterize and parameterize the on-body channel from several items starting from path loss, following the power delay profile and arriving time of multi-path ray. A small set of model parameters will be given for five typical transmission links in which the transmitter is fixed at the left chest and the receivers are located at the right chest, the left and right waist, and both ears respectively. Measurement will be also performed in order to verify the derived channel model. Finally, the channel model will be implemented in Matlab in the form of discrete impulse response.

3.2 Computation Method

3.2.1 Frequency-dependent FDTD method

Due to the frequency-dependent dielectric properties of the human body with respect to UWB signals, the frequency-dependent FDTD method was employed in the numerical approach. The details of this method can be found in (Kunz and Luebbers 1993), and some minor modifications have been made in our algorithms in order to better simulate human tissue (Wang and Nishikawa 2007). This method is based on the time-domain Maxwell curl equations, where the electric flux density $D$ is related to the electric field $E$ through the complex permittivity of human body tissue.

The time-domain Maxwell curl equations are written as

$$\nabla \times E(t) = -\mu_0 \frac{\partial H(t)}{\partial t}$$

(3-1)
\[ \nabla \times H(t) = \frac{\partial D(t)}{\partial t} + J_0(t) \]  

(3-2)

where \( H \) is the magnetic field, \( J_0 \) is the current density due to static electric conductivity, and \( \mu_0 \) is the free-space permeability. The complex permittivity of human tissue can be approximated by the Debye equation

\[ \varepsilon_r(\omega) = \varepsilon_\infty + \chi(\omega) + \frac{\sigma_0}{j\omega\varepsilon_0} \]  

(3-3)

where \( \varepsilon_0 \) is the free-space electric permittivity, \( \varepsilon_\infty \) is the relative electric permittivity at frequencies approaching infinity, \( \chi(\omega) \) is the frequency-domain susceptibility, and \( \sigma_0 \) is the static electric conductivity.

We can relate the first and second items in Equation (3-3) to \( D(\omega) \) and the third item to \( J_0(\omega) \) because only the former two terms are due to the frequency dispersion of tissue. Therefore we have

\[ D(\omega) = \varepsilon_0[\varepsilon_\infty + \chi(\omega)]E(\omega) \]  

(3-4)

\[ J_0(\omega) = \sigma_0 E(\omega) \]  

(3-5)

Since Equation (3-1) and Equation (3-2) need to be solved iteratively in the time-domain by using the FDTD method, Equation (3-4) and Equation (3-5) must be transferred to the time domain. This can be realized by performing a Fourier transform as follows

\[ D(t) = \varepsilon_0 \varepsilon_\infty E(t) + \varepsilon_0 \chi(t) \ast E(t) \]  

(3-6)

\[ J_0(t) = \sigma_0 E(t) \]  

(3-7)

where the symbol \( \ast \) denotes the convolution operator.

Substituting the two equations into the time-domain Maxwell curl equations, we can obtain the frequency-dependent FDTD formulation for analyzing UWB transmission on the human body. The derivation is given in follows:

\[ \therefore D(t) = \varepsilon_0 \varepsilon_\infty E(t) + \varepsilon_0 \chi(t) \ast E(t) \]

\[ \therefore D^n = \varepsilon_0 \varepsilon_\infty E^n + \varepsilon_0 \sum_{m=0}^{n-1} E^{n-m} \chi^m \]
\[ D_n = \varepsilon_0 \varepsilon_\infty E_n + \varepsilon_0 \sum_{m=0}^{n-2} E_n^{1-m} \chi^{m+1} \]  
(3-8)

\[ D_n^{n-1} = \varepsilon_0 \varepsilon_\infty E_n^{n-1} + \varepsilon_0 \sum_{m=0}^{n-2} E_n^{1-m} \chi^m \]  
(3-9)

\[ \varepsilon \frac{\partial D(t)}{\partial t} = \nabla \times H(t) - J_0(t) \]  
(3-10)

\[ \varepsilon \frac{D_n^{n-1}}{\Delta t} = \nabla \times H_n^{n-1/2} - \sigma_0 E_n^{n-1/2} \]  
(3-11)

Assume one-relaxation Debye approximation i.e.,

\[ \chi(\omega) = \frac{\varepsilon_s - \varepsilon_\infty}{1 + j\omega \tau} \]  
(3-12)

in Equation (3-3), then

\[ E_n = \frac{\varepsilon_\infty - \sigma_0 \Delta t}{\varepsilon_\infty + \chi^0 + \frac{\sigma_0 \Delta t}{2 \varepsilon_0}} E_n^{n-1} + \frac{1}{\varepsilon_\infty + \chi^0 + \frac{\sigma_0 \Delta t}{2 \varepsilon_0}} \phi^{n-1} + \frac{\Delta t/\varepsilon_0}{\varepsilon_\infty + \chi^0 + \frac{\sigma_0 \Delta t/2 \varepsilon_0}{\varepsilon_\infty + \chi^0}} \nabla \times H_n^{n-1/2} \]  
(3-13)

where

\[ \phi^{n-1} = \sum_{m=0}^{n-2} E_n^{1-m} (\chi^m - \chi^{m-1}) = E_n^{n-1} (\chi^0 - \chi^1) + e^{-\Delta t/\tau} \phi^{n-2} \]  
(3-14)

\[ \chi^m = \int_{m \Delta t}^{(m+1) \Delta t} \chi(\tau) d\tau = (\varepsilon_s - \varepsilon_\infty) e^{-\frac{m \Delta t}{\tau}} \left( 1 - e^{-\frac{\Delta t}{\tau}} \right) \]  
(3-15)

The fundamental flow chart of this frequency-dependent FDTD method with recursive convolution scheme is shown in Figure 3-1. In the flow chart, the PML means perfectly matched layer which is an artificial absorbing layer for wave equations. It is used to truncate computational regions in numerical methods to simulate problems with open boundaries. It can absorb outgoing waves from the interior of a computational region without reflecting them back into the interior.

### 3.2.2 Human body model

The human body model used in this research was developed on the basis of average statistical values of the body parameters of Asian adults (RIHEL 1997. The derivation of generalized
channel characteristics can benefit from the construction of a model of a statistically average human body.

The model has been developed based on magnetic resonance imaging data. As shown in Figure 3-2 (Nagaoka et al. 2004), the model has a height of 173 cm, a weight of 65 kg, and a spatial resolution of 2 mm. It consists of 51 tissue types. For incorporating the tissue properties into the frequency-dependent FDTD method, as described in section 3.1.1, we assume one-relaxation Debye approximation, i.e. Equation (3-3), and determine the parameters $\varepsilon_s$, $\varepsilon_\infty$, and $\tau$ for all of the 51 tissue types. Assuming Gabriel's measurement-based data (Gabriel 1996) as the true values, we determine these parameters by using the least square error fitting in the UWB frequency band from 3.1 to 10.6 GHz. Figure 3-3 shows the fitted complex relative permittivity values for muscle, skin and bone as a function of frequency. The imaginary parts are negative in mathematics. The straight lines are measurement-based data, and the broken lines are one-relaxation Debye approximation. They have fair agreement at this frequency band between 3.1 to 10.6 GHz. Figure 3-4 shows the fitting errors averaged over the entire frequency band for all of the tissues in the human body.
model. The average difference is found to be within ±10% between the one-relaxation Debye approximation and measurement-based data in the interested frequency band. Moreover, the maximum difference at any frequency between 3.1 to 10.6 GHz is within ±20% for all of the 51 tissues. Although a second-order (Gandhi and Furse 1997) or high-order Debye relaxation approximation (Gabriel 1996) offers the advantage of better fit to the measurement data, the one-relaxation Debye approximation has a reasonable accuracy.

Since the UWB frequency band ranges from 3.1 to 10.6 GHz, the practically usable skin depth of the biological tissue in this frequency band is only several millimeters. This suggests that the surface tissue layers should dominate the propagation characteristics of the body. In fact, based on the computational comparison between the received voltages for a homogeneous body model comprising only skin and a heterogeneous body model comprising 51 different tissue types, as reported in (Tayamachi et al. 2007), we found that the differences in the received voltages under these two situations never exceed 5%. We therefore assumed a homogeneous human body for the model, with dielectric properties of the skin, and determined the parameters for the one-relaxation Debye equation by applying the least square error fitting method to the reported measured data (Gabriel 1996).

![Figure 3-2 Anatomically based human body model](image)

(Male, 173 cm, 65 kg, 2 mm resolution, 51 tissue types)
Figure 3-3 Fitted complex relative permittivity values for muscle, skin and bone as a function of frequency

Figure 3-4 Fitting errors averaged over the entire frequency band of 3.1 to 10.6 GHz for all of the tissues in the human body model

3.2.3 Software implementation

The frequency-dependent FDTD code was developed in our laboratory and is described in detail in (Wang and Nishikawa 2007).
The validity of this FDTD code with respect to simulating biological tissue has been experimentally confirmed in (Wang et al. 2006), and its accuracy has been carefully examined in (Beard et al. 2006) through numerical comparisons performed by various international research groups.

Moreover, a computer tool was developed for modeling various human body postures and movements. The tool has a user-friendly interface, and the created body model can be easily incorporated into the frequency-dependent FDTD algorithm.

3.3 Channel Characterization and Parameterization

The characteristics of the propagation channels corresponding to different types of on-body propagation links can be expected to differ considerably, due to the variability of the link geometry. Here we fixed the transmitter which was a Hertzian dipole on the left chest and move the receiving points along the body. For the multi-path fading channel characterization, five receiving points on the right chest, the left and right waists, and both ears as shown in Figure 3-5, are taken as the main research objects. They correspond to five transmission links which can be considered as typical transmission links for medical and health care applications.

The effects of various body postures were accounted for by simulating 35 different postures, as seen in Figure 3-6. The details of the postures are as follows:

1) Standing: 9 postures
2) Walking: 10 postures
3) Running: 10 postures
4) Sitting: 6 postures.

Simulating various postures is indispensable in order to obtain a statistical characterization of the transmission channels, since the posture of the user is generally not fixed in medical and health care applications.
It is known that any deterministic impulse response can be represented by a tapped delay line model as long as the system is band-limited. Since the impulse response of the Saleh-Valenzuela model (Saleh and Valenzuela 1987) is applicable after appropriate modifications have been implemented, as in (Molisch et al. 2006) for UWB applications, here we apply and modify the Saleh-Valenzuela model in order to characterize the on-body UWB channel and to obtain the parameters required for the implementation. The modified Saleh-Valenzuela model in the time domain deserves practical implementation in system design and simulation.

In this chapter, first the path loss is derived based on a static heterogeneous body model.

### 3.3.1 Pulse shape

The choice of the transmitted pulse is crucial since it affects the power spectral density (PSD) of the transmitted signal. The pulse shape that can be generated in the easiest way by a pulse generator actually has a bell shape such as a Gaussian. A Gaussian pulse $p(t)$ can be described by the following expression:

$$p(t) = \pm \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{t^2}{2\sigma^2}} = \pm \frac{\sqrt{2}}{\alpha} e^{-\frac{2\pi t^2}{a^2}}$$  \hspace{1cm} (3-16)

where $\alpha = \sqrt{4\pi\sigma^2}$ is the shape factor and $\sigma^2$ is the variance.

To be radiated in an efficient way however, a basic feature of the pulse is to have a zero dc (direct current) offset. Several pulse waveforms might be considered, provided that this condition is verified. Gaussian derivatives are suitable. Actually, the most currently adopted pulse shape is modeled as the second derivative of a Gaussian function (Win and Scholtz 2000), described by:

$$\frac{d^2p(t)}{dt^2} = \left(1 - 4\pi\frac{t^2}{a^2}\right) e^{-\frac{2\pi t^2}{a^2}}$$  \hspace{1cm} (3-17)

Other pulse shapes have also been proposed such as the Laplacian (Conroy et al. 1999), compositions of Gaussian pulses (Hämäläinen et al. 2001), and Hermite pulses (Ghavami et al. 2002). We will use a 2nd-derivative Gaussian pulse as shown in Figure 3-7 as the transmitted
UWB pulse. Pulse width is tightly related to the shape factor $\alpha$. Reducing the value of $\alpha$ shortens the pulse, and thus enlarges the bandwidth of the transmitted pulse. As a consequence the same waveform can be used to occupy different bandwidths by adjusting the value of the pulse width. Pulse width of 280 ps is used in order to have most of the energy between 3.1 and 10.6 GHz.

Figure 3-5 Locations of the transmitter and some representative receiving points on the human body model

Figure 3-6 Typical body postures
3.3.2 Path loss

Measurements (Zasowski et al. 2005) and FDTD simulations (Fort et al. 2005) have shown that paths traveling through the body in the gigahertz range are significantly attenuated. Instead, waves diffract around the torso. Therefore, we measure the distance around the perimeter of the body when modeling the path loss.

The transmitter is a dipole located at the left chest, and the receiving points are located along the whole body as shown in Figure 3-8. Both the transmitter dipole and the received fields are normal to the body to facilitate the waves diffract around the body. The receiving points are separated by 10 cm along the body, while the separation between receiving points on the same plane is 2 mm. In the FDTD calculations totally 869 receiving points are calculated which consists of 250 points on front, 345 points on side and 274 points on back of the body. Both the transmitter and the receiving points have a spacing of 2 mm above the body surface.

The path loss can be obtained from the following equation:

\[ PL_{dB}(d) = -10 \log_{10} \left[ \frac{u_r(d)}{u_s(d)} \right] \]  \hspace{1cm} (3-18)
where $U_s(d)$ and $U_r(d)$ are the amounts of transmitted and received energy, which can be calculated from the transmitted and the received pulse voltage waveforms. Figure 3-9 shows the calculated path loss versus distance on the human body model. It is clear that the path loss increases with distance as expected, and that there is a large variance around the mean path loss. In Figure 3-9, the horizontal axis is not the straight-line distance but the one traveled by the wave around the perimeter of the body between the transmitter and the receiving points, because the major components reaching the receivers are due to the diffraction around the body.

![Figure 3-8 Transmitter and receiving points locations on the body](image)

![Figure 3-9 Path loss fitting vs. distance along the human body for front y-y](image)
Two path loss models are investigated for possible fitting to the calculated results. One is according to the empirical power decay law as

\[ PL_{dB}(d) = PLo_{dB} + 10n \log_{10}\left(\frac{d}{d_0}\right) \]  

(3-19)

where \( PL_{0,dB} \) is the path loss at distance \( d_0 \) and \( n \) is called the path loss exponent.

The other is according to an exponential fitting as

\[ PL_{dB}(d) = Ae^{\alpha d} \]  

(3-20)

where \( A \) is the excitation coefficient and \( \alpha \) is the attenuation coefficient. Based on the path loss data from the case in which both the transmitter and the receiving point are in front of body and the directions are along the y axis, the fitting results by using the two equations in the least-square-error sense are also shown in Figure 3-9 with the solid line and the dashed line, respectively. As can be seen, compared to Equation (3-20), Equation (3-19) gives a better fitting, which suggests that the power decay law is more appropriate for the UWB path loss modeling on the human body. This finding is identical to the results in (Fort et al. 2006b) but different from the result in (Zhao et al. 2006).

Table 3-1 shows the estimated \( PL_0 \) and \( n \) for various directed transmitter dipoles and receiving components. \( PL_{0,dB} \) is obtained at \( d_0 = 0.1 \) m. It is found that the path loss at the first 0.1 m is around 40 – 85 dB, and the exponent \( n \) is around 2 – 4. Besides, the path loss due to diffraction around the body is higher than the path loss due to waves traveling along the length of the body. This result is consistent with the measurement result in (Fort et al. 2006a). Furthermore, for the transmitter, the y-directed dipole yields a smaller reference path loss \( PL_{0,dB} \). On the other hand, at the receiving points, the z-directed field components always have the smallest path loss exponent \( n \). Altogether taking the two parameters into consideration, the y-directed transmitter dipole and y-directed receiving component may be the best option among all of the results.

For the typical five transmission links, which are taken as the main focus of this research, the parameters in Equation (3-19) are concluded as \( PL_{0,dB} = 43 \) dB at \( d_0 = 0.1 \) m and \( n =3.8 \).
The values of the respective path loss in each of these five transmission links are summarized in Table 3-2, from which it is clear that the path loss values are in the same order as those given in (Fort et al. 2006b) or (Zhao et al. 2006).

Table 3-1 Estimated $P_{L_0}$ and $n$ in the power decay law for various directed transmitter antenna and receiving components

<table>
<thead>
<tr>
<th>Transmitting Antenna</th>
<th>Received Components</th>
<th>Receiving Antenna</th>
<th>Front</th>
<th>Side</th>
<th>Back</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$P_{L_0}$ [dB]</td>
<td>$n$</td>
<td>$P_{L_0}$ [dB]</td>
<td>$n$</td>
<td>$P_{L_0}$ [dB]</td>
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<td>X</td>
<td>50.2</td>
<td>4.05</td>
<td>48.9</td>
<td>4.61</td>
</tr>
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<td></td>
<td>Y</td>
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<td>3.69</td>
<td>48.5</td>
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<td></td>
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<tr>
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<td>Z</td>
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<tr>
<td>Z</td>
<td>X</td>
<td>47.7</td>
<td>3.81</td>
<td>53.2</td>
<td>3.82</td>
</tr>
<tr>
<td></td>
<td>Y</td>
<td>50.1</td>
<td>3.62</td>
<td>52.3</td>
<td>3.66</td>
</tr>
<tr>
<td></td>
<td>Z</td>
<td>49.6</td>
<td>2.60</td>
<td>55.9</td>
<td>2.88</td>
</tr>
</tbody>
</table>

Table 3-2 Path loss values at five receiving points

<table>
<thead>
<tr>
<th>Receiving point</th>
<th>Distance [m]</th>
<th>Path loss [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Right ear</td>
<td>0.31</td>
<td>61.7</td>
</tr>
<tr>
<td>Left ear</td>
<td>0.26</td>
<td>58.8</td>
</tr>
<tr>
<td>Right chest</td>
<td>0.16</td>
<td>50.8</td>
</tr>
<tr>
<td>Left waist</td>
<td>0.53</td>
<td>70.5</td>
</tr>
<tr>
<td>Right waist</td>
<td>0.56</td>
<td>71.4</td>
</tr>
</tbody>
</table>

3.3.3 Power delay profile

The power delay profile (PDP), $\rho(\tau)$, is a statistical expression of the transmission channel characteristics. It can be derived from the impulse response $h(t)$, i.e.,

$$h(t) = F^{-1}(H(f)) = F^{-1}\left\{\frac{F[p_{\tau}(f)]}{F[p_\tau(f)]}\right\}$$

(3-21)

and

$$p(\tau) = \langle h(\tau) \cdot h^*(\tau) \rangle$$

(3-22)
where \( v_s(t) \) and \( v_r(t) \) are the transmitted and the received pulse voltages respectively, \( H(f) \) is the frequency-domain transfer function, and \( F\{ \cdot \} \) and \( F^{-1}\{ \cdot \} \) denote Fourier transform and inverse Fourier transform, respectively. The power delay profile characterizes the mean power of different multi-paths.

In the derivation of the impulse response \( h(t) \), Hamming windows were applied in the frequency domain in order to limit the signal to effective frequency components. \( F[V_s(t)] \) and \( F[V_r(t)] \) were extracted in the frequencies below 14 GHz in view of the dominating frequency components of the employed 2nd-derivative Gaussian pulses. The time resolution of the inverse Fourier transform for \( h(t) \) can be approximated as the reciprocal of the bandwidth (1/14GHz = 0.07 ns) multiplied by the additional window function bandwidth. Since the coefficient of the Hamming window is 2, this results in a 0.14-ns time resolution for the impulse response and the power delay profile.

Based on the data from all 35 postures, the average power delay profiles (APDP) for five typical transmission links are shown in Figure 3-10. It is concluded that one cluster is sufficient for describing the power delay profile in all 5 representative transmission links. Each peak in the cluster may be attributed to a multi-path ray, which results from the diffraction from the body surface or the reflection from a body part. No clusters corresponding to the surrounding environment were included, and the channel model extraction was limited solely to the human body. From Figure 3-10, as expected, the average power delay profile decays exponentially with the arrival time, i.e., it can be approximately expressed as

\[
p(\tau) = \Omega_0 e^{-\frac{\tau - \tau_0}{\gamma}}
\]

(3-23)

where \( \Omega_0 \) and \( \tau_0 \) are the mean power gain and the arrival time of the first ray, respectively. With the exponential fittings to the data in Figure 3-10, it was found that the ray decay time constants \( \gamma \) are 0.21 ns, 0.26 ns, 0.38 ns, 0.30 ns and 0.5 ns respectively for chest-to-right-chest link, chest-to-left-ear link, chest-to-right-ear link, chest-to-left-waist link and chest-to-right-waist link.
Figure 3-10 Average power delay profiles for five typical transmission links based on all 35 postures
3.3.4 Amplitude distribution

The lognormal distribution is reported in (Fort et al. 2006b) and (Zhao et al. 2006) as an excellent fit to the amplitude data for all receiver locations. In our case, we assume that all of the obvious peaks in the power delay profile correspond to the multi-path rays. We characterize the multi-path rays by identifying the corresponding peaks, and then obtain the arrival time and the amplitude of the individual rays. The peaks that have amplitudes 30 dB lower than the maximum peak value are taken into account in order to extract the channel parameters, while the other peaks are small enough to be ignored. We also carefully studied the amplitude distribution in the first and the second rays. Some possible candidates, such as the lognormal distribution, the Rayleigh distribution, the Rice distribution and the Weibull distribution, were considered for the amplitude model.

Similar to (Fort et al. 2006b), we also used the second-order Akaike information criterion ($AIC_c$) rather than a hypothesis test to rank the fitting results from best to worst.

The second-order $AIC_c$ is defined as follows (Burnham K. P. 2002):

$$AIC_c = -2 \log_e(\ell(\hat{\theta}|data)) + 2K + \frac{2K(K+1)}{(n-K-1)}$$

where $\log_e(\ell(\hat{\theta}|data))$ is the value of the maximized log-likelihood over the unknown parameters ($\theta$), given the data and the model, $K$ is the number of parameters estimated in that model, and $n$ is the sample size. This equation is straightforward to compute since the log likelihood is readily available from the maximum likelihood estimation. Intuitively, the first term indicates that better models have a lower $AIC_c$ because the log-likelihood reflects the overall fit of the model to the data. The second part of the equation penalizes additional parameters ensuring we select models that best fit the data with the least number of parameters. The $AIC_c$ also has a strong theoretical motivation since it provides an estimate of the Kullback–Leibler information loss (Akaike 1973). In this way, the model with the lowest $AIC_c$ approximates the “true” distribution with the minimum loss of information.

In practice, the value of the $AIC_c$ by itself has no meaning. However, the relative values of $AIC_c$ among the models can be used to rank the models from best to worst and to provide
strength of evidence that one model is better than another. To facilitate this, the two following related metrics are normally reported:

$$\Delta_i = AIC_{c,i} - \text{min}(AIC_c)$$  \hspace{1cm} (3-25)

$$w_i = \frac{\exp\left(-\frac{\Delta_i}{2}\right)}{\sum_{r=1}^{R} \exp\left(-\frac{\Delta_r}{2}\right)}$$  \hspace{1cm} (3-26)

where $AIC_i$ is the AIC value for model index $i$, and $R$ is the number of models. Clearly, the best model among the set of models has a delta AIC of 0. As a rule-of-thumb, $\Delta_i < 2$ suggests substantial evidence for the model, values between 3–7 indicate that the model has considerably less support, while values greater than ten indicate that the model is very unlikely (Burnham and Anderson 2002). The Akaike weights ($w_i$) provide a more precise measure of the strength of evidence and can be interpreted as the probability that the model is the best among the whole set of candidates. In addition, the ratio of two AIC weights indicates how much more likely one model is better compared to the other. Clearly, these metrics are more informative than a simple hypothesis test that can only pass or fail a model based on an arbitrary significance level without providing any strength of evidence or ranking. These advantages will become more apparent as we apply the metrics to our data in the following sections. For more information about the Akaike criterion, refer to (Burnham and Anderson 2002).

As seen from Figure 3-11, the lognormal distribution provides a superior fit to the amplitude distribution in the first and the second rays. Furthermore, Table 3-3 gives a comparison of the amplitude distribution fitting models for the chest-to-right-waist link and the AIC parameters show further that the lognormal distribution gives the best fitting result. For the chest-to-right-waist link, the average standard deviation of the amplitude variation is 7.87 dB, which will be used in the channel modeling. For other representative links, the conclusion that the lognormal distribution provides a superior fit to the amplitudes still holds, although naturally the standard deviations are different. The physical meaning of the amplitude variation following a lognormal distribution is easily understood, since the wave can be
considered as propagating along the body area with amplitudes which are affected by statistically varying reflection and diffraction coefficients.

Figure 3-11 Cumulative distributions of multi-path amplitudes
Table 3-3 Comparison of fitting models of amplitude distribution for the chest-to-right-waist link

<table>
<thead>
<tr>
<th>Model</th>
<th>$\Delta$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lognormal</td>
<td>0</td>
<td>0.88</td>
</tr>
<tr>
<td>Rayleigh</td>
<td>173.75</td>
<td>0.0</td>
</tr>
<tr>
<td>Rice</td>
<td>14</td>
<td>0.0</td>
</tr>
<tr>
<td>Weibull</td>
<td>4</td>
<td>0.12</td>
</tr>
</tbody>
</table>

3.3.5 Arrival time of the first ray

For various transmission links, the arrival time of the first multi-path ray varies, and is determined mainly by the direct transmission distance. For all five transmission links, the Gamma distribution is fitted to the arrival time data of the first ray, in accordance with the Akaike criterion. The Gamma fittings of the arrival time of the first ray for the five links are shown in Figure 3-12. For example, for the chest-to-right-waist link, the mean value and the standard deviation of the Gamma distribution are found to be 2.0 ns and 0.03 ns respectively. The extremely small standard deviation suggests that fixing the arrival time of the first ray at the mean value is reasonable when modeling the transmission links.

3.3.6 Inter-ray delay distribution

An impulse response can be represented by a tapped delay line model. In the on-body UWB transmission, adjacent taps may be influenced by a single physical multi-path component, which suggests a correlation. In this case, it is possible to realize an impulse response based on a uniformly-spaced tapped delay line model (Molisch et al. 2006). In the present study, however, the dominant multi-path components correspond to parts of the body such as the arms, legs and so on due to their varying positions. Most of the multi-path rays are distinguishable due to the high time resolution of 0.14 ns in our simulation. Moreover, the simulation results indicate that the correlation coefficient between the first and the second distinguishable multi-path rays is as weak as 0.2, which allows us to characterize the inter-ray delay without resorting to uniformly spaced taps.

The inter-ray delay, which corresponds to the temporal delay between two successive rays, represents the characteristics of the arrival time for all multi-path rays in the on-body
Figure 3-12 Cumulative distributions of the arrival time of the first ray

Arrival time of the first ray [ns]

Chest-to-right-chest link

Chest-to-left-ear link

Chest-to-left-waist link

Chest-to-right-waist link
transmission channel. In view of the above observations, we attempted to derive the statistical model for the inter-ray delay. Therefore, we identified the delay time of each ray from the corresponding peak in the power delay profile, after which we calculated the difference between the arrival times of two successive rays in order to obtain the inter-ray delay. The inter-ray delay data obtained in this way were fitted to some candidate statistical distributions, such as the exponential distribution, the Weibull distribution, the lognormal distribution and the inverse Gaussian distribution.

The results of the inverse Gaussian fitting for the inter-ray delay data are shown in Figure 3-13 for all five links. The inverse Gaussian distribution provides a superior fit. Furthermore, Table 3-4 gives a comparison of the two metrics of the second-order $AIC_c$ for the results of the fitting to the chest-to-right-waist FDTD-calculated data. The second-order $AIC_c$ indicates further that the inverse Gaussian distribution is the best-fitting model. For the chest-to-right-waist link, the mean value of the inter-ray delay is 0.33 ns and the standard deviation is 0.2 ns. For other representative links, the inverse Gaussian distribution also provides a superior fit in comparison to other distributions, although the mean values and the standard deviations are, of course, different.

In probability theory, the inverse Gaussian distribution (also known as the Wald distribution) is a two-parameter family of continuous probability distributions with support on $(0, \infty)$. Its probability density function is given by

$$f(x, \mu, \lambda) = \left[ \frac{\lambda}{2\pi x^3} \right]^{\frac{1}{2}} \exp \left\{ -\frac{\lambda(x-\mu)^2}{2\mu^2x} \right\}$$  \hspace{1cm} (3-27)$$

for $x > 0$, where $\mu > 0$ is the mean and $\lambda > 0$ is the shape parameter.

As $\lambda$ tends to infinity, the inverse Gaussian distribution becomes more like a normal (Gaussian) distribution. The inverse Gaussian distribution has several properties analogous to a Gaussian distribution. The name can be misleading. It is an "inverse" only in that, while the Gaussian describes the distribution of distance at fixed time in Brownian motion, the inverse Gaussian describes the distribution of the time a Brownian motion with positive drift takes to
Figure 3-13 Cumulative distributions of inter-ray delay
reach a fixed positive level. This physical explanation shows the time distribution characteristic of the inverse Gaussian distribution and verifies further the fact that inverse Gaussian distribution fits to the inter-ray delay data.

Table 3-4 Comparison of fitting models of inter-ray delay distribution

<table>
<thead>
<tr>
<th>Model</th>
<th>$\Delta$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inverse Gaussian</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>Lognormal</td>
<td>604</td>
<td>0.0</td>
</tr>
<tr>
<td>Weibull</td>
<td>792</td>
<td>0.0</td>
</tr>
<tr>
<td>Exponential</td>
<td>852</td>
<td>0.0</td>
</tr>
<tr>
<td>Gamma</td>
<td>718</td>
<td>0.0</td>
</tr>
</tbody>
</table>

3.3.7 Summary of the derived model parameters

Based on the above characterizations and parameterizations, we have determined all statistical models and parameters required for the construction of the channel model. These are summarized in Table 3-5.

Table 3-5 Model parameters for five representative transmission links

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Description</th>
<th>Characteristics</th>
<th>Right ear</th>
<th>Left ear</th>
<th>Right chest</th>
<th>Left waist</th>
<th>Right waist</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\gamma$  [ns]</td>
<td>Time constant for ray power decay</td>
<td>Exponential law</td>
<td>0.38</td>
<td>0.26</td>
<td>0.21</td>
<td>0.30</td>
<td>0.47</td>
</tr>
<tr>
<td>$\sigma$ [dB]</td>
<td>Standard deviation of amplitude distribution</td>
<td>Lognormal</td>
<td>7.5</td>
<td>12.56</td>
<td>15.6</td>
<td>8.46</td>
<td>7.87</td>
</tr>
<tr>
<td>$\tau_0$ [ns]</td>
<td>Average arrival time of first ray</td>
<td>Constant</td>
<td>1.05</td>
<td>0.92</td>
<td>0.68</td>
<td>1.89</td>
<td>2.01</td>
</tr>
<tr>
<td>$\tau_0 - \tau_{t-1}$ [ns]</td>
<td>Distribution of inter-ray delay</td>
<td>Inverse Gaussian</td>
<td>$\mu_0 = 0.30$</td>
<td>$\mu_1 = 0.56$</td>
<td>$\mu_2 = 0.37$</td>
<td>$\mu_3 = 0.38$</td>
<td>$\mu_4 = 0.33$</td>
</tr>
<tr>
<td>$\lambda_0$</td>
<td>Mean power gain of first ray</td>
<td>$\Omega_0 = \mu \gamma$</td>
<td>-60.7</td>
<td>-62.1</td>
<td>-53.3</td>
<td>-71.5</td>
<td>-69.9</td>
</tr>
</tbody>
</table>

In Table 3-5, according to (Saleh and Valenzuela 1987), the mean power gain $\Omega_0$ of the first ray is related to $G$ (the reciprocal of the path loss $PL_0$) as $\Omega_0 = \mu G/\gamma$, where $\mu_0$ is the mean time interval between two rays.
3.4 Measurement Validation

In order to verify the computational results and the statistically constructed model result, measurement was carried out in a full anechoic chamber. The chest-to-right-waist transmission link was taken as the measurement object. Figure 3-14 presents a view of the measurement. Two small-size, low-profile Skycross SMT-3TO10M (Skycross, Melbourne, FL [Online]. Available: http://www.skycross.com) UWB antennas were mounted on the body surface. The transmitting antenna was fixed on the left side of the chest, and the receiving antenna was fixed on the right side of the waist. The measurement was conducted for 8 persons, and for 10 body postures for each person, including standing, walking and sitting.

![Measurement antennas mounted on the body](image)

3.4.1 Measurement setup

The measurement method was as follows:

1) The S21 parameter for the two antennas on the body was measured by using an Agilent E5071B network analyzer, where the S21 is the frequency domain transfer function.
2) The measured frequency domain transfer function was converted to the time domain by using an inverse Fourier transform.

3) Based on Equation (3-21) and Equation (3-22), the impulse responses and the average power delay profile were computed. The average power delay profile was derived from the average over 80 readings (8 persons, 10 body postures each).

4) All modeling parameters were extracted from the impulse response and the average power delay profile.

3.4.2 Measurement results

By using the approach described in the previous section, we performed a characterization and parameterization of the measurement results for the chest-to-right-waist transmission link. Table 3-6 presents a comparison of the parameters obtained from the FDTD modeling and the measurements. It is clear that good agreement has been achieved, even though it appears that the $\lambda_r$ values of the inverse Gaussian distribution of the inter-ray delay are somewhat different. In fact, $\lambda_r$ refers to the shape of the probability density function (PDF) of the inverse Gaussian distribution, and in spite of the twofold difference between the values, there is almost no significant difference between the shapes of the respective probability density functions. In short, the model parameters are in good agreement with the parameters obtained from the experiment, which proves the validity of this modeling method.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Model</th>
<th>Measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\gamma$ [ns]</td>
<td>0.47</td>
<td>0.41</td>
</tr>
<tr>
<td>$\sigma$ [dB]</td>
<td>7.87</td>
<td>8.87</td>
</tr>
<tr>
<td>$\tau_0$ [ns]</td>
<td>2.01</td>
<td>-</td>
</tr>
<tr>
<td>$\tau_s - \tau_{s-1}$ [ns]</td>
<td>$\mu_s=0.33$</td>
<td>$\mu_s=0.30$</td>
</tr>
<tr>
<td></td>
<td>$\lambda_s=0.85$</td>
<td>$\lambda_s=2.14$</td>
</tr>
</tbody>
</table>
3.5 Channel Model Implementation

3.5.1 Implementation method

One cluster was observed in all five transmission links as a means of visualizing the average power delay profile as derived in the previous section. Therefore, based on the modified Saleh-Valenzuela model and the channel characterization, a discrete time impulse response function applied to these five transmission links is written as follows:

\[ h(t) = \sum_{k=0}^{K} \alpha_k \delta(t - \tau_k) \]  

(3-28)

where \( \alpha_k \) is the multi-path gain coefficient, and \( \tau_k \) is the delay of the \( k \)-th multi-path component relative to the arrival time of the first ray.

First, the time delays of the multi-paths are induced as follows: the first ray is induced at a fixed arrival time, after which a temporal delay between two successive rays is created according to the inverse Gaussian distribution and added to the arrival time of the previous ray.

Next, the gain coefficient for each ray is defined as follows:

\[ \alpha_k = p_k \beta_k \]  

(3-29)

where \( p_k \) takes a value of +1 or -1 with equal probability, and

\[ 10 \log_{10}(\beta_k^2) \sim \text{Normal}(\mu_k, \sigma^2) \]  

(3-30)

since \( \alpha_k \) belongs to a lognormal distribution. From Eq. (3-23), we have

\[ E[\beta_k^2] = \Omega_0 e^{\frac{\tau_k - \tau_0}{\gamma}} \]  

(3-31)

and \( \mu_k \), the mean in Eq. (3-29), is written as

\[ \mu_k = \frac{10 \ln(\Omega_0) - 10 \tau_k/\gamma - \sigma^2 \ln(10)}{\ln(10)} - \frac{\sigma^2 \ln(10)}{20} \]  

(3-32)

3.5.2 Implemented model

All parameters required for the modeling of this simplified impulse response function have already been described in the previous section, and are summarized in Table 3-5.
The propagation models for all five transmission links were implemented in Matlab, and Figure 3-15 shows the implemented impulse response sample for the chest-to-right-waist link with $\Omega_0 = 1$ for the sake of simplicity.

### 3.5.3 Model validation

We can compare the channels which have been computed and statistically modeled using the root mean square (RMS) delay spread $\sigma_\tau$ and the mean excess delay $\tau_m$, which measure the effective duration of the channel impulse response. They are two representations of the impulse response profile and frequently used to verify channel models. The two metrics are defined as follows (Hashemi 1993):

\[
\sigma_\tau = \sqrt{\frac{1}{P_R} \int_0^\infty (\tau - \tau_m)^2 p(\tau) d\tau}
\]

(3-33)

\[
\tau_m = \frac{1}{P_R} \int_0^\infty \tau^2 p(\tau) d\tau
\]

(3-34)

where $P_R$ is the multi-path mean power. The above expressions show that $\sigma_\tau$ is defined as the square root of the second central moment of the power delay profile and $\tau_m$ is defined as the first moment of the power delay profile. $\sigma_\tau$ is a good measure of multipath spread and it gives an indication of the potential for inter-symbol interference.

Figure 3-16 and 3-17 show the cumulative density function (CDF) distributions of the RMS delay spread and mean excess delay for two receiving points. As can be seen from the figures, the model matches closely the FDTD-computed results, and therefore adequately characterizes the transmission link.
Figure 3-15 Sample impulse responses
Figure 3-16 Comparison of the FDTD-derived and modeled RMS delay spread distribution

Figure 3-17 Comparison of the FDTD-derived and modeled excess delay distribution
3.6 Conclusion

We have proposed a UWB channel model for some representative on-body transmission links, which is based on the frequency-dependent FDTD numerical method in which various body postures are taken into account. The derived channel model is a discrete impulse response model which is convenient for characterizing the multi-path effect. By virtue of a high time resolution in the numerical simulation and concluded low correlation between rays, this impulse response model is implemented based on distinguishable multi-path rays without resorting to a uniformly-spaced tapped delay line model. Based on the data derived from the FDTD method, we have provided a small set of parameters for implementing the channel model in order to generate discrete time impulse responses. Furthermore, the statistically implemented model and the data derived from the FDTD method in terms of key communication metrics are in good agreement, which demonstrates the validity of the channel model. Moreover, measurements were also conducted, and the results provided additional verification of the FDTD-derived data and the modeling results.

A reasonable channel model is of utmost importance for system design and testing, as it estimates the system performance. The proposed on-body UWB channel model can contribute to the standardization of body area networks such as the IEEE 802.15.6, which is now in the progress for medical applications. In practice, however, the channel model should include the effects of the surrounding environment such as floors, walls and so on. A complete on-body UWB channel model can be regarded as a combination of two parts. At first, the model generates components which diffract around and reflect from parts of the body, as proposed in this study, after which it adds additional components from the surrounding environment, as modeled in (Molisch et al. 2006).
Chapter 4 Communication Performance

4.1 Introduction

On-body communication performances are mainly decided by the multi-path propagation channel. As shown in Chapter 3, the multi-paths in on-body communication are essentially induced by the reflection due to movement of one or more parts of the body. The multi-path propagation channel makes electromagnetic waves traveling along the body reflected, diffracted, and scattered, so that a received signal ends up being the superimposition of several attenuated, delayed, and eventually distorted replicas of a transmitted waveform. System performance is therefore in fact significantly degraded by the distortion of pulses due to propagation over such a real channel. On the other hand, the performance degradation can be mitigated if a detailed characterization of the multi-path channel is available at the receiver. Chapter 3 provides detailed parameterization for 5 typical multi-path channels on upper body, which makes possible the simulation and analysis of communication performances.

In this chapter, the communication performances of PPM-TH-UWB with and without the RAKE receiver are analyzed and evaluated in the multi-path-affected chest-to-right-waist channel. We will first determine the maximum data rates under maximum probability of error constraints. And then evaluate the communication performances with and without the RAKE receiver under the maximum data rates. Finally, based on the path loss model derived in Chapter 3, we will determine the maximum communication distances at the given maximum data rates.
4.2 Analysis Setup

4.2.1 UWB radio signal and modulation scheme

The most common and traditional way of emitting an UWB signal is by radiating pulses that are very short in time. This transmission technique goes under the name of impulse radio (IR). For UWB-IR, the way by which the information data symbols modulate the pulses may vary; PPM and pulse amplitude modulation (PAM) are commonly adopted modulation schemes (Welborn 2001; Guvenc and Arslan 2003). In addition to modulation and in order to shape the spectrum of the generated signal, the data symbols are encoded using pseudorandom or pseudonoise (PN) codes. In a common approach, the encoded data symbols introduce a time dither on generated pulses leading to the so-called time-hopping UWB (TH-UWB).

In this research, TH-UWB combined with binary PPM (binary PPM-TH-UWB or 2PPM-TH-UWB) is adopted for the communication performances analysis. Even orthogonal frequency division multiplexing and multi-carrier code division multiple access are also capable of generating UWB signals at appropriate data rates, they are out of the scope of this thesis and we do not go into detail for this category.

The 2PPM-TH-UWB signal can be schematized to be generated as Figure 4-1 (Benedetto and Giancola 2004b), which represents the transmission chain. The binary information data is encoded by code repetition coder and transmission coder, and then modulated by PPM modulation, and then transmitted via the antenna as a pulse train.

![Figure 4-1 Transmission scheme for a PPM-TH-UWB signal](image)
4.2.2 Emission masks and power limits

As described in preceding chapters, the extensively approvable emission mask for UWB radio communications is issued by the FCC. It imposes limit on the emitted radiation. With the emission mask which is not allowed to exceed -41.3 dBm/MHz from 3.1 GHz to 10.6 GHz, the maximum allowed total power is derived approximately as 0.55 mW in Section 4.2. We will use the maximum allowed transmit power as the worst-case prerequisite in the communication performance analysis.

4.2.3 Noise characteristics

Decision at the receiver is based on the observation of a received energy $E$ over a finite time interval, which is composed of mainly two terms: a signal term $E_r$ and a noise term $E_{noise}$. The noise term may include several independent noise sources such as thermal noise, multi-user interference, external interference, and so on. Let us assume that the only noise source at the receiver is additive white Gaussian noise (AWGN). This noise is typically thermal, introduced by the circuitry of the receiving antenna and the receiver. The bilateral thermal noise PSD $N_0/2$ expressed in Joule, that is, W/Hz is given by:

$$\frac{1}{2} N_0 = \frac{1}{2} k T_s(f) = \frac{1}{2} k (T_a + (F(f) - 1) T_0) \quad (4-1)$$

Where $T_s$ is the so-called spot noise temperature at the receiver, that is, referring to a particular spot or frequency in the spectrum; $T_a$ is the receiving antenna temperature; $T_0 = 290K$ (17°C or 62.3°F) is a standard temperature; $k = 1.38 \times 10^{-23} J/K$ is the Boltzmann constant; and finally, $F(f)$ is the so-called spot noise figure of the receiving device.

In the case of narrow-band transmissions, $F(f)$ is usually averaged over the frequencies of interest and the average noise figure $F$ is used instead of $F(f)$:

$$\frac{1}{2} N_0 = \frac{1}{2} k T_s = \frac{1}{2} k (T_a + (F - 1) T_0) \quad (4-2)$$
Even averaging $F(f)$ over the frequencies of interest for UWB systems might not be significant due to the very large bandwidth of the signal, given the current absence of UWB noise figure models, we assume a noise figure narrow-band model. Since the receiving device is adomed on the human body, $T_0$ is here adopted as 300 K. And furthermore, we introduce a reasonable hypothesis that $T_a = T_0 = 300$ K.

4.2.4 Analysis channel

Chapter 3 describes the on-body channel model implementation and gives a set of model parameters for five typical transmission channels on the upper body. In this chapter, we take the chest-to-right-waist as the objective channel in which the communication performances analysis is carried out. It is reasonable to pick it up since it can represent the maximum propagation distance on the front of the upper body, where worst-case hypothesis can hold.

Figure 4-2 shows 50 samples of the impulse response realizations for chest-to-right-waist channel. The dominating multi-paths range from 5 to 6.

![Figure 4-2](image)

Figure 4-2 50 samples of the impulse response realizations for chest-to-right-waist channel
4.3 Maximum Data Rate

The different signal paths between a transmitter and a receiver correspond to different transmission times. For an identical signal pulse from the transmitter, multiple copies of signals are received at the receiver at different moments. The signals on shorter paths reach the receiver earlier than those on longer paths. The direct effect of these asynchronous arrivals of signal causes the spread of the original signal in time domain. This delay spread puts a constraint on the maximum transmission capacity on the UWB on-body channel, which also means data rates are therefore restricted. Specifically, if the period of baseband data pulse is smaller than that of delay spread, inter-symbol interference (ISI) will be generated at the receiver. That is, the data signals on two neighboring pulse periods are received at the same time, which causes the receiver not to be able to distinguish them. This means the error probability performance depends on the data rate in multi-path-affected channel. The higher the data rate, the more the error bits. Therefore, a maximum data rate can be concluded at a given BER requirement.

On the other hand, PPM modulation is inherently sensitive to multi-path-affected channel. Since the information is encoded in the time of arrival, the presence of one or more paths can make it difficult, if not impossible, to accurately determine the correct pulse position corresponding to the transmitted pulse.

4.3.1 Conventional correlation receiver

The correlation receiver has been typically brought forward for the multi-path-free AWGN channel. In this case, the correlation receiver is called optimal receiver, which consists of a multiplication module and an integrator. In a digital system, the integrator would then be followed by a threshold detector.

By the correlation receiver in the multi-path-free AWGN channel, the average BER for binary PPM is expressed as follows:
\[ Pr_b = \frac{1}{2} \text{erfc} \left( \frac{E_b}{\sqrt{2N_0}} \right) \]  \hspace{1cm} (4-3)

where \( E_b \) is the received energy per bit and

\[ \text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{+\infty} e^{-t^2} \, dt \]  \hspace{1cm} (4-4)

Figure 4-3 shows average BER as a function of \( E_b/N_0 \) under multi-path-free AWGN channel.

![Figure 4-3 Average BER for binary PPM under multi-path-free AWGN channel](image)

**4.3.2 Data rate requirements**

The RMS delay spread for the chest-to-right-waist channel is summarized in Table 4-1. As can be seen, the RMS delay spread is around 0.47 ns. The delay spread will put a constraint on the data rate. When the data rate goes up, the error bits will increase and the BER performance will deteriorate. We simulated the BER performance with increasing data rates using the conventional correlation receiver described in 5.2.1. Table 4-2 shows the simulation specifications setup. Most of the parameters are taken from IEEE 802.15.4a (Available on:
http://www.ieee802.org/15/pub/TG4a.html). More than 10 discrete impulse response functions were taken out from Figure 4-2 for the BER simulation to have an average performance since the BER result depends on the particular realization of the channel impulse response.

Table 4-1 FDTD-derived and modeled RMS delay spread

<table>
<thead>
<tr>
<th>RMS delay spread</th>
<th>FDTD-derived</th>
<th>Modeled</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mean</td>
<td>0.469</td>
<td>0.470</td>
</tr>
<tr>
<td>Standard deviation</td>
<td>0.145</td>
<td>0.177</td>
</tr>
</tbody>
</table>

Table 4-2 Data rate simulation specifications

<table>
<thead>
<tr>
<th>UWB Frequency Band</th>
<th>3.1-10.6 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pulse</td>
<td>2nd derivative Gaussian pulse</td>
</tr>
<tr>
<td>Chips Per Bit</td>
<td>16</td>
</tr>
<tr>
<td>Modulation Scheme</td>
<td>PPM-TH</td>
</tr>
<tr>
<td>Receiver</td>
<td>Correlation Receiver</td>
</tr>
<tr>
<td>Detector Scheme</td>
<td>Synchronization Detection</td>
</tr>
</tbody>
</table>

Figure 4-4 BER vs. data rate performance with correlation receiver
Figure 4-4 gives the BER performance with the increasing of data rate under the FCC permissible maximum UWB transmit power. With fixed chip number of per bit, different data rate will have different bit period. The energy per bit $E_b$ will also vary with a fixed transmitting power so that the corresponding $E_b/No$ in Figure 4-4 ranges from 13 to 15 dB. As expected, when the data rate goes up to 100 Mbits/s, the BER becomes very bad.

In the error correction theory, the probability of error as 0.01 and 0.001 is usually taken as a criterion threshold in which the error correction code can work effectively. For example, in (Bic et al. 1991) it is shown that the intersection point of BER with coding and BER without coding, i.e., the point of coding gain equal to zero, is between $BER = 0.01$ and $BER = 0.001$ for forward error correction. It is therefore reasonable to use 0.01 and 0.001 as indices. That is to say, as long as the BER is not worse than 0.01 or 0.001, the error bits can be detected and corrected using specific error correction codes. In turn, if the error probability is very high surpassing the threshold, the error correction coding might not be practical.

Based the error correction theory, we can conclude that the maximum desirable data rates for chest-to-right-waist channel are 70 Mbits/s and 90 Mbits/s, which correspond the BER threshold of 0.001 and 0.01 respectively.

Under the two maximum data rates, the on-body BER performance with correlation receiver is given in Figure 4-5 to compare with the ideal performance under Gaussian channel. It seems that the performance of the correlation receiver is not so satisfactory in the multi-path-affected on body channel since the BER performance deteriorated badly. Actually, it is not appropriate for multi-path-affected on body channel since its structure foresees the presence of a correlator that is matched to one single waveform, while in the multi-path-affected on-body channel multiple signal superposition is not evitable. Advanced receiver structure might be needed to improve the communication performance. Yet, the simple structure of correlation receiver still keeps an advantage to on-body communications.
4.3.3 Conclusion

By conventional correlation receiver, the maximum desirable data rates for chest-to-right-waist channel are 70 Mbits/s and 90 Mbits/s which correspond to the BER threshold of 0.001 and 0.01 respectively. At the maximum data rates of 70 Mbits/s and 90 Mbits/s, the receiver performance of correlation receiver seems not so satisfactory. Actually, it is not appropriate for multi-path-affected on-body channel since its structure foresees the presence of a correlator that is matched to one single waveform.
4.4 BER Performance

4.4.1 RAKE receiver

The presence of multiple paths between transmitter and receiver demands improvement in receiver structure. The optimum receiver for the AWGN channel seems not appropriate for the on-body multipath channel since its structure foresees the presence of a correlator that is matched to one single waveform. As it should be, the performance degradation can be mitigated if a detailed characterization of the multi-path-affected channel is available. UWB-IR systems can take advantage of multi-path propagation by combining different and independent replicas of the same transmitted pulse. In this case, the receiver exploits "temporal diversity" of the multi-path channel to improve performance of the decision process.

Different strategies for exploiting diversity can be adopted by the receiver: selection diversity (SD), equal gain combining (EGC), and maximal ratio combining (MRC). With the SD method, the receiver selects the multi-path contribution exhibiting the best signal quality and operates the decision on the transmitted symbol based on the observation of this contribution only. With the EGC method, the different contributions are first aligned in time and then added without any particular weighting. In MRC, the different contributions are weighted before the combination and the weights are determined to maximize the signal noise ratio (SNR) before the decision process. In the presence of Gaussian noise at the receiver, the SNR is maximized by applying a weighting factor that is proportional to the amplitude of the corresponding received signal to each multi-path contribution.

For different diversity strategies described above, the receiver takes advantage of multi-path under the hypothesis that different replicas of the same transmitted pulse can be analyzed separately and eventually combined before decision. The optimum correlator for the multi-path-affected channel must include additional correlators associated with different replicas of a same transmitted waveform. Such a scheme was invented by (Price and Green 1958), and is called the RAKE receiver. Figure 4-6 shows the structure of the RAKE receiver, which
consists of a parallel bank of \( N_R \) correlators, followed by a combiner that determines the variable to be used for the decision on the transmitted symbol. Each correlator is locked on one of the different replicas of the transmitted symbol. \( \{m_1(t), m_2(t), ... m_R(t)\} \) are the correlator masks which are the delayed replicas of the transmitted symbol. \( \{\omega_1, \omega_2, ... \omega_R\} \) are the weighting factors which depend on the diversity method implemented at the receiver. In

\[
\begin{align*}
\text{r}(t) & \xrightarrow{\text{Integrator}} \omega_1 \\
\text{m}_1(t) & \xrightarrow{\text{Integrator}} \omega_2 \\
\text{m}_2(t) & \xrightarrow{\text{Integrator}} \omega_R \\
& \xrightarrow{\text{Detector}} \text{Estimated Symbol}
\end{align*}
\]

Figure 4-6 RAKE receiver with \( N_R \) parallel correlators

the SD case, the weighting factors are equal to zero, except for the factor on the branch corresponding to the signal with highest amplitude, which is equal to one. In the case of EGC, all factors are equal to 1, that is, the combiner simply adds the outputs of the correlators without applying any weighting. Finally, in the MRC case, the output of each branch is multiplied by a weighting factor, which is proportional to the signal amplitude on that branch.

Since on-body UWB channel is modeled with a discrete time impulse response, the RAKE scheme can be greatly simplified, shown in Figure 4-7. The correlator integrates the product between no-delayed correlation mask \( m(t) \) and the received waveform \( r(t) \). The output of the correlator is sampled with period \( \Delta t \) before passing through a delay unit and a combiner, which implements one of the previously described diversity methods: SD, EGC, or MRC.

The MRC diversity method is employed in this study for the communication performance evaluation. In the RAKE receiver the MRC method has the maximum signal-to-noise power
ratio since different multi-math contributions are weighted before the combination by a weighting factor that is proportional to the amplitude of the corresponding received signal.

The adoption of a RAKE considerably increases the complexity of the receiver. This complexity increases with the number of multi-path components analyzed and combined before decision, and can be reduced by decreasing the number of components processed by the receiver. However, a reduction of the number of paths leads to a decrease of energy collected by the receiver. A quasi-analytical investigation of the existing tradeoff between receiver complexity and percentage of captured energy in a RAKE receiver for UWB-IR systems is presented in (Win and Scholtz 1998).

Based on the FDTD-calculated results, we calculate the probability of multi-paths and also the captured energy of multi-paths. Figure 4-8 shows the probability versus the number of multi-paths. Four effective multi-paths have the highest probability in all of the possible multi-paths as shown in Figure 4-8. Figure 4-9 shows the energy percentage captured by multi-paths. About 80% of the received energy is captured by the first 2 multi-paths, and 92% of the received energy is captured by the first 4 multi-paths. Therefore, 2 and 4 fingers RAKE receivers, which correspond to the first two and four multi-path components, are employed in the simulation scheme.
4.4.2 BER and data rate

The RAKE receiver is supposed to improve the BER performance since it takes advantage of multi-paths to enhance the SNR. We carry out the simulation for different data rates given that perfect channel knowledge for the fingers is assumed in the RAKE receiver.

![Probability versus number of multi-paths](image1)

**Figure 4-8** Probability versus number of multi-paths

![Energy percentage captured by multi-paths](image2)

**Figure 4-9** Energy percentage captured by multi-paths
Figure 4-10 and Figure 4-11 show the BER performances with 2 and 4 fingers RAKE receivers under the two maximum data rates 70 Mbits/s and 90 Mbits/s respectively. As expected, both the 2-finger and 4-finger RAKE receiver significantly improves the BER performance compared with conventional correlation receiver. At the data rate of 70 Mbits/s, the degradation in performance of the 2-finger and 4-finger RAKE receivers is less than 2 dB at a BER of 0.01 with respect to the ideal Gaussian noise channel, while at data rate of 90 Mbits/s, the performance degradation is less than 4 dB. The BER of 4-finger RAKE receiver is superior to 2-finger RAKE receiver. At a BER of 0.01, the performance degradations of 4-finger RAKE for 70 Mbits/s and 90 Mbits/s are respectively about 1 dB and 2 dB compared with the ideal Gaussian channel. The performance degradation of 2-finger RAKE compared with 4-finger RAKE is less 1 dB and 2 dB at BER of 0.01 for 70 Mbits/s and 90 Mbits/s respectively. These results demonstrate that both the 2-finger and 4-finger RAKE receiver are effective in the on-body multi-path channel and they can considerably improve the communication performance.

However, simplicity in the device structure is also one important issue which cannot be ignored. The complicated structure for 4-finger RAKE receiver may hinder its application in wearable body area communication. In this sense, the 2-finger RAKE receiver can be an option to compromise the communication performance and structure complication, since it can also get satisfactory performance and at the same time holds a relatively simple structure compared with 4-finger RAKE receiver.

4.4.3 Conclusion

The conventional correlation receiver is obvious here not appropriate in multi-path-affected on-body channel, while the 2-finger and 4-finger RAKE receiver considerably improve the BER performance in multi-path-affected on-body channel compared with the normal correlation receiver. At the data rate of 70 Mbits/s, the degradation in performance of the 2-finger and 4-finger RAKE receivers is less than 2 dB at a BER of 0.01 with respect to the ideal Gaussian noise channel, while at data rate of 90 Mbits/s, the performance
Figure 4-10 BER under data rate 70 Mbits/s

Figure 4-11 BER under data rate 90 Mbits/s
degradation is less than 4dB. Although the performance of 4-finger RAKE receiver is superior to 2-finger RAKE receiver, the 2-finger RAKE receiver might be more practical for body area communication since it can get a satisfactory performance and at the same time holds a relatively simple structure compared with 4-finger RAKE receiver.

4.5 Communication Distance

4.5.1 BER and distance

Figure 4-10 and Figure 4-11 describe the relationship between the $E_b/N_0$ values and the probability of error. Based on the path loss Equation (3-19) and the maximum allowed transmitted power Equation (4-1), the maximum receiving power can be obtained and therefore the $E_b$ term can be derived as a function of propagation distance. For noise power density $N_0$, Section 5.1.3 has given a detailed description. In this case, we can get the probability error as a function of propagation distance at a given data rate. These analysis and derivation procedures hold, given that the channel statistical characteristics will not be a big difference when the communication distance spreads to the whole body, which is basically reasonable.

Figure 4-12 and Figure 4-13 show the BER vs. communication distance under the maximum data rates of 70 Mbits/s and 90 Mbits/s respectively. At data rate of 70 Mbits/s, the maximum communication distance with correlation receiver is about 0.6 m while with RAKE receiver is about 0.8 m. At data rate of 90 Mbits/s, the distance with correlation receiver deteriorate heavily.

Table 4-3 gives the quantitative description of the communication distances at two threshold probability of error. At the given BER thresholds and the given data rates, the communication distance is around 0.6-0.8 m. It can be concluded that under the FCC emission spectral mask, the effective communication distance for UWB-PPM-TH with 2 and 4 fingers RAKE receiver is at least about 0.6-0.8 m on the human body since it is concluded
Figure 4-12 BER vs. distance under data rate of 70 Mbits/s

Figure 4-13 BER vs. distance under data rate of 90 Mbits/s
Table 4-3 Maximum communication distances under different data rates

<table>
<thead>
<tr>
<th>Communication distance [m]</th>
<th>70Mbits/s</th>
<th>90Mbits/s</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>2-finger</td>
<td>4-finger</td>
</tr>
<tr>
<td>BER=0.01</td>
<td>0.8</td>
<td>0.83</td>
</tr>
<tr>
<td>BER=0.001</td>
<td>0.64</td>
<td>0.72</td>
</tr>
</tbody>
</table>

under the maximum data rates. If a low data rate less than 70 Mbits/s and 90 Mbits/s is acceptable, the communication distance on human body could be longer than 0.8 m.

4.5.2 Communication distance and human safety

If the communication link is from a transmitter at the head to a receiver at the leg, the on-body communication distance will be about 1.6 m. In this case, twice the above distance may be required. It is obvious that a transmitted power under the FCC emission spectral mask is difficult to meet such a communication distance. From Figure 4-12 and Equation (3-19), it can be derived that the received power at 0.8 m is -83 dBm, which yields a BER of 0.01. Since the path loss for a 1.6m on-body transmission is 88.8 dB, a transmitted power of 5.8 dBm is required to guarantee the BER of 0.01 at this communication distance. This means an increase of 11.3 dB on the transmitter power. However, even so, as shown in the next chapter in detail, the SAR still has a margin of 18.7 dB (30dB-11.3dB) compared to the ICNIRP safety guideline, although the countermeasure for FCC emission spectral mask should be considered.

4.5.3 Communication distance and data rate

The BER vs. data rate performance in Figure 4-4 is based on the chest-to-right-waist link, in which the communication distance is 0.56 m and the corresponding path loss is 71.4 dB. If the communication link is from a transmitter at the head to a receiver at the leg, the on-body communication distance will be about 1.6 m and the path loss will be 88.7 dB as described in 4.5.2. So the path loss difference compared with 0.56 m is 17.3 dB (88.7 dB-71.4 dB). If the maximum transmitted power keeps not be raised, in order to keep the BER at 0.01 or 0.001, the data rate has to be decreased to compensate the energy loss due to the longer communication distance.
From Figure 4-10, it is concluded that at BER of 0.001, the improvement of EbNo due to 2-finger RAKE compared with correlation receiver is about 4 dB, and the improvement of EbNo due to 4-finger RAKE compared with correlation receiver is about 5.5 dB. So the compensation terms the data rate reduction should account for are 13.3 dB (17.3 dB - 4 dB) and 11.8 dB (17.3 dB - 5.5 dB) respectively for 2-finger RAKE and 4-finger RAKE.

\[ E_b[dB] = P_r[dB] + T_b[dB] \]  
(4-5) \[ \Delta T_b[dB] = 10 \log_{10}\left(\frac{DataRate_{max}}{DataRate_{reduced}}\right) \]  
(4-6)

As shown in Equation (4-5), the required improvement of EbNo will be put on the bit period \( T_b \) provided that the maximum received power \( P_r \) keeps not be raised. Therefore the reduced data rate can be derived based on Equation (4-6) where \( \Delta T_b \) means the compensation term the data rate reduction should account for and \( DataRate_{max} \) and \( DataRate_{reduced} \) represent the maximum data rate and the reduced data rate respectively. So the data rate will be reduced to 3.3 Mbits/s and 4.6 Mbits/s based on Equation (4-6) for 2-finger RAKE and 4-finger RAKE.

From Figure 4-11, similar process can be done at BER of 0.01 for original 90 Mbits/s. The improvement of EbNo due to 2-finger RAKE and 4-finger RAKE are about 8 dB and 10 dB read from Figure 4-11. So the compensation item will be 9.3 dB and 7.3 dB. The data rate will be 10.6 Mbits/s and 16.7 Mbits/s for 2-finger RAKE and 4-finger RAKE.

4.5.4 Conclusion

Under the FCC emission spectral mask, the effective maximum communication distance for UWB-PPM-TH with 2 and 4 fingers RAKE receiver is around 0.6-0.8 m on the human body since it is concluded under the maximum data rates. If the data rates are lower than the maximum data rates of 70 Mbits/s and 90 Mbits/s, the available communication distances could be longer than 0.8 m. On the other hand, even the communication link is cover the whole body and the corresponding required transmitted power exceeds the power limitation by 11.3 dB, the SAR still has a considerable margin compared to the ICNIRP safety guideline. Moreover, if the maximum transmitted power keeps not be raised, the data rate has to be
decreased to compensate the energy loss due to the longer communication distance. The whole body communication can still keep a data rate in the order of Mbits/s.

4.6 Conclusion

The maximum desirable data rates for upper body channel with correlation receiver are 70 Mbits/s and 90 Mbits/s which correspond to the BER threshold of 0.001 and 0.01 respectively. The 2-finger and 4-finger RAKE receiver considerably improve the BER performance in multi-path-affected on-body channel compared with the normal correlation receiver. At the data rate of 70 Mbits/s, the degradation in performance of the 2-finger and 4-finger RAKE receivers is less than 2 dB at a BER of 0.01 with respect to the ideal Gaussian noise channel, while at data rate of 90 Mbits/s, the performance degradation is less than 4 dB. The 2-finger RAKE receiver might be a good option for body area communication since it can get a satisfactory performance and at the same time holds a relatively simple structure compared with 4-finger RAKE receiver. The effective maximum communication distance for UWB-PPM-TH with 2 and 4 fingers RAKE receiver is around 0.6-0.8 m on the human body since it is concluded under the maximum data rates. On the other hand, if the communication link is needed to cover the whole body, the data rate has to be decreased to the order of several Mbits/s in order to compensate the energy loss due to the longer communication distance.
Chapter 5 Human Safety and EMC

5.1 Introduction

UWB wearable body area communications for medical applications inevitably brings forward human body safety and bio- electromagnetic compatibility (Bio-EMC) issues since it operates on human body. Wearable body area communication has communication appliances adorned on the human body which acts as a transmission medium. Electromagnetic emission from body worn communication devices may give rise to body tissue energy absorption as well as possible interference to implanted medical device. These issues have to be considered in the design of the BAN communication system, but almost no study has been done until now.

Because UWB wearable body area communications operate on human body, safety to human body has a higher priority than the other wireless communications. This is also one of the major issues to be addressed by IEEE 802.15.6 standard (Astrin et al. 2009). For wearable body area communications, the transmit power should be limited as low as possible in order to assure the safety to humans. It has been known that high intensities of RF radiation can be harmful to the human body due to the ability of RF energy to heat biological tissue rapidly. Exposure to high RF power can result in heating of the human body and an increase in body temperature. Under local exposure conditions, the SAR of about 1W/kg could result in measurable heating of biological tissue. Biological effects that result from heating of tissue by RF energy are often referred to as thermal effects. If the absorbed heat energy is greater than the heat released by the tissue into the environment, the temperature of the interior of the tissue will rise. Tissue damage can result primarily because of the body's inability to cope with or dissipate the excessive heat. When the body temperature of the exposed tissue rises...
from its normal value, important biological effects can occur. If the electromagnetic emission from body worn communication devices is over high, the SA and SAR may violate the human safety guideline which is used to ensure the human safety under electromagnetic exposure. For UWB wearable body area communications, although the signal from one UWB device is very low, however, it is unclear that the energy absorption will increase to what extent when many UWB devices are adorned simultaneously to a human body, which is the actual situation for a body area network. An analysis method is therefore required from the point of view of biological safety evaluation.

On the other hand, the electromagnetic (EM) interaction of on-body communication signals with the human body implies a potential for EM interference (EMI) to an implanted medical device such as a cardiac pacemaker. The cardiac pacemaker consists of a shielded housing with electronic circuits inside and an electrode. It is connected to the heart by an electrode to read the electrocardiogram (ECG) and to simulate the heart beat by voltage pulses if necessary. External EM fields can couple into the pacemaker to cause an interference voltage at the input of the internal sensing circuit. The induced interference voltage at the input of the sensing circuit of pacemaker will be amplified and low-pass filtered. When the output voltage of the amplifier and low-pass filter exceeds a threshold, the pulse voltage to simulate the heart beat may be triggered and a malfunction of cardiac pacemaker occurs. Extensive investigations for the EMI problem of pacemakers by mobile telephones have been conducted experimentally or analytically (Wang J.Q. et al. 2000), while any EMI prediction of on-body communication devices on pacemakers has not yet been carried out.

This chapter mainly consists of two parts which will elaborate the energy absorption for UWB pulse exposure and EMI modeling for cardiac pacemaker by UWB wearable body communications respectively. For the first issue, we will first propose two approaches in the time-domain and frequency-domain to calculate the SA and SAR for UWB pulse exposure. We first calculate the SA for a single UWB pulse exposure and then derive the SAR based on the UWB pulse radio technique. The 10-gram averaged SA and SAR are calculated and analyzed under single exposure and multiple exposures with multiple UWB devices adorned
on body simultaneously. Besides, the variation relationship of SA/SAR and the distance between the antenna and the human body are also concluded. For the second issue, we will propose a two-step approach to model the induced EMI voltage at the cardiac pacemaker by UWB on-body communication signals. In the first step, we calculate the input voltage of the pacemaker circuit using a full-wave EM field simulator by considering the pacemaker as a receiving antenna. In the second step, we employ a Volterra kernel description to predict the output voltage of the amplifier and low-pass filter circuit in the pacemaker for evaluating the EMI effect.

5.2 SA and SAR

5.2.1 UWB radio signal and emission mask

Currently, the only available and extensively accepted emission mask for UWB radio communications are those issued by the FCC in the United States as mentioned in Section 2.2. It imposes limit on the emitted radiation. As shown in Fig. 2-4, for the whole UWB frequency band, i.e., from 3.1 GHz to 10.6 GHz, the maximum emission power density is not allowed to exceed -41.3 dBm/MHz.

It is important to note that the FCC emission mask refers to a unilateral PSD \( P_M(f) = -41.3 \text{ dBm/MHz} \); therefore, the maximum allowed transmit power \( P_{M,\text{max}} \) for a signal occupying the whole UWB frequency band is:

\[
P_{M,\text{max}}[\text{dBm}] = 10 \log_{10} \int_{3.1 \times 10^3}^{10.6 \times 10^3} P_M(f) \, df
\]

\[
= -41.3 + 10 \log_{10}(10.6 \times 10^3 - 3.1 \times 10^3) \quad \text{(5-1)}
\]

\[
\approx -2.8 \text{dBm}
\]

i.e.

\[
P_{M,\text{max}} \approx 0.55mW
\]
The FCC regulation illustrates the fact that the UWB radio signal must meet the emission mask and the maximum allowed transmit power requirements. Signal power up to 0.55 mW can be considered as the worst case in simulation and analysis.

5.2.2 SA/SAR safety guidelines

The SA is defined as the energy absorbed per unit mass of biological tissue with unit of joule per kilogram and the SAR is defined as the power absorbed per unit mass of biological tissue with unit of watts per kilogram. SAR is a measure of the rate at which RF energy is absorbed by the body when exposed to radio-frequency electromagnetic field. SA and SAR are usually averaged either over the whole body, or over a small sample volume, typically 1g or 10g of tissue.

The IEEE and many national governments have established safety limits for exposure to various frequencies of electromagnetic energy based on SA/SAR, mainly based on the International Commission on Non-Ionizing Radiation Protection (ICNIRP) guideline (ICNIRP 1998) which can be accessed from the web site http://www.icnirp.org/documents/emf.gdl.pdf. This guideline is guarding against thermal damage. In ICNIRP guideline, the safety limits for 10g-averaged SA and SAR are 2 mJ/kg and 2 W/kg respectively in frequency band of 100 KHz-10 GHz for general public exposure. These are the extensively accepted guideline values for human safety.

5.2.3 Antenna and excitation

As a UWB antenna in the numerical calculation, we choose an elliptic disk dipole with a major axis of 25 mm diameter and a minor axis of 21 mm diameter, which yields a voltage standing wave ratio (VSWR) nearly 2.0 between 3.1 and 10.6 GHz. Figure 5-1 gives the antenna dimension picture and the VSWR plot. The antenna will be placed at 2mm from the surface of the body.
The pulse shape that can be generated in the easiest way by a real pulse generator and be radiated in an efficient way is considered as Gaussian derivatives. For a UWB pulse to be transmitted, we choose a 5th-derivative Gaussian pulse with a pulse width of nearly 500 ps. Figure 5-2 shows the employed 5th-derivative Gaussian pulse waveform and Figure 5-3 shows the equivalent isotropically radiated power (EIRP). EIRP is the amount of power that a theoretical isotropic antenna (that evenly distributes power in all directions) would emit to produce the peak power density observed in the direction of maximum antenna gain. The EIRP is derived as follows.

In general,

\[ EIRP = P_a \cdot G_a \]  \hspace{1cm} (5-2)

where \( P_a \) is the power supplied to the UWB antenna and \( G_a \) is the absolute gain of antenna. The determination of \( P_a \) depends on the UWB radio technique. For the UWB-IR, we can first calculate the antenna power density

\[ P_{SD}(f) = \text{Re}\left(\frac{V(f)V^*(f)}{Z_{in}(f)}\right) \]  \hspace{1cm} (5-3)

where \( V(f) \) is the Fourier transform of the UWB pulse voltage with unit of V/Hz, \( Z_{in}(f) \) is the antenna input impedance and \( T \) is the pulse duration. Then we have the EIRP as
\[ EIRP \left[ \frac{\text{dBm}}{\text{MHz}} \right] = 10 \log_{10} \left[ 2P_{SD}(f) \right] + 10 \log_{10} G_a + 90 \] (5-4)

Figure 5-2 5-th derivative Gaussian pulse waveform

Figure 5-3 FCC indoor emission spectrum mask and the equivalent isotropically radiated power

It can be found from Figure 5-3 that the maximum power density of the employed 5th-derivative Gaussian pulse is limited to \(-43.1 \text{ dBm/MHz}\) between 3.1 and 10.6 GHz, i.e., it meets with the FCC indoor emission spectrum mask.

Pulse with other order of Gaussian derivatives can also produce a power density that meets with the FCC UWB emission limit, i.e., less than the limitation of \(-43.1 \text{ dBm/MHz}\). However, the 5th-derivative Gaussian pulse is able to produce a fair good PSD that can closely
approximate the FCC emission mask, which is desirable in view of the excitation source requirement in the SA/SAR analysis. (Benedetto and Giancola 2004a) also brings forward the combination of the Gaussian pulse and its derivatives combination with different shape factor and differentiation order to adjust the PSD to the FCC masks. In a recent solution based on Gaussian waveforms, (Sheng et al., 2003) proposed an algorithm to select the best pulse differentiation order and shape factor values for fitting the mask in the bandwidth 3.1-10.6GHz. A limitation of this approach is in the difficulty of meeting the mask outside the above bandwidth by using a single derivative.

5.2.4 Analysis method and human model

The SA/SAR analysis method we employed is also based on the frequency-dependent FDTD numerical method which has been described in detail in chapter 3. For the human body model used for the SA and SAR analysis, we employ the same Japanese adult male model also from (Nagaoka et al. 2004). The difference lies in the fact that a homogeneous human body is employed with dielectric properties of the skin for the propagation channel modeling, while for SA/SAR analysis a heterogeneous human body model with 51 tissue types is required.

Under the FCC UWB limit, we perform the SA/SAR calculation for the anatomical human body model in two methods which are in time domain and frequency domain respectively. At first, we investigate the validation of the two methods for the SA calculation.

The first method is to calculate the current density in the time domain in the frequency-dependent FDTD method, i.e.,

\[ j(t) = \sigma_0 E(t) + \varepsilon_0 \frac{d}{dt} [\chi(t) * E(t)] \]  \hspace{1cm} (5-5)

Then we obtain the SA from

\[ SA = \int_0^T \frac{j(t)E(t)}{\rho} \cdot dt \]  \hspace{1cm} (5-6)

where \( \rho \) is the mass density. This calculation is straightforward in the frequency-dependent FDTD method, and it does not need additional calculation burden.
The second method is the frequency domain expression of the first approach. According to Parseval theorem,

\[
SA = \frac{\int_\omega^\omega |\sigma(\omega)|^2 E(\omega)^2}{\rho} \cdot df
\]

(5-7)

\[
\sigma(\omega) = -\omega \varepsilon_0 \text{Im}[\varepsilon_r(\omega)]
\]

(5-8)

where \( \sigma(\omega) \) is the lossy component of the dielectric properties and \( E(\omega) \) is the Fourier transfer of \( E(t) \). We get \( E(\omega) \) directly in the frequency-dependent FDTD method as a running summation at each time step based on the discrete Fourier transfer. This approach requires additional calculation burden.

Although the two methods are mathematically equivalent, the corresponding numerical algorithms are different. We carry out the SA calculation using these two methods. The UWB antenna is placed in the front of a cube. The cube has a length of 200 mm and the dielectric properties of muscle. Figure 5-4 shows the calculated SA profile taken from the front to the back of the muscle cube. As can be seen, both the first and the second methods give the same SA values. In view of the calculation efficiency, the first approach in the time domain is more appropriate to the SA calculation in the frequency-dependent FDTD method.

Comparison of the calculation results between the two approaches can confirm the validation of the time-domain algorithm in the SA and SAR calculations.

5.2.5 Results and discussion

We then use the first approach to first calculate the SA with the UWB antenna placed on the body at the chest, ear, eye and waist respectively. Figure 5-5 shows the antenna radiation pattern in the horizontal plane at a central UWB frequency of 5 GHz in the situation that the
UWB antenna is placed on the chest. It can be seen that the power radiates towards the front or the outside of the body. Similar radiation pattern can be also observed in other situations.

In the single exposure case, the UWB antenna is placed at one of the four locations, and in the multiple exposure case, two or four UWB antennas are simultaneously placed at two of or all of the four locations as shown in Figure 5-6. Figure 5-6 shows the SA distributions on the body surface for both single and multiple UWB pulse exposure. The 0 dB corresponds to 10 pJ/kg. As can be seen, the SA concentrates on very small area on the body surface near the
UWB antenna location, and the attenuation is more than 30 dB at a location 10 cm far from the UWB antenna. Especially, in the multiple-exposure situation the enhancement effect from neighbor UWB antenna is not obvious, which may be attributed to the rapid attenuation of the UWB signal on the body surface.

![Image](image.png)

Figure 5-6  SA distributions under the FCC UWB limit for various antenna locations
(0 dB corresponds to 10 pJ/kg)

For the UWB-IR pulse signals, the SAR can be simply obtained from the ratio of SA to the pulse duration $T$. That is, $\text{SAR} = \frac{\text{SA}}{T}$. It is reasonable since the excitation signal is in a form of successive pulses in UWB-IR. Of course UWB pulse should be repeated for communication usages. For SA, it is generally evaluated on single pulse. For SAR, however,
the values for single pulse and for pulse series are the same because the SAR is the ratio between $SA$ and $T$ for the above-mentioned modulation schemes. That is to say, the quantities for $SA$ and $T$ will increase simultaneously $N$ times for a series of $N$ pulses. It is therefore reasonable to use single pulse to investigate the SA and SAR.

Table 5-1 gives the peak SA, the ten-gram-averaged SA and the ten-gram-averaged SAR in all of the six situations as shown in Figure 5-6. As can be seen from Table 5-1, compared to single exposure, almost no obvious increase in the SA and SAR are observed in the multiple exposure cases. In order to verify this conclusion further, we observed the E-field at the chest under multiple exposure and the result is shown in Figure 5-7. Since the separation distance between other devices and the device at the chest is about 30 cm, the transmission time is at least 1 ns. It means that if the effect of signals from other devices is not negligible, there should be some significant components around 1 ns in the observed E-field. This does not depend on whether the devices are excited simultaneously or un-simultaneously. From Figure 5-7, no significant component is observed around 1 ns. The conclusion is therefore drawn that as long as the separation between two UWB devices is larger than 30 cm, the additional effect of signals from other devices is insignificant due to the rapid surface attenuation of UWB signals.

The ten-gram-averaged SA is in the order of $\text{pJ/kg}$ and is much smaller than the ICNIRP safety guideline of $2 \text{ mJ/kg}$. The ten-gram-averaged SAR is in the order of $\text{mW/kg}$ and is smaller than $1/2000$ of the ICNIRP safety guideline of $2 \text{ W/kg}$.

Besides, the energy absorbed by the whole body is found to be $0.01 \text{ pJ}$ which is about a quarter of the energy radiated from the antenna which is around $0.04 \text{ pJ}$.

The SA and SAR vary with the distance between the antenna and the human body. Table 5-2 shows the 10g-averaged SA and SAR corresponding to distances of 2 mm, 1 cm and 2 cm when one UWB antenna is adorned on the chest. It can be seen that the 10g-averaged SA and SAR values will decrease by $12 \text{ dB}$ and $17 \text{ dB}$ respectively when the distances become 1 cm.
Table 5-1  Ten-gram averaged SA and SAR under FCC UWB limit

<table>
<thead>
<tr>
<th>Antenna Location</th>
<th>Peak SA [pJ/kg]</th>
<th>10g SA [pJ/kg]</th>
<th>10g SAR [mW/kg]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Chest</td>
<td>7.00</td>
<td>0.473</td>
<td>0.946</td>
</tr>
<tr>
<td>Ear</td>
<td>2.17</td>
<td>0.037</td>
<td>0.074</td>
</tr>
<tr>
<td>Eye</td>
<td>13.26</td>
<td>0.268</td>
<td>0.536</td>
</tr>
<tr>
<td>Waist</td>
<td>7.39</td>
<td>0.232</td>
<td>0.464</td>
</tr>
<tr>
<td>Chest+Waist</td>
<td>7.40</td>
<td>0.474</td>
<td>0.948</td>
</tr>
<tr>
<td>Chest+Ear+Eye+Waist</td>
<td>13.27</td>
<td>0.476</td>
<td>0.952</td>
</tr>
</tbody>
</table>

and 2 cm compared with the initial distance of 2 mm. In practice, the antenna will be adorned as closely as possible to the body surface. Therefore, the 2mm case we calculated may represent the worst case. The further the distance becomes, the larger the safety margin will be.
5.2.6 Conclusion

Based on the frequency-dependent FDTD method, two analysis approaches have been proposed to calculate the SA and SAR for UWB pulse exposure. The two approaches have the same accuracy but the time-domain approach is more straightforward to the SA and SAR analysis for UWB-IR systems. We have also demonstrated the SA and SAR levels under the FCC UWB limit for various antenna locations on an anatomical human body model. The results have shown that the SA and SAR levels are much smaller than the ICNIRP safety limits and the multiple exposures do almost not obviously increase the SA and SAR as long as two UWB devices have a separation as large as 30 cm. The ten-gram-averaged SA is in the order of pJ/kg and is much smaller than the ICNIRP safety guideline of 2 mJ/kg. The ten-gram-averaged SAR is in the order of mW/kg and is smaller than 1/2000 of the ICNIRP safety guideline of 2 W/kg. Besides, the 10g-averaged SA and SAR values will decrease by 12 dB and 17 dB respectively when the distances become 1cm and 2 cm compared with the initial distance of 2 mm. The further the distance becomes, the larger the safety margin there will be.

5.3 EMI Modeling for Cardiac Pacemaker

A hybrid method, which consists of EM field modeling and electric circuit modeling, is used for EMI modeling for cardiac pacemaker in human body communication.

5.3.1 EM field modeling

Figure 5-8 shows a basic configuration for an implanted cardiac pacemaker. The cardiac pacemaker consists of a shielded housing with electronic circuits inside and an electrode as well as the lead wire. By considering the internal impedance seen from the connector to the internal circuit as a load, and the metal portions consisting of the pacemaker housing and the lead wire of the electrode as two elements of a receiving antenna, the resultant equivalent circuit for the pacemaker can be shown in Figure 5-9. Here, $Z_R$ is the radiation impedance of the pacemaker, $V_M$ is the open voltage induced between the pacemaker housing and the lead
wire due to the EM fields from the external communication devices, $Z_I$ is the internal impedance of the pacemaker seen from the connector, and $V_I$ is the voltage induced through the connector onto the internal circuit, which is referred here as the input interference voltage to the internal sensing circuit of pacemaker, respectively.

Figure 5-8 A basic configuration of cardiac pacemaker

Figure 5-9 Equivalent circuit for EMI pacemaker
The open voltage $V_M$ at the connector can be obtained by simulating the pacemaker as a receiving antenna with an open load at the connector using a full-wave EM field simulation tool such as the FDTD method. Figure 5-10 shows an FDTD model for a pacemaker user with an on-body communication device near the chest. This situation imagines a user certification case. The human body model is a homogeneous one with the dielectric properties of muscle and a spatial solution of 5 mm. The on-body transmitter is an electrode structure consisting of two metal plates on the human body surface. The pacemaker is modeled as a metal housing and a metal lead line of electrode. Using the FDTD method, the open voltage $V_M$ between the metal housing and the lead wire, i.e., at the connector, can be obtained when the transmitter electrode is excited. Then the input interference voltage $V_I$ of the sensing circuit can be obtained as

$$V_I = \frac{z_I}{z_R+z_I}V_M \quad (5-9)$$

Figure 5-10 A model for a pacemaker user with an on-body transmitter

5.3.2 Electric circuit modeling

Figure 5-11 shows the block diagram of the analogue sensing circuit of pacemaker (Schenke S. et al 2007). The input signal $V_I$ is amplified and low-pass filtered. The resulting output
voltage $V_o$ is compared with a sensing threshold $V_c$. When the voltage exceeds the sensing threshold, it will switch the pulse output, and then yield a malfunction of pacemaker.

![Analog Input Circuit for Sensing Signal](image)

**Figure 5-11** Block diagram of the internal pacemaker circuit

The amplifier and low-pass filter can be considered as an operational amplifier (opamp) combined with external elements. Figure 5-12 shows a general negative feedback opamp configuration. Although the on-body communication signal is at higher frequencies compared to the heart-beat-simulating pulse signal, due to the demodulating properties of the opamp, high frequencies can be down converted and are able to pass through the low-pass filter. Measured results in (Schenke S. et al 2007) also show the fact that the ideal linear opamp model is no longer applicable at higher frequency due to the demodulating properties of an opamp. A nonlinear model is therefore necessary for the opamp with low-pass filter character in order to precisely predict the interference output.

![General Negative Feedback Operational Amplifier Configuration with Differential Inputs](image)

**Figure 5-12** General negative feedback operational amplifier configuration with differential inputs

The Volterra Series is known as a powerful and accurate tool for weakly nonlinear analogue circuit analysis (Schetzen, M., 1980). According to Volterra series method, the output signal $y(t)$ of a nonlinear system can be expressed in the form
\[ y(t) = \sum_{i=1}^{+\infty} y^{(i)}(t) \]  

where

\[ y^{(i)}(t) = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} H_{k_1,\ldots,k_i}(f_1, \ldots, f_i) \times X_{k_1}(f_1) \cdots X_{k_i}(f_i) \exp{j2\pi(f_1 + \cdots + f_i)t} \, df_1 \cdots df_i \]

\( X_{k_i}(f_i) \) is the Fourier transform of the system input and \( H_{k_1,\ldots,k_i}(f_1, \ldots, f_i) \) with \( 1 \leq k_i \leq d \) (\( d \) is the potential highest order of the system) are the \( i \)th order frequency-domain Volterra kernels. If the system is taken as a 2-order system (\( d=2 \)), the \( H_{k_1,\ldots,k_i}(f_1, \ldots, f_i) \) item will contain \( H_1(f_1), H_2(f_1), H_{11}(f_1, f_2), H_{12}(f_1, f_2), H_{21}(f_1, f_2), H_{22}(f_1, f_2) \). Volterra series method depicts that any nonlinear system, in principle, can be modeled through multidimensional convolution. However, in practice, the kernels are usually only considered up to order 3 or rarely 5 due to complex kernel.

Fiori F. et al employed the Volterra series to represent the distortion of a weakly nonlinear negative feedback opamp driven by conveyed radio frequency interference (RFI) and in particular, the output voltage expression for two arbitrary input signals has been derived using second-order Volterra series (Fiori F. et al 2003). Based on the derived Volterra kernels in (Fiori F. et al 2003) and with input voltage \( V^+ = 0 \) in our case, we can get the opamp output voltage as follows:

\[ v_o(t) = \int_{-\infty}^{\infty} H_2(f_1) V^-(f_1) \exp{j2\pi f_1 t} \, df_1 \]

\[ + \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} H_{22}(f_1, f_2) V^-(f_1) V^-(f_2) \exp{j2\pi(f_1 + f_2)t} \, df_1 \, df_2 \]

(5-11)

where \( H_2(f_1) \) and \( H_{22}(f_1, f_2) \) are the frequency domain Volterra kernels, that depend on the opamp parameters and the external components.

Now we start the derivations for \( H_2(f) \) and \( H_{22}(f_1, f_2) \):

(1) Assumptions for an ideal opamp:

\[ Z_3(f) = 0, \quad Z_{in}(f) = \infty, \quad Z_0(f) = 0, \quad Z_L(f) = \infty. \]
Then the negative feedback opamp in Figure 5-12 can be simplified as in Figure 5-13, where the relationship is as follows:

\[ Z_1(f) = R_1 \]
\[ Z_2(f) = \frac{R_2}{1 + j\omega R_2 C_2} \]
\[ A_d(f) = A_0 \frac{1}{1 + jf/f_1} \]  \hspace{1cm} (5-12)

Here \( A_0 \) is the amplification factor and \( f_1 \) is the cut-off frequency of the low pass filter.

(2) Derivation of \( H_2(f) \):

\[ H_2(f) = \frac{1 - B(f) A_d(f)}{1 + B(f) A_d(f)} \]  \hspace{1cm} (5-13)

\[ \therefore B(f) = - \frac{Z_0(f)}{Z_0(f) + Z_1(f)} \cdot \frac{Z_2(f)}{Z_1(f) + Z_2(f)} \cdot \frac{Z_{\text{in}}(f)}{Z_{\text{in}}(f) + Z_3(f)} \]  \hspace{1cm} (5-14)

but

\[ Z_1'(f) = \frac{Z_3(f)[Z_2(f) + Z_{\text{in}}(f)]}{Z_1(f) + Z_3(f) + Z_{\text{in}}(f)} = Z_1(f) \]  \hspace{1cm} (5-15)
\[ Z_2'(f) = \frac{Z_4(f)[Z_3(f) + Z_{\text{in}}(f)]}{Z_2(f) + Z_3(f) + Z_{\text{in}}(f)} = Z_1(f) + Z_2(f) \]  \hspace{1cm} (5-16)

\[ \therefore B(f) = 0 \]  \hspace{1cm} (5-17)
\[ H_2(f) = A_d(f) \] (5-18)

(3) Derivation of \( H_{22}(f_1, f_2) \):

\[ H_{22}(f_1, f_2) = G_{12}(f_1)G_{22}(f_2)H_0(f_2, f_1) + G_{22}(f_1)G_{12}(f_2)H_0(f_1, f_2) \] (5-19)

where

\[ H_0(f_1, f_2) = \frac{1}{2} \frac{A_d(f_1 + f_2)}{2I_0 \left( 1 + B(f_1 + f_2)A_d(f_1 + f_2) \right)} \times \frac{2g_m j2\pi f_1 C_T}{j2\pi f_1 (2C_{gs} + C_T) + 2g_m} \]

\[ H_0(f_1, f_1) = \frac{1}{4I_0} A_d(f_1 + f_2) \cdot \frac{2g_m j2\pi f_2 C_T}{j2\pi f_2 (2C_{gs} + C_T) + 2g_m} \]

\( I_0 \) is the bias current at the transistors, \( g_m \) is the transfer conductance of the opamp, \( C_{gs} \) is the gate-to-source capacitance of each transistor, \( C_T \) is the sum of the parasitic capacitances related to the ground and the power supply, and

\[ G_{12}(f) = \frac{-D(f)Y_1(f)}{Y_1(f) + Y_2'(f)[1 + A'_d(f)D(f)] + Y_1'(f)} \] (5-22)

\[ G_{22}(f) = \frac{2[D(f)]Y_1'(f)}{2[1 + [D(f)]Y_2'(f)[1 + A'_d(f)D(f)] + Y_1'(f)]} \] (5-23)

\( \therefore \)

\[ D(f) = \frac{Z_{in}(f)}{Z_0(f) + Z_{in}(f)} = 1 \] (5-24)

\[ Y_1(f) = \frac{1}{Z_1(f)} \] (5-25)

\[ Y_2'(f) = \frac{1}{Z_2(f) + Y_0(f) + Y_L(f)} = \frac{1}{Z_2(f)} \] (5-26)

\[ Y_3'(f) = \frac{1}{Z_3(f) + Z_{in}(f)} = 0 \] (5-27)

\[ A'_d(f) = A_d(f) \frac{Z_1(f)}{Z_L(f) + Z_0(f)} = A_d(f) \] (5-28)

\( \therefore \)

\[ G_{12}(f) = -\frac{1}{\frac{1}{Z_1(f)} + \frac{1}{Z_2(f)}[1 + A_d(f)]} = -\frac{Z_2(f)}{Z_2(f) + Z_1(f)[1 + A_d(f)]} \] (5-29)
\[ G_{22}(f) = \frac{\frac{1}{Z_2(f)}}{2 \left( \frac{1}{Z_1(f)} + \frac{1}{Z_2(f)}[1+A_d(f)] \right)} = \frac{Z_2(f)}{2(Z_2(f) + Z_1(f)[1+A_d(f)])} \quad (5-30) \]

### 5.3.3 Modeling validation

The interference output voltage induced by a narrow-band on-body communication signal at carrier frequency \( f_c \) is as (Fiori F. et al 2003):

\[
v_0(t) = V_i H_2(f_c) \cos[2\pi f_c t + \angle H_2(f_c)] + 0.5V_i^2 Re[H_{22}(f_c - f_c)] \quad (5-31)
\]

where \( V_i \) is the amplitude of the input narrow band signal. Since the term at \( f_c \) is beyond the opamp circuit bandwidth, only the second term in Equation (5-31) is effective. This indicates that the EMI effect is actually an offset in the output voltage.

Based on Equations from (5-19) to (5-30), we have the interference output voltage \( V_0 \) as

\[
V_0 = 0.5V_i^2 Re[H_{22}(f_c - f_c)] = V_i^2 A_0 \frac{g_m}{g_i} Re\left\{ \frac{j2\pi f_c C_T}{j2\pi f_c (2C_{gs} + C_T) + 2g_m} \times \left[ \frac{Z_2(f_c)}{R_1 + Z_2(f_c)} \right]^2 \right\} \quad (5-32)
\]

According to (Schenke S. 2007), the low-pass filter of the pacemaker circuit has a cut-off frequency of 1 kHz. Assuming a 10 dB gain of the negative feedback opamp circuit, we determined the component parameters for the circuit in Figure 5-13, i.e., \( R_1 = 1K\Omega, R_2 = 3K\Omega \) and \( C_2 = 53.05nF \). Furthermore, with typical values of \( A_0, I_0, g_m, C_{gs} \) and \( C_T \), we can obtain the output voltage \( V_0 \) for a known interference input voltage at the pacemaker sensing circuit.

According to the measured values given in (Schenke S. 2007), the critical input voltage from 10 to 100 MHz is almost 1 V. This means that when the input interference voltage is 1 V, the output interference voltage will be the threshold voltage \( V_t \). When the input interference voltage exceeds 1 V, the pulse output switches on and this is rated as heart beat.

According to (Imich W. et al 1996), the mean value for the sensing threshold voltage \( V_t \) of a pacemaker is around 2.2 mV. The white circles in Figure 5-14 show the measurement-based output voltage as a function of frequency between 10 MHz and 100 MHz. It can be found that the output voltage is almost flat within this frequency band and has a constant level of about 2.2 mV.
On the other hand, based on the proposed nonlinear opamp model in Equation (5-32), we calculated the output voltage $V_o$ for the same input voltage $V_t$ as given in (Schenke S. 2007). The results are also plotted in Figure 5-14 with black circles. The predicted interference output voltage is found to agree well with the measurement-based one, which assured the validity of the nonlinear circuit model. The circuit parameters used in the model are listed in Table 5-3.

<table>
<thead>
<tr>
<th>Table 5-3 Opamp circuit parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Amplification factor $A_0$</td>
</tr>
<tr>
<td>Bias current $I_0$</td>
</tr>
<tr>
<td>Transfer conductance $g_m$</td>
</tr>
<tr>
<td>Gate-to-source capacitance $C_{gs}$</td>
</tr>
<tr>
<td>Parasitic capacitance $C_r$</td>
</tr>
</tbody>
</table>

Figure 5-14 Comparison of measurement-based and calculated output voltages $V_o$

5.3.4 EMI evaluation

Based on the above consideration and derivation, the prediction of EMI voltage at the analogue sensing circuit output of pacemaker can be now described in details as follows:
1. Calculate the open voltage $V_M$ using the FDTD method by modeling the pacemaker as a receiving antenna. Since the radiation impedance $Z_R$ is much smaller than the input...
impedance $Z_I$ of the opamp circuit, based on Equation (5-9), it is reasonable to use the open voltage $V_M$ as the input voltage $V_I$, which actually considers a worst case.

(2) Calculate the output voltage $V_O$ using Equation (5-11) at the carrier frequency and compare it with the sensing threshold $V_t$. If $V_O > V_t$, the heart-beat-simulating pulse will be triggered and a malfunction may occur.

This approach is now applied to a UWB wearable body sensor communication scenario. In this application scenario, the pacemaker user adorned his chest with an on-body UWB transmitter for some medical diagnose purpose. Since the transmitter was just on the chest surface, induced EMI voltage to the implanted pacemaker should have a significant level.

The UWB pulse is generally generated in the form of a nth-derivative Gaussian pulse. The nth-derivative Gaussian pulse has the following pulse expression:

$$v^{(n)}(t) = -\frac{n-1}{\sigma^2} v^{(n-2)}(t) - \frac{t}{\sigma^2} v^{(n-1)}(t)$$  \hspace{1cm} (5-33)

and the corresponding Fourier transform is

$$X_n(f) = A(j2\pi f)^n e^{-\frac{(2\pi \sigma^2 f)^2}{2}}$$  \hspace{1cm} (5-34)

where $A$ is the amplitude and $\sigma^2 = \alpha^2/4\pi$ is the variance and $\alpha$ is the shape factor. Then we have $V^-(f) = X_n(f)$ in Equation (5-11).

For the EMI evaluation, we choose a second-derivative Gaussian pulse as the transmitted UWB pulse. The second-derivative Gaussian pulse has the following pulse expression:

$$v(t) = -\frac{A}{\sqrt{2\pi \sigma^2}} (1 - \frac{t^2}{\sigma^2}) e^{-\frac{t^2}{2\sigma^2}}$$  \hspace{1cm} (5-35)

The pulse amplitude peak keeps 0.3 V and the pulse shape factor has been adjusted to $\alpha = 1.1802e - 10$ so as to have a spectrum close to the FCC emission requirement as much as possible. The pulse width is 350 ps as shown in Figure 5-15. It has been derived in (Wang J.Q. 2009) that the UWB pulse attenuates about 54 dB when it propagates from the left chest surface to the location where the pacemaker is implanted. Therefore, we can derive the maximum UWB interference input voltage at the implanted pacemaker connector will be 0.6 mV. Figure 5-16 shows the interference voltage spectral density of the second-derivative
Gaussian pulse under FCC UWB emission limit at the pacemaker connector. Based on Equation (5-11), the interference output voltage waveform is shown in Figure 5-17. It can be seen that the amplitude peak value of the interference output voltage is 0.037 mV. Compared with the sensing threshold voltage $V_t = 2.2$ mV, the interference output voltage peak is very small and the safety margin is around 35 dB. The direct current (dc) component of the interference output voltage, which is only 0.06 μV, is much smaller compared with the sensing threshold voltage. Figure 5-18 (a) and (b) show the input-output voltage relationships for output amplitude peak and output dc component respectively. It can be concluded that when the input pulse peak is less than 1 mV, both the peak and dc component of the output pulse keep linear relationships with the input pulse peak. This is reasonable since when the input amplitude is very small, the second item in Equation (5-11) which contains a quadratic of the input signal will become extremely minute. Moreover, the frequency domain $2^{nd}$-order Volterra kernel $H_{22}(f_1, f_2)$ is much smaller than the $1^{st}$-order Volterra kernel $H(f)$ at UWB frequency band. Therefore the nonlinear effect on input-output amplitude is negligible and a linear amplitude relationship can be maintained as shown in Figure 5-18 (a) and (b).

![Figure 5-15 Second-derivative Gaussian pulse waveform under FCC UWB emission limit](image)
Figure 5-16 Input interference voltage spectral density of the second-derivative Gaussian pulse under FCC UWB emission limit at the pacemaker connector.

Figure 5-17 Predicted interference output voltage waveform of the pacemaker circuit for maximum permissible UWB pulse input.

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Figure 5-18 Predicted interference output voltage amplitude (a) peak (b) dc vs. input UWB pulse amplitude peak for the pacemaker circuit

However, the demodulating property of the opamp's nonlinearity changes the output spectral components. Figure 5-19 (a) and (b) show the interference input and output signal spectrums between 0-15 GHz. As can be seen, the interference output spectrum is not proportional to the input spectrum within the whole UWB frequency band. Some spectral components have been amplified while some others have been attenuated. The transmission characteristics at different frequency points within UWB band are shown in Figure 5-20 using the output and input spectrum amplitude ratio. Input high frequency components will be demodulated and then contribute to the output low frequency components. Figure 5-21 shows the output and input spectral amplitude ratio at 1 KHz (the cut-off frequency of the opamp) with the input pulse peak ranging from 10 μV to 1 V. The spectral component ratio at 1 KHz is as high as 233 dB which declares a high amplitude increase of low frequency components. The high gain of the low frequency component is far beyond the opamp amplification ability which is attributed to the demodulating property of the opamp. Moreover, Figure 5-22 derives the ratio of output spectral amplitude at 1 KHz and input spectral amplitude at 7 GHz is -32 dB which declares that the demodulating property keeps constant when the input pulse peak is
less than 1 V. Around the heartbeat frequency 0.1 KHz, nearly the same results and conclusions as at 1 KHz have been drawn. The interference output voltage component at 0.1 KHz is in a order of $1e^{-12}$ V and will not make any influence on heartbeat behavior.

Figure 5-19 Interference input (a) and output (b) signal spectrums under FCC UWB emission limit

Figure 5-20 Interference input and output spectral amplitude ratio between 0 Hz and 10 GHz.
Figure 5-21  Output and input spectral amplitude ratio at 1 KHz vs. input pulse peak

Figure 5-22  Ratio of output spectral amplitude at 1 KHz and input spectral amplitude at 7 GHz vs. input pulse peak

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5.3.5 Conclusion

Based on the EM field analysis and nonlinear electric circuit analysis, a two-step hybrid approach has been proposed to model the induced EMI voltage at the cardiac pacemaker by on-body communication signals. The nonlinear opamp model with Volterra series representation in the electric circuit analysis has been validated by available measurement data form literatures. For UWB wearable body sensor communication scenario, the interference output voltage of a pacemaker under the FCC UWB emission limit is much smaller than the actual sensing threshold of pacemaker. A safety margin of around 35 dB has been concluded. Moreover, evaluation results expound the fact that the demodulating property of the operational amplifier of pacemaker converts the high frequency components in the input UWB body communication signals to low frequency components. This reveals the basic mechanism of a high-frequency UWB signal interfering with the low-frequency pacemaker circuit.
Chapter 6 Conclusions

This study has developed an effective UWB wearable body area communication system for medical and healthcare applications. Three major issues are covered: (1) channel modeling, (2) communication performance, and (3) human safety and EMC. The major concluded results are summarized as following.

1. UWB wearable body area channel modeling
   
   (1) A frequency-dependent FDTD numerical method and a human body model capable of various body postures have been developed in order to develop a dynamic channel model with an emphasis on body movements and postures.

   (2) Based on the data derived from FDTD simulation, a small set of parameters has been provided to implement the discrete time impulse response channel model for five typical transmission links in which the transmitter is fixed at the left chest and the receivers are located at the right chest, the left and right waist, and both ears respectively.

   (3) The derived channel model is a discrete impulse response model which is convenient for characterizing the multi-path effect. By virtue of a high time resolution in the numerical simulation and concluded low correlation between rays, this impulse response model is implemented based on distinguishable multi-path rays without resorting to a uniformly-spaced tapped delay line model.

   (4) The statistically implemented model and the data derived from FDTD simulation in terms of key communication metrics are in good agreement, which demonstrates the validity of the channel model.
(5) Measurement has been conducted and the results provides additional verification of the FDTD-derived data and the modeling results.

2. Communication performance evaluation

(1) The maximum desirable data rates for upper body channel with correlation receiver are 70 Mbits/s and 90 Mbits/s which correspond to the BER threshold of 0.001 and 0.01 respectively.

(2) The 2-finger and 4-finger RAKE receiver considerably improve the BER performance in multi-path-affected on-body channel compared with the normal correlation receiver. At the data rate of 70 Mbits/s, the degradation in performance of the 2-finger and 4-finger RAKE receivers is less than 2 dB at a BER of 0.01 with respect to the ideal Gaussian noise channel, while at data rate of 90 Mbits/s, the performance degradation is less than 4 dB.

(3) The 2-finger RAKE receiver might be a good option for body area communication since it can get a satisfactory performance and at the same time holds a relatively simple structure compared with 4-finger RAKE receiver.

(4) The effective maximum communication distance for UWB-PPM-TH with 2 and 4 fingers RAKE receiver is around 0.6-0.8 m on the human body since it is concluded under the maximum data rates. On the other hand, if the maximum transmitted power keeps not be raised, the data rate has to be decreased to compensate the energy loss due to the longer communication distance. The whole body communication can still keep a data rate in the order of Mbits/s.

3. Human safety and EMC analysis

(1) Two analysis approaches in time-domain and frequency-domain repectively have been proposed to calculate the SA and SAR for UWB pulse exposure. The two approaches have the same accuracy but the time-domain approach is more straightforward to the SA and SAR analysis for UWB-IR systems.

(2) SA and SAR levels under the FCC UWB limit have been demonstrated for various antenna locations on an anatomical human body model. The results have shown that the SA and SAR levels are much smaller than the ICNIRP safety limits and
the multiple exposures do almost not obviously increase the SA and SAR as long as two UWB devices have a separation as large as 30 cm.

(3) The ten-gram-averaged SA is in the order of pJ/kg and is much smaller than the ICNIRP safety guideline of 2 mJ/kg. The ten-gram-averaged SAR is in the order of mW/kg and is smaller than 1/2000 of the ICNIRP safety guideline of 2 W/kg.

(4) The 10g-averaged SA and SAR values will decrease by 12 dB and 17 dB respectively when the distances become 1 cm and 2 cm compared with the initial distance of 2 mm. The further the distance becomes, the larger the safety margin will be.

(5) A two-step hybrid approach has been presented to model the induced EMI voltage at the cardiac pacemaker by on-body communication signals. The demodulating property of nonlinear opamp of pacemaker converts the high frequency components in the input UWB body communication signals to low frequency components. The nonlinear opamp model with Volterra series representation in the electric circuit analysis has been validated by available measurement data form literatures.

(6) For UWB wearable body sensor communication scenario, the interference output voltage of a pacemaker under the FCC UWB emission limit is much smaller than the actual sensing threshold of pacemaker. A safety margin of around 35 dB has been concluded.

The proposed on-body UWB channel model and communication evaluation conclusions can contribute to the standardization of body area networks such as the IEEE 802.15.6, which is now in the progress for medical applications. The communication performance analysis and EMC evaluation have demonstrated the feasibility of a body area communication system for medical applications. In practice, however, the channel model should include the effects of the surrounding environment such as floors, walls and so on. A complete on-body UWB channel model can be regarded as a combination of two parts. At first, the model generates components which diffract around and reflect from parts of the body, as proposed in this study, after which it adds additional components from the surrounding environment, as
modeled in (Molisch et al. 2006). In fact, adding the effects of the surrounding environment is included in the plan of our future work. Besides, other UWB radio signals like direct spreading-UWB (DS-UWB) and multi-band-UWB (MB-UWB) can also be used to perform the communication performance analysis and evaluation. Comparison between multiple radio techniques and modulation schemes will be more helpful to contribute to the development of the body area communication system for medical and healthcare applications.
Channel model realization algorithm in Matlab

Main routine

% On-body channel model generation

clear;

% Body model parameters
% Ray decay factor
gamma=0.5;
% sampling time (nsec)
ts=0.00385;
% total std of log-normal shadowing for each
std_ln=7.7;

% Ray arrival
std_ln_1=std_ln/sqrt(2);
% Standard deviation of log-normal
variable associated with Clusters
std_ln_2=std_ln/sqrt(2);
% Standard deviation of log-normal
variable Rays within a cluster

num_channels=100;
F=20;
C=3;

index_imp_response=floor((F*gamma+C)/ts)+1;
imp_response=zeros(index_imp_response,num_channels);

for k=1:num_channels
    if (mod(k,10)==0)
        k;
    end

    tmp_imp_response=zeros(floor((F*gamma+C)/ts)+1,1);

% Determine Ray arrivals for the only cluster
Tr0=2.0;
Tr=Tr0;

while (Tr<F*gamma+Tr0)
    tmp_imp_response_index=floor(Tr/ts)+1;
    mu=(-10*(Tr-Tr0)/gamma)/log(10)-(std_ln)^2*log(10)/20; % assume
    log(Omega)=0
    ln_rv_temp=std_ln_1*randn;
    ln_rv=mu+std_ln_2*randn+ln_rv_temp; % Ray fading
    pk=2*round(rand)-1;
    tmp_imp_response(tmp_imp_response_index)=10^(ln_rv/20)*pk;
    Tr=Tr+1;
end
% determine excess delay
sq_imp_response=imp_response(:, k).^2;
max_tap=floor((F*gamma+C)/ts)+1;
t=(0:(max_tap-1))'*ts;
excess_delay(k)=t'*sq_imp_response;

% determine RMS delay spread
RMS_delay(k)=sqrt((t.^2)'*sq_imp_response-(excess_delay(k))^2);
SUM_p=sum(sq_imp_response);

mean_excess_delay=mean(excess_delay);
mean_RMS_delay=mean(RMS_delay);

figure(1)
plot(t,imp_response)
grid
title('Impulse response realizations')
xlabel('Delay (nsec)')
ylabel('Amplitude')

figure(2)
plot(excess_delay)
grid
title('Excess delay')
xlabel('Multipath realization number')
ylabel('Excess delay (nsec)')

figure(3)
plot(RMS_delay)
grid
title('RMS delay')
xlabel('Multipath realization number')
ylabel('RMS Delay (nsec)')

temp_average_power=sum(imp_response'*imp_response)/num_channels;
temp_average_power=temp_average_power/max(temp_average_power);
average_decay_profile_DB=10*log10(temp_average_power);
pt_mean=sum(imp_response'*imp_response)/num_channels;
pt=imp_response(:,1)'*(imp_response(:,1));
figure(4)
plot(t,average_decay_profile_dB)
grid
title('Average Power Decay Profile')
xlabel('Delay (nsec)')
ylabel('Average power (dB)')

figure(5)
plot(t,imp_response(:,1))
grid
title('Example impulse response realization')
xlabel('Delay (nsec)')
ylabel('Amplitude');

e_d=excess_delay';
RMS_d=RMS_delay';
save excess_delay_Model.dat e_d -ascii;
save RMS_delay_Model.dat RMS_d -ascii;
Subroutine

function f=IGNum(mu, lambda)
  y=chi2rnd(1,1,1);
  xl=mu*(2*lambda+mu*y.^2-
         power((4*lambda*mu*y.^2+mu.^2*power(y,4)), 0.5))/ (2*lambda);
  x2=mu.^2/xl;
  p1=mu/(mu+xl);
  p2=xl/(mu+xl);
  u=rand(1,1);
  if u>p1
     f=x2;
  else
     f=x1;
  end
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Acknowledgements

I would like to express my profound gratitude to my supervisor Professor Jianqing Wang. My thanks go first for accepting me at his laboratory, giving me all the orientation and support that allowed me to develop and conclude this research. My thanks go also for his supervision, discussion and inspiring through the whole research and his tolerance, patience and kindness for my various discussion and questions. Thanks go for his guidance, which has been making me realize and experience the fact that the scientific research work is really like a kind of adventure. The rewarding pleasure through the whole research adventure is unparalleled and extraordinary whenever the difficulties are cleared up one by one. Great gratitude for his setting himself as an example to teach me the essential dispassion, rigor, perceptivity, concentration, humility, persistence and confidence when man is occupied with scientific research, which has had and will ever have a lasting impact on my future research work. Without him, I could have not enjoyed the pleasure of research work and I cannot even be the one I am now.

My great gratitude to Professor Osamu Fujiwara, who always guided me in the research work and at the daily laboratory life since my first day in Nagoya Institute of Technology. Without his practical advices and constructive inspiration along the doctoral course, this Ph.D. thesis work would have not been completed in such a limited time. Special thanks for his frequent inviting me to his lab’s parties and also the delicious wine offer, thanks for his humorous talking which always made my life in Japan happy times. My thanks go also to his colleagues Associate Professor Hirada, Assistant Professor Yoshinori Taka and Secretary Saiki for all the kind help.
My great gratitude goes also to Professor Yasunori Iwanami, who gave a great deal of constructive suggestions and comments for my research work. Thanks for his strict and careful examination of my thesis and I really learn and benefit a lot from his modifications and comments.

My special gratitude to Harumi Hasegawa and Tomonari in the General Affair Office and Fubu in the posting office, for all their help, support and friendliness, for treating me with respect and trust. With them, my life in Nagoya Institute of Technology is always full of happy memories.

My deep gratitude for my tutor and friend Kun Sun and Masami, for the friendliness since my first day at the laboratory and all the help they provided me which made my life and study in Japan easy and smooth. Many thanks to all past and present laboratory members, in special to Nishikawa, Tayamachi, Takahashi, Ipei, Shaolang Chen, Suzuki, Sanpei, Okawa, Urami, Takeoka, for all the discussions, support, translations, technical help and companionship.

My gratitude to Keigo Kuma, my best friend in Japan, for all his help, friendship and trust. My thanks also go to Mauricio Kugler, for his kindness on all times I asked him for help on Latex and Microsoft Word. Thanks to all my friends in Nagoya Institute of Technology, in special to Shaolang Chen, Cong Li, Victor Benso, Cheng Wan, Puwu Chuan and so on.

I am inexpressibly grateful to all my family, specially my father Jianping and my mother Guangzhen, for their incentive, patience, support, love and respect and for believing on my dreams. Special thanks also go to my dear grandmother, for her love and for enduring my absence.

Finally, I express gratitude to all people that contributed to the realization of this work but unfortunately I forgot to mention their names.
Publications

Journals


International Conferences


**Domestic Conferences**


Awards

2009.6.9  Student Research Award of the Institute of Electronics, Information and Communication Engineers, Tokai Branch

2009.7.22 Student Award of the 2009 International Symposium on Electromagnetic Compatibility, Kyoto (EMC'09/Kyoto)