Performance analysis and integrated circuit design for ultra-wideband transceiver

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Performance Analysis and Integrated Circuit Design for Ultra-Wideband Transceiver

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
Doctor of Philosophy

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Summary

The ultra-wideband (UWB) technology has unique features over a wide range of applications. Comparing with traditional narrowband wireless technologies, UWB technology employs ultra-wide bandwidth and permits high data-rate and very low power transmissions. Since 2003, UWB technology has been undergoing rapid development in both academia and industry. An UWB front-end has to accommodate design challenges over exceptional wide bandwidth, and is of essential importance for overall transceiver system.

The thesis studies the implementation-oriented aspects of UWB transceiver front-end. It includes two aspects of work: analysis of system performance with realistic front-end; and design of radio-frequency (RF) integrated circuits (IC) for UWB transceiver.

A general framework for performance evaluation of realistic UWB impulse radio system is proposed, which is based on the study of waveform distortion due to performance fluctuation of nonideal front-ends. The nonidealities are modeled by the transfer functions, whose fluctuation or ripple in the frequency domain results in the waveform distortion in the signal path. A pulse-position modulation (PPM) and correlation based UWB transceiver with three widely proposed UWB antennas is studied in four deterministic multipath channels

which represent various UWB applications. The inter-symbol interference (ISI) effects on waveform distortion and system performance degradation are also demonstrated.

The low power property of UWB impulse radio is particularly suitable for low data-rate wireless body-area networks (BAN), where the human body has significant impact on the wideband signal propagation. The UWB on-human-body channel is studied and its effects on the system performance are evaluated. Various pulse shapes and modulation schemes are discussed and their efficiencies under different on-human-body scenarios are simulated.

An inductorless low-noise amplifier (LNA) capable of UWB applications is designed and fabricated. Without on-chip inductors, the ultra-wide -3 -dB bandwidth is achieved by a syncretic adoption of thermal noise canceling, capacitor peaking, and current reuse. Fabricated in $0.13\text{-}\mu\text{m}$ CMOS process, the LNA achieves a small signal gain of 11-dB and a -3-dB bandwidth of $2\text{--}9.6\text{-GHz}$ with $S_{11} < -8.3\text{-dB}$ in-band impedance matching. The noise figure is $3.6\text{--}4.8\text{-dB}$ over $2\text{--}9.6\text{-GHz}$. The LNA consumes 19-mW from a low supply voltage of 1.5-V . The performance achieved in this design is comparable and sometimes better than inductor-based designs. Benefiting from its inductorless architecture, the LNA with test pads occupies only 0.17 mm^2 die area, which is among the smallest UWB LNA designs.

Transmit/receive (T/R) switch is among the several RF circuits for which the CMOS integration is quite challenging. The thesis studied CMOS T/R switch design towards high power handling capability and high frequency applications. A 6-GHz differential T/R switch for higher power handling capability is designed preliminarily to demonstrate the feasibility of differential architecture

SUMMARY

v

for RF switches. The study is subsequently moved towards full-band UWB operations and higher frequencies up to 20-GHz. Comprehensive analysis and performance tradeoffs are presented, and techniques for minimizing parasitics and increasing linearity are discussed. Customized switch transistor layout and triple-well body-floating techniques are proposed and verified by experimental results. Fabricated in 0.13- μm triple-well CMOS, the T/R switch exhibits less than 2-dB insertion loss and higher than 21-dB isolation up to 20-GHz. With resistive body floating and differential architecture, the high linearity is of ultra-wideband characteristic, more than 30-dBm power 1-dB compression point ($P_{1\text{dB}}$) is obtained up to 20-GHz in only 0.03 mm² active area.

Table of Contents

Acknowledgements	i
Summary	iii
Table of Contents	vii
List of Figures	xi
List of Tables	xix
1 Introduction	1
1.1 Motivation	1
1.2 Objectives	4
1.3 Major Contributions of the Thesis	5
1.4 Organization of the Thesis	7
2 UWB Transceiver Design: A Brief Review	10
2.1 Origins and Regulations	10
2.2 Systems and Considerations	13
2.2.1 Impulse Radio	13
2.2.2 UWB Standard	18
2.2.3 Wideband Performance Variations	19
2.3 Integrated Circuits for UWB Transceiver	21
3 Performance Analysis of Realistic UWB System	25
3.1 Introduction	25

3.2	Methodology of Analysis	27
3.3	Antennas and Deterministic Channels	31
3.3.1	Antennas	31
3.3.2	Deterministic Multipath Channels	35
3.4	Numerical Results and Discussions	39
3.4.1	Received Waveform Distortion	40
3.4.2	ISI Distorted Waveform	42
3.4.3	BER Performance	43
4	UWB for Body-Area Networks	47
4.1	Introduction	47
4.2	Impulse Radio WBAN	49
4.3	UWB On-Human-Body Channel	51
4.3.1	Description of Measurements	52
4.3.2	Statistical Parameters of the Channel	57
4.4	UWB WBAN Performance	62
5	Inductorless UWB Low-Noise Amplifier	69
5.1	Introduction	69
5.2	Review of Wideband Noise Canceling	72
5.3	Gain-Enhanced Wideband Noise Canceling Architecture	74
5.3.1	Architecture and Analysis	74
5.3.2	Properties and Limitations	78
5.3.3	Circuit Design	82
5.4	Capacitive Peaking	84
5.5	Measurement Results and Discussions	89
6	Differential T/R Switch	95
6.1	Introduction	95
6.2	Architecture and Considerations	97
6.2.1	Architecture	97
6.2.2	Design Considerations	98

TABLE OF CONTENTS	ix
6.2.3 Layout Considerations	102
6.3 Measurement Results and Discussions	104
6.3.1 Differential-Mode Performance	105
6.3.2 Common-Mode Performance	106
6.3.3 Common-Mode Rejection	107
6.3.4 Power Handling Capability	108
7 High-Frequency T/R Switch	112
7.1 Introduction	112
7.2 Switch Transistor in Cutoff Region	113
7.3 Double-Well Body Floating	120
7.4 Measurement Results and Discussions	125
7.4.1 Differential-Mode Performance	127
7.4.2 Common-Mode Performance	127
7.4.3 Common-Mode Rejection	129
7.4.4 Power Handling Capability	131
8 Conclusions and Recommendations	133
8.1 Conclusions	133
8.2 Recommendations for further research	136
Author's Publications	139
Bibliography	143

List of Figures

- 2.1 Regulations of indoor UWB emission power limit in US, Singapore and Europe. The regulations for Singapore and Europe have not been finalized. 12
- 2.2 A brief description of transmitted binary-PPM signal with time-hopping. 14
- 2.3 Architecture of a correlation-based impulse radio receiver for PPM demodulation. 15
- 2.4 Block diagram of an UWB impulse radio system with PPM and correlation-receiver. 16
- 2.5 Transmitted signals in two alternative impulse radio modulations: (a) anti-phase PSK and (b) OOK. 17
  - (a) anti-phase PSK 17
  - (b) OOK 17
- 3.1 Waveform of Gaussian second-order derivative in the time domain. The amplitude of the waveform is normalized to 1 V. 29
- 3.2 Flowchart for analyzing waveform distortion and system performance considering nonidealities in the individual blocks. 30
- 3.3 Structures and dimensions of 3 planar antennas for UWB transmission. The unit of scales is meter. 32
  - (a) Slot antenna 32
  - (b) Diamond dipole antenna 32
  - (c) Planar volcano-smoke slot antenna (PVSA) 32

3.4 Voltage transfer functions of the three antenna pairs. Each pair of transmitting and receiving antennas are identical and placed in parallel with 1 meter distance, only the LOS path is considered. The input variable is defined as the transmitting source voltage with 50 Ω series impedance, the output variable is the voltage across the 50 Ω load. 34

(a) Amplitude 34

(b) Phase 34

3.5 Received waveform by realistic antennas. The Gaussian second-order derivative pulse of Figure 3.1 is distorted by the antenna pairs with transfer functions of Figure 3.4. There is also a 3.3 ns time delay due to the 1 m LOS path. 35

3.6 Configuration of deterministic multipath channels. The unit of scales is meter. 38

(a) Urban environment. 38

(b) Inter-floor environment. 38

3.7 Received waveforms by diamond dipole antennas in the two-ray channel. Comparing with the original transmitted waveform of Figure 3.1, the received waveform encounters a significant distortion because of the nonideal antennas and multipath channels. The 0.32 m longer reflected path leads to a 1.08 ns delay between the two rays. The pulse duration is too short comparing to the inter-path time delay. 40

3.8 Received waveforms by diamond dipole antennas in the urban and inter-floor deterministic multipath environments. Note that due to the visibility of the figures, only the strongest waveforms are displayed among all multipath components. 41

(a) Urban environment of Figure 3.6(a). 41

(b) Inter-floor environment of Figure 3.6(b). 41

3.9 Waveform distortion considering ISI with diamond dipole antenna in the two-ray channel. The ISI occurs by assuming both symbol time interval and time delay due to reflection are about 1 ns. Comparing with Figure 3.7, the ISI further distorts the received waveform in both of the two rays. 43

3.10 BER of overall UWB impulse radio system considering realistic antennas in the LOS channel. 44

3.11 BER of overall UWB impulse radio system considering realistic antennas in the two-ray channel. 44

3.12 BER of overall UWB impulse radio system considering realistic antennas in the urban environment of Figure 3.6(a). 45

3.13 BER of overall UWB impulse radio system considering realistic antennas in the inter-floor environment of Figure 3.6(b). 45

4.1 The antenna used in the measurement of on-human-body channel. It is made in silver on a 0.8mm-thick substrate with $\epsilon_r = 5.9$. 53

4.2 The antenna and its measured return loss. The antenna return loss is improved to below -10 dB for the on-body cases over the entire UWB band due to the lossy nature of the body. 53

4.3 The measured antenna radiation patterns at 6.85-GHz. It has an omni directional radiation pattern in the E-plane and an ∞ shaped radiation pattern in the H-plane. 54

    (a) E-plane (the YZ plane) 54

    (b) H-plane (the XZ plane) 54

4.4 The measurement environment of (a) anechoic chamber and (b) staff lounge room. 55

    (a) anechoic chamber 55

    (b) staff lounge room 55

4.5 Location of test points around human body. All test points form a cylinder with 1 m perimeter. 56

4.6 Transfer functions at test point A in Figure 4.5 in anechoic chamber and staff lounge room. The phase is unwrapped. The loss is relatively small and amplitude fluctuations is high in the staff lounge room due to environment reflections. 58

    (a) Magnitude 58

    (b) Phase 58

4.7 The radiograph of channel path loss on and in close proximity to the body. The unit of path loss is dB. The result is obtained by the measurements in staff lounge room. The position of Tx antenna is (0, 1.2m). 59

4.8 The radiograph of channel rms delay spread on and in close proximity to the body. The unit of rms delay spread is ns. The result is obtained by the measurements in staff lounge room. The position of Tx antenna is (0, 1.2m). 61

4.9 BER performance of UWB impulse radio when the receiver is placed at point A in (a) anechoic chamber and (b) staff lounge room. 64

    (a) anechoic chamber 64

    (b) staff lounge room 64

4.10 BER performance of UWB impulse radio when the receiver is placed at point B in (a) anechoic chamber and (b) staff lounge room. 65

    (a) anechoic chamber 65

    (b) staff lounge room 65

4.11 BER performance of UWB impulse radio when the receiver is placed at point C in (a) anechoic chamber and (b) staff lounge room. 66

    (a) anechoic chamber 66

    (b) staff lounge room 66

LIST OF FIGURES

xv

5.1	Simplified wideband LNA architectures exploiting thermal noise canceling with (a) resistive shunt-feedback and (b) common-gate input.	73
	(a) shunt-feedback input	73
	(b) common-gate input	73
5.2	Architecture of gain-enhanced wideband LNA with thermal noise canceling.	75
5.3	Numerical simulation of EF_{MD} with regard to the variations of $A_{v,2}$ and $A_{v,3}$	82
5.4	Simulated performance of the gain-enhanced noise canceling circuit in Figure 5.2.	84
5.5	Example of a gain-enhanced noise canceling LNA with inductor peaking.	85
5.6	Simulation result of Figure 5.5 with $L_1=0.6nH$, $L_2=1.5nH$	86
5.7	Schematic of a capacitive peaking gain stage.	87
5.8	Complete schematic of gain-enhanced noise canceling LNA with capacitive peaking.	88
5.9	Simulation result of the proposed noise canceling LNA with gain-enhancement and capacitive peaking.	89
5.10	Die microphotograph of the fabricated LNA chip.	90
5.11	Measured performance of the fabricated LNA.	91
5.12	The input 1-dB compression point (ICP) and input third-order intercept point (IIP3) at different frequencies.	92
6.1	Architecture of the proposed differential T/R switch.	98
6.2	Small signal equivalent circuit of a MOS transistor switch.	99
6.3	Simulated insertion loss and isolation performance.	102
6.4	Layout of the proposed differential T/R switch.	103
6.5	Die photograph of the fabricated differential T/R switch.	104
6.6	Measured insertion loss and isolation parameters for differential-mode small-signals.	106

6.7 Measured insertion loss and isolation parameters for common-mode small-signals. 107

6.8 Measured common-mode rejection ratio in the *ON* and *OFF* state of the differential T/R switch. 108

6.9 Linearity of the differential T/R switch at 5.2-GHz. The control voltage is 1.8/0-V and the TX/RX nodes are biased at 0.8-V. . . . 109

6.10 Linearity of the differential T/R switch at different frequencies. The TX/RX nodes are biased at 0.8-V with 1.8/0-V control voltage, and 1.6-V with 3.6/0-V control voltage. 109

7.1 Schematic of series/shunt type single-ended T/R switch. 113

7.2 Parasitic capacitances of a MOS transistor that is used as a switch and turned off. 114

7.3 Layout sketch of drain-source interconnection for an interdigitized MOS transistor that is used as a switch. 115

7.4 Foundry provided p-cell (a) and customized (b) layout of an NMOS transistor operating as a switch. $W = 108\mu m$, $L = 0.13\mu m$, fingers = 6. The drain-source distance in the customized layout is 4 times of that in the standard p-cell. 117

    (a) 117

    (b) 117

7.5 Simulated insertion loss and isolation of T/R switch with and without shunt arm. The switch without shunt arm employs customized layout of Figure 7.4(b). 119

7.6 Simplified cross-sectional view of an NMOS transistor in triple-well technology. 121

7.7 Simulated linearity performance for T/R switch with and without shunt arms, and with different types of body-floating. 123

7.8 Simulated insertion loss and isolation with and without body-floating. Both switches use the proposed customized layout and have no shunt transistors. Double-well body-floating is used. . . . 124

LIST OF FIGURES

7.9 Final schematic and circuit parameters of the differential T/R switch. 125

7.10 Die microphotograph of the fabricated differential T/R switch. . . 126

7.11 Measured insertion loss and isolation parameters for differential-mode small-signals. 128

7.12 Measured insertion loss and isolation parameters for common-mode small-signals. 128

7.13 Measured common-mode rejection ratio in the *ON* and *OFF* state of the differential T/R switch. 130

7.14 Power handling capability in terms of P_{1dB} 130

List of Tables

4.1   Extracted PL_0 and n from Radiograph   60

5.1   Summary of the LNA Performances and Comparison with Re-
ported State-of-the-Art Designs   94

6.1   Summary of Differential Switch Performance   111

7.1   Parasitic Capacitances of NMOS Transistor in Cutoff Region with
 $W=108\mu m, L=0.13\mu m$, fingers = 6   116

7.2   Summary of the T/R Switch Performances and Comparison with
Reported State-of-the-Art Designs   132

Chapter 1

Introduction

1.1 Motivation

The wireless technology has been under extensive research and development during the past decade. There are remarkable advances in both theory and implementation aspects that have been made to enable various wireless applications such as radar, personal communication system (PCS), wireless local-area network (WLAN), radio-frequency identification (RFID), etc.. The capacity, distance, multi-access and power consumption are of the highest importance for a wireless communication system. Traditionally, wireless communication system is narrowband, which employs only a small portion of bandwidth around a certain frequency to transmit information. The design and operation of such narrowband system depend severely on the carrier frequency. Considering the high power spectrum density (PSD) and the resulting interferences, narrowband systems can hardly share the spectrum. As frequency is a rare resource and the transmit power can not be arbitrarily high, narrowband system encounters

severe restraints in achieving high performance with low power consumption concurrently. To date, the highest speed of a commercial narrowband WLAN transceiver is 54-Mbit/s.¹

On the other hand, the demand of higher speed and lower power consumption has been increasing since the emergence of commercial wireless applications. With the development of internet and multimedia technologies, the demand of higher data rate is further boosted. There are two different approaches that have been proposed to further pushing up the data rate of wireless system. One direct approach extends the narrowband concepts to 60-GHz range, where over 5-GHz spectrum can be used for transmission so that a significant speed improvement can be made available² [1, 2, 3, 4]. This approach succeeds to traditional narrowband architectures while placing most of difficulties in hardware implementation. As a result, the integration of 60-GHz transceiver requires sophisticated design techniques as well as advanced semiconductor and package technologies. Significant advances have been made, however, the single-chip solution for 60-GHz transceiver has not yet been demonstrated. In addition, the regulation and standardization of 60-GHz wireless transmission are still not clear.

Ultra-wideband (UWB) technology is another approach to further improving the data rate of wireless systems. Comparing with narrowband system, UWB system utilizes ultra-wide bandwidth for data transmission. As defined by the Federal Communications Commission (FCC) in US, the term *Ultra-Wideband* refers to a minimum -10 -dB bandwidth of 500-MHz or minimum fractional

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¹The maximum 54-Mbit/s refers to the IEEE 802.11a/g WLAN standard. Some of the ongoing and future narrowband standards may have higher data rate, but they have not been realized to the date of the thesis.

²According to FCC's definition, 60-GHz radio is also a kind of UWB.

bandwidth of 20%, whichever is less [5]. UWB technology has unique features because of its wideband nature in a number of future applications including wireless communications, imaging systems, radar, positioning, wireless sensor network, measurement systems, biomedical applications, etc. [6, 7, 8].

- The large bandwidth occupied permits higher achievable data rate than narrowband systems in theory, especially for short-range multiple-access wireless applications.
- The ultra-wide bandwidth results in fine time resolution, which allows accurate ranging and localizing.
- The transmission power can be lowered, so that interferences with existing systems are reduced. Spectrum segments can be reused to a certain extent.
- The wideband nature of signals implies fading robustness.
- Difficult to intercept.
- Easy trade-off among capacity, power, distance, etc..

The validation of unlicensed frequency band for UWB device has permitted and stimulated the commercial development of UWB technology. The research on UWB system and transceiver implementation have been the subject of extensive research in recent years. The issues should be studied regarding the implementation of UWB transceiver include system architecture, performance analysis, channel characterization, antennas, circuit design, baseband signal processing, synchronization, etc..

A radio frequency (RF) front-end is of high importance for transceiver implementation because it directly performs the transmit and receive functions. This is particularly true for UWB applications where the transmitted signal is of ultra-wide spectrum and very short duration. The silicon integration of UWB transceiver is also highly demanded by the portable and low power applications. The unique properties and challenging requirements have urged the integrated UWB transceiver front-end design an imperative for the implementation of UWB wireless systems.

1.2 Objectives

The research work presented in this dissertation is implementation oriented. Issues regarding the integrated UWB transceiver design are studied. This includes two aspects of work: system performance evaluation with nonideal front-end components and RF integrated circuits (IC) design for UWB transceiver.

As a newly emerged technology, UWB transceiver has unique architecture and requirements comparing with those of narrowband transceivers. There are a lot of issues in the system level that should be studied prior to circuit implementations. In particular, the realistic system performance should be evaluated considering nonideal front-end components. Although the theoretical expressions of UWB system performance can be found in literatures [9, 10], it is not efficient enough when it applies to realistic systems. This is an unique issue that results from the ultra-wide bandwidth of operation, where performance fluctuation or ripple regarding frequency can not be avoided for realistic components. Narrowband system does not encounter this problem as the frac-

tional bandwidth is relatively very small and performance fluctuation effects are not significant. The performance evaluation includes also the comparisons among different antenna types, multipath channel environments, and modulation schemes.

The design of RF integrated circuits is essential for integrated UWB transceiver implementation. This part of work includes transmit/receive (T/R) switch design and low-noise amplifier (LNA) design, both in state-of-the-art CMOS technology. As an analog circuit whose transistors do not operate in the saturation region, T/R switch in CMOS is only developed to a limit extent. Further integration of T/R switch in CMOS is an emerging topic. High power handling capability and high frequency design of CMOS T/R switch are going to be studied. The performance of LNA determines directly the overall receiver performance including gain, noise figure and linearity. An inductorless low noise amplifier design capable of UWB operation is going to be explored. In both designs, minimizing power and area consumption is basic. For high frequency analog circuits, silicon verification is a must and measurement results should be presented and discussed.

1.3 Major Contributions of the Thesis

The research on UWB wireless system implementation has been undergoing rapid development since 2003. The work presented in the thesis was conducted in parallel with the trends. The contributions achieved are towards four aspects within the scope of UWB wireless transceiver analysis and design.

The implementation-oriented performance analysis and evaluation of UWB

impulse radio considering realistic antennas and deterministic channels are studied. Based on the point of view of waveform distortion, the nonideal effects due to the realistic components on the overall system are evaluated in terms of bit-error rate. Three antennas and four typical multipath channel environments are considered. This part of work is focused on the pulse-position modulation based impulse radio for high data-rate personal area networks (PAN).

Another important application of UWB is for low data-rate body-area networks (BAN), where the UWB on-human-body channel effects are evaluated. Various modulation schemes are discussed and their efficiencies under different on-human-body scenarios are simulated. The evaluation also includes three pulse shapes.

The RF integrated circuit design for UWB transceiver front-end is another emphasis of the thesis, where an inductorless low-noise amplifier capable of 2–9.6-GHz UWB operation is designed. Based on the study of basic noise canceling architecture, a gain-enhancement technique is proposed and its properties and limitations are discussed. Without on-chip inductors, the bandwidth is achieved by a capacitive peaking technique. The LNA is also designed at low voltage and low power consumption. The LNA is silicon proven.

T/R switch is among the several RF circuits whose CMOS integration is quite challenging. The thesis also contributes to the further development of CMOS T/R switch technology. A 6-GHz differential T/R switch for higher power handling capability is designed and verified in CMOS. The study is subsequently improved towards full-band UWB and higher frequencies up to 20-GHz. Comprehensive analysis and tradeoffs are investigated. A custom transistor layout and triple-well body-floating techniques are discussed. Significant frequency

push-up has been observed from silicon measurements.

1.4 Organization of the Thesis

This chapter presents an overview of the thesis. The following chapters are organized as follows.

Chapter 2 reviews the origins of UWB technology and the not-so-long development of UWB transceiver system and circuits. The standard and proposals for UWB are discussed, with emphasis on the pulse based UWB impulse radio technology. From circuit aspects of view, this chapter introduces the reported state-of-the-art integrated UWB transceiver designs to date.

Chapter 3 studies the waveform distortion and performance of a UWB impulse radio system considering both realistic antennas and deterministic multipath channels. The system observed includes a pulse position modulation and correlation based transceiver, three types of realistic antennas, i.e. slot, diamond dipole and planar volcano-smoke slot antenna, and four cases of deterministic multipath channels which represent four typical propagation environments in UWB applications. The effects of nonidealities in the real systems are evaluated by analyzing the waveform distortion along the signal path. Performance of the overall system in terms of bit-error rate is obtained by observing the eventually received waveforms before the correlator. The simulated waveform distortion and performance degradation resulting from realistic antennas and multipath channels are discussed. This work demonstrates the implementation oriented analysis of pulse-based UWB systems. The selection of antenna over different propagation environment is also demonstrated.

Chapter 4 explores the impacts of 3.1–10.6-GHz on-human-body UWB channel on the impulse radio for wireless body-area networks. Succeeding to the performance evaluation method presented in Chapter 3, a further study is conducted considering more pulse shapes and modulation schemes. The measurement and characterization of the 3.1–10.6-GHz on-human-body UWB channel are devised to generate the radiographs of path loss and delay spread for the first time. The performance of the UWB impulse radio transceiver in terms of bit error rate is evaluated based on the waveform distortion analysis and on-human-body channel measurement. The result shows that human body effect is more significant than the environment multipath effect, especially in the case that the propagation contains no line-of-sight path.

Chapter 5 presents an inductorless low-noise amplifier design for UWB receiver front-end. A current-reuse gain-enhanced noise canceling architecture is proposed, and the properties and limitations of the gain-enhancement stage are discussed. Capacitive peaking is employed to improve the gain flatness and -3 -dB bandwidth, at the cost of absolute gain value. The LNA circuit is fabricated in a $0.13\text{-}\mu\text{m}$ triple-well CMOS technology. Measurement result shows that a small signal gain of 11-dB and a -3 -dB bandwidth of 2–9.6-GHz are obtained. Over the -3 -dB bandwidth, the input return loss is less than -8.3 -dB, and the noise figure is 3.6–4.8-dB. The LNA consumes 19-mW from a low supply voltage of 1.5-V. It is shown that the LNA designed without on-chip inductor achieves comparable performance with inductor-based designs. Benefiting from its inductorless architecture, the LNA circuit with test pads occupies only 0.17 mm^2 die area, which is among the smallest UWB LNA designs.

Chapter 6 discusses a differential T/R switch integrated in a $0.18\text{-}\mu\text{m}$ standard CMOS technology for up to 6-GHz operations. This switch design employs

fully differential architecture to accommodate the design challenge of differential transceiver and improve the linearity performance. It exhibits less than 2-dB insertion loss, higher than 15-dB isolation, in $60\mu\text{m} \times 40\mu\text{m}$ area. 15-dBm power 1-dB compression point is achieved without using additional techniques to enhance the linearity. This switch is suitable for up to 6-GHz low-band UWB transceiver as well as other wireless applications with a moderate power level.

Chapter 7 presents the comprehensive considerations of CMOS T/R switch design towards ultra-wideband and higher frequencies. Techniques for minimizing parasitics and increasing linearity are discussed. A customized transistor layout is proposed for T/R switch design and its effects on insertion loss and isolation are studied. The analysis shows that a series only architecture using the customized transistor layout achieves better insertion loss and reasonable isolation. A double-well body-floating technique is proposed and the tradeoffs are discussed. A differential switch architecture without shunt arms is designed and verified by experimental results. Fabricated in 0.13- μm triple-well CMOS, the T/R switch exhibits less than 2-dB insertion loss and higher than 21-dB isolation up to 20-GHz. With resistive body floating and differential architecture, the high linearity is of ultra-wideband characteristic, more than 30-dBm power 1-dB compression point is obtained up to 20-GHz in only 0.03 mm² active area.

Chapter 8 summarizes the studies and contributions of the thesis. The recommendations for further research are also discussed.

Chapter 2

UWB Transceiver Design: A Brief Review

There are various topics regarding the UWB technology from communication theory to hardware implementation and applications. This chapter reviews briefly the development of UWB wireless system implementations, with particular focus on the architectures and RF integrated circuits for UWB transceiver.

2.1 Origins and Regulations

The history of employing signals with ultra-wide fractional bandwidth can be traced back to the earliest wireless transmissions and radios [11]. However, the radio technology later on adopted carrier based approaches because of the easy realization of narrowband transceivers. In particular, the foremost target of earlier radio was to transmit informations over longest possible distance rather than the capabilities of high data rate and multiple access. The carrier based

narrowband radio then became the prior architecture due to the propagation advantages of carrier signal.

The modern ultra-wideband technology, impulse radio, originated from the work in time-domain electromagnetics began in 1962 [12]. The concept is later used mainly in radar applications. In 1970s, efforts turned towards communications using UWB impulse signals. The development of UWB technology for wireless communications had not been a topic of extensive interests until late 1990s, when the demand of short range, high data rate and multiple access began to increase significantly. On the other hand, most of the usable frequency spectrum has been occupied by the dedicated users and devices. It is necessary to develop a more frequency-efficient communication scheme, or a scheme that can reuse the spectrum with existing users.

The basic idea of the UWB impulse radio scheme is using precisely-timed pulses with extremely short duration for transmissions. In the frequency domain, the spectrum is spread over an ultra-wide bandwidth with very low power spectrum density, which avoids the interferences to existing frequency users. The multiple access is realized by pseudorandom time-hopping of the pulses. Therefore, the users can share the same bandwidth that is overlapped with each other. Ideally, the spectrum of UWB signal can be kept as low as the noise floor with reasonable performance [6, 13, 14].

The unlicensed spectrum and emission power level for indoor UWB devices can be found in Figure 2.1. The bandwidth is significantly wider while the emission power is much lower than the narrowband standards [5]. The interests of UWB development are mainly focused on the bandwidth of 3.1–10.6-GHz, where the emission power can be slightly higher at maximum -41.3-dBm/MHz.

Figure 2.1: Regulations of indoor UWB emission power limit in US, Singapore and Europe. The regulations for Singapore and Europe have not been finalized.

Although the regulations in other countries are not officially issued, the proposals are in the similar manner, around 7.5-GHz unlicensed bandwidth can be used for UWB wireless devices. A lot of efforts have been put on the short-range high-speed UWB developments within 3.1–10.6-GHz since the official release of FCC’s regulation.

The spectrum below 960-MHz is proposed for UWB applications for ranging, positioning, wireless sensor network, etc., where the data rate need not to be high but the hardware implementation can be of very low power and cost.

Note that although the emission power of UWB devices has been restricted to a very low level, the interferences with narrowband devices can not be ignored and there are various effects that should be considered regarding the interferences between UWB and narrowband devices [15, 16, 17, 18, 19].

2.2 Systems and Considerations

There are several approaches that have been proposed to UWB wireless system implementations. Due to the different ways of using the available 7.5-GHz bandwidth, the transceiver architectures and implementation challenges can be quite different from each other. A real UWB system also encounters unique problems due to wideband performance variations of the RF front-end.

2.2.1 Impulse Radio

The deployment of UWB technology originates from pulse based approach, impulse radio (IR) [9, 10, 12, 20, 21]. Impulse radio basically employs precisely-timed pulses with extremely short duration for transmissions, which occupies a wide range of spectrum in the frequency domain. The original impulse radio proposal employs pulse-position modulation (PPM) scheme and time-domain hopping for multiple access. Figure 2.2 shows the transmitted UWB signal with binary-PPM and 3-bit time hopping. Generally, the transmitted signal can be written as

$$s_{tr}^{(k)}(t) = \sum_i \sum_{j=0}^{N_s-1} w(t - iT_b - jT_f - c_j^{(k)}T_c - \delta d_{[j/N_s]}^{(k)}), \quad (2.1)$$

where $w(t)$ represents the monocycle waveform transmitted by the k^{th} transmitter in a multi-user environment. i is the bit index and j is the pulse index. Normally each bit contains several monocycle pulses. T_f is the frame time or pulse repetition time which is typically hundreds to thousands times of the pulse width. The time hopped pulse $w(t)$ occupies within a certain chip time T_c that is decided by the user-dependent pseudorandom code $c_j^{(k)}$. In PPM, a small time

Figure 2.2: A brief description of transmitted binary-PPM signal with time-hopping.

delay δ is employed to modulate the transmitted data sequence $d^{(k)}$. N_s is the number of frames which represent a single bit, the bit time is $T_b = N_s T_f$. In Figure 2.2, $T_f = 3T_c$, $T_b = 3T_f$, and the time-hopping code is $[1\ 0\ 2]$. Note that these values may be quite different in the real implementations, especially for pseudorandom time-hopping code $c_j^{(k)}$ that should be designed more lengthy and hard to be deciphered.

The typical demodulation in a PPM receiver is based on correlation. It is in principle the same as a matched-filter based receiver. Figure 2.3 shows the architecture of a correlation-based receiver. The correlator contains a multi-

Figure 2.3: Architecture of a correlation-based impulse radio receiver for PPM demodulation.

plier (mixer) and an integrator. The received signal, $S(t)$, is multiplied by a synchronized template signal, $V_{templ}(t)$, the product is then integrated within the bit duration T_b . The demodulated bit is decided by the comparison between integrator result and a decision threshold. The template signal can be written as

$$V_{templ}^{(k)}(t) = \sum_i \sum_{j=0}^{N_s-1} v(t - iT_b - jT_f - c_j^{(k)}T_c), \quad (2.2)$$

where $v(t)$ is the monocycle of template signal and is given by

$$v(t) = w(t) - w(t - \delta). \quad (2.3)$$

Note that integration and decision can also be applied to each pulse, i.e. within each frame time T_f , the bit decision is made based on each pulse decision. This gives more flexibilities and potentially higher bit-error rate performance [22]. However, it requires high-speed correlator and consumes more power.

Figure 2.4 shows the block diagram of a complete UWB impulse radio system. The monocycle waveforms are generated by the pulse generator and tem-

Figure 2.4: Block diagram of an UWB impulse radio system with PPM and correlation-receiver.

plate generator, which should be controlled by the same time-hopping sequence and synchronized clock signal. For binary PPM, the analog-to-digital converter (ADC) is simply a two-bit comparator that can be of very high speed.

From the communication point of view, the impulse radio offers great advantages over traditional systems. It is a baseband and carrierless approach, thus the transceiver architecture can be simplified significantly. For instance, intermediate-frequency (IF) circuits and image problems do not exist anymore due to the absence of frequency conversions. Moreover, the transceiver architecture can be reconfigured to provide more flexibilities in system applications. If the ADC is able to sample the received RF signal directly, all the demodulation and correlation functions can be performed in the digital main, which permits good flexibilities in both design and applications [23]. On the other hand, this poses much challenges to the ADC design, sampling a RF signal of 3.1–10.6-GHz with reasonable resolution and power is very difficult for state-of-the-art silicon ADCs. For UWB systems that target on low data rate applications, 960-MHz and below spectrum can be used, where the digital architecture is relatively viable [24, 25].

From the implementation point of view, PPM requires very precise timing

Figure 2.5: Transmitted signals in two alternative impulse radio modulations: (a) anti-phase PSK and (b) OOK.

and synchronization, which is a great challenge to IC design in current silicon technologies. As the UWB impulse signals are of extremely short duration, normally less than 1 nanosecond, precise control of the PPM time-delay in terms of several tens of picosecond becomes almost impractical for integrated circuits. This problem occurs also in the template generation. Variations in either process or temperature will ruin the timing condition and make PPM inefficient. Therefore, in the real implementation, PPM is often not favorable comparing to phase shift keying (PSK) and on-off keying (OOK) [26, 27]. Figure 2.5 shows the transmitted signals under anti-phase PSK and OOK. They are actually two special cases of pulse amplitude modulation (PAM). As two alternatives of PPM,

they are of comparable performances and easier to realize [27].

2.2.2 UWB Standard

The IEEE standardization of UWB wireless systems contains two aspects of work: high data-rate short-range standard for data communications [28] and low data-rate standard for sensor, localization, positioning, etc. [29]. Neither standard has been enacted. For high data-rate standard, there are two comparable proposals, pulse-based proposal and multi-band OFDM proposal, which have been selected as the candidate standard. The low-rate standard is currently dominated by pulse-based proposals.

The pulse-based standard, direct sequence code division multiple access (DS-CDMA), is a modified and improved architecture originates from the idea of impulse radio [23]. To avoid potential interferences with existing wireless LAN (WLAN) standards, it does not employ the bandwidth of 5.15–5.825-GHz. As a result, it has three spectral modes of operation: low band (3.1–5.15-GHz), high band (5.825–10.6-GHz) and multi-band. M -ary bi-orthogonal keying (MBOK) modulation is in principle similar to the BPSK/QPSK. The receiver is correlation based, with I/Q channel for bi-orthogonal demodulation. This proposal inherited the advantages of impulse radio, the pulse-based nature permits much design flexibilities and wide range of potential applications.

The multi-band orthogonal frequency division multiplexing (OFDM) is of different concept from impulse radio [30, 31]. Rather than a pulse based transmission technology, it regards UWB as just a way of utilizing the available 7.5-GHz bandwidth. From this point of view, the multi-band OFDM proposal concerns much on the requirement of easy implementation and is close to existing

wireless technologies. It divides the available spectrum to 13 sub-bands, each of them has 528-MHz to meet FCC's regulation. The data transmissions across all bands are interleaved. In each band, OFDM is employed for data modulation. As a result, the transmission contains a lot of carrier frequencies, and the transceiver architecture is similar to that of conventional OFDM systems. This makes the multi-band OFDM UWB system easier to be implemented, and a lot of proven techniques used in narrowband transceiver can be adopted, which is especially time- and cost-efficient for industrial developments.

In brief, the two proposals are from different concepts and differ severely from each other. The pulse-based approach is ideal from communication point of view while multi-band OFDM approach is better in terms of circuit implementations. Neither can get through the final selection, which has been voted for several times since 2003 [28]. Nowadays it seems both proposals are in the development, both have demonstrated their integrated transceivers [32, 33]. Maybe only can the market determine the future direction of UWB standards.

2.2.3 Wideband Performance Variations

In spite of the standard issues, the system studied in this thesis is basically an impulse radio UWB transceiver. Single-band pulses occupying the whole available spectrum are employed for the transmission, which permits maximum available flexibilities of the UWB technology, and is applicable to a wide range of systems [8, 11, 12, 13, 21]. Meanwhile, the transceiver front-end of impulse radio is rather less explored comparing to that of OFDM front-end. The system issues for impulse radio UWB front-end are unique and of higher interests for academic research.

From Figure 2.4, the UWB impulse radio front-end contains several blocks: transmit and receive antennas, T/R switch, transmitter driver (power amplifier) and low-noise amplifier. Multi-path channel should also be included since the antenna and propagation are not separable in many cases [34].

The issues regarding impulse radio front-end implementation include many topics [8, 21, 24]. Among them an unique problem due to the ultra-wide bandwidth is the frequency-domain performance ripple.

In principle, UWB impulse radio is a time-domain technology. Besides the problems of timing and synchronization, the time-domain pulse shape is also of significant impact on the system performance. Ideally, the effect of various monocycle pulse shapes on the system performance in terms of bit-error rate (BER) can be studied through either theoretical analysis or simulation [35, 36, 37, 38]. Gaussian second-order derivative is proposed in most studies [9, 10, 35]. However, in the real system scenario, the received waveform is much different from the transmitted one. This is due to the frequency-domain gain fluctuation or ripple of realistic UWB front-end.

The problem of gain flatness occurs only in wideband or ultra-wideband systems, where it is impractical for front-end components to have flat characteristics over the exceptional wide bandwidth. Traditional narrowband front-end does not encounter this problem because it operates only over a small portion of bandwidth. The gain variation in the frequency domain results in a pulse shape distortion effect in the time domain. For impulse radio receiver whose demodulation is based on the pulse shapes, such effect is of severe impact on the system performance. Furthermore, some front-end components are of un-flat gain characteristics in nature, e.g. the propagation loss is frequency depen-

dent and can not be flat, which is decided by physics [34, 39]. Other front-end components can be designed and the gain flatness is relatively controllable, but flat gain is still a big challenge considering the system integration. Meanwhile, the phase of gain, normally evaluated by group delay, is also a factor that can distort the time-domain pulse shapes.

Therefore, performance evaluation of UWB system with realistic components is a stringent topic. Previous work has been done from various aspects including realistic antenna, coding, channel, or architectures [10, 40, 41, 42]. However, there is no approach to realistic system evaluation considering these issues. Chapter 3 proposes an implementation oriented approach and demonstrates it on the system performance evaluation considering realistic antenna and deterministic channels.

The performance degradations due to realistic components influence the efficiency of modulation schemes. In the real system environment, the most efficient modulation scheme may not be the same as that in ideal analysis. Chapter 4 studies the power efficiency of different modulation schemes for an impulse radio UWB for body-area network applications.

2.3 Integrated Circuits for UWB Transceiver

System-on-chip (SoC) and system-in-package (SiP) are currently the trend of electronic system implementations. The idea is to integrate all components on a single chip or in a chip and its package. SoC/SiP is an ideal approach to the implementations of low power and portable devices. For a radio transceiver SoC/SiP, the most challenging part lies in the integration of RF front-end. This

is particularly true for UWB front-end that must be able to accommodate high frequency and ultra-wideband requirements. For the receiver front-end, noise performance is also crucial as it basically determines the noise performance of the whole receiver.

The semiconductor technology determines the fundamental limit of RF circuits. State-of-the-art RF IC design is often based on CMOS or SiGe BiCMOS technology. After several years' development since mid-1990s, the CMOS RF IC has been matured and fully integrated transceiver SoC designed in CMOS has been demonstrated [43]. The CMOS design is of relatively lower cost and can be easily integrated with digital and mixed-signal circuits. However, the CMOS transistor is normally of lower transconductance under same current driving, and the noise performance of CMOS transistors is less superior comparing with that of bipolar transistors. Recently, the SiGe BiCMOS technology has begun to receive more and more interests for RF IC design. Comparing with silicon CMOS, SiGe BiCMOS is of higher cutoff frequency (f_T) and lower noise, and thus it is more suitable for high frequency applications e.g. RF and millimeter-wave ICs [44]. The cost of SiGe BiCMOS is relatively higher, which make it less preferred if CMOS solution is available.

The development of integrated UWB transceiver began from 2003. Most of reported designs are focused on the RF front-end, especially the receiver front-end, which is regarded as the most challenging circuits among all other analog/mixed-signal and digital parts.

In 2003–2004, the reported designs were focused on individual building blocks [45, 46]. From 2004, the design of integrated UWB transceiver has been undergoing rapid developments. As conventional transceiver architectures and

circuits can be adopted directly for multi-band OFDM design, most of the early reported UWB transceivers are based on this scheme. A 3.1–4.8-GHz multi-band OFDM receiver designed 0.25- μm SiGe BiCMOS is reported in [47, 48]. The design is focused on the interference robustness to 2.4- and 5-GHz wireless systems. A 3.1–8.2-GHz multi-band OFDM receiver designed in 0.18- μm SiGe BiCMOS is reported in [49, 50]. The design follows the direct conversion architecture and achieves very low noise figure of 3.3–4.1-dB. A 3.1–4.8-GHz 0.13- μm CMOS transceiver for multi-band OFDM UWB is reported in [51, 52], the whole transceiver occupies only 1-mm². A 0.18- μm SiGe BiCMOS UWB front-end is reported in [53], the chip is packaged and bond-wire inductances are employed in the design. A 0.18- μm CMOS direct sequence spread spectrum (DSSS) UWB transceiver is reported in [54], which is the first reported pulse-based UWB transceiver capable of 1-Gbps chip rate.

Since 2005, the pulse-based transceiver design has begun its rapid growth. An impulse radio UWB transceiver for low data-rate and accurate ranging applications is reported in [55, 56], where sub-mW operation is permitted and realized. A 0.18- μm CMOS analog correlation receiver is reported in [57]. The original idea of impulse radio is implemented in [58]. These impulse radio transceivers normally follow the typical architecture of Figure 2.4 with slight modifications. Due to the short duration of high frequency pulses, these designs normally employ only the lower band of UWB. The multi-band OFDM transceivers are also developed towards further integration and sophisticated architectures [59, 60, 61]. In particular, a fully integrated UWB physical layer IC based on multi-band OFDM has been implemented [62].

Since pulse-based UWB is a new concept and the transceiver requires new architecture, the design of pulse-based UWB transceiver is often more chal-

lenging and is of higher value in academic research. All RF components in impulse radio front-end have to accommodate the full-bandwidth operations rather than 528-MHz sub-bandwidth for some of RF components in multi-band OFDM front-end. In circuit part of the thesis, the design of UWB transceiver front-end RF components, LNA and T/R switch, is presented in Chapter 5, Chapter 6 and Chapter 7. An inductorless UWB LNA design technique is proposed and a high-performance T/R switch is demonstrated.

Chapter 3

Performance Analysis of Realistic UWB System

3.1 Introduction

In the system level, an unique challenge to UWB radio implementation is the performance fluctuation over the exceptionally wide operating bandwidth of transceiver front-ends. Traditional NB front-end does not encounter this problem because it operates only over a fractional bandwidth as small as 1%; performances of individual blocks are almost invariable within the band, which ensures small distortion of carrier waveforms. Unfortunately, broadband design techniques are often restrained and confined by manufacture process and cost; as a result, it is impractical to obtain an unrippled performance over the UWB operating bandwidth. In the frequency domain, it exhibits a fluctuation of the transfer function; while it shows a waveform distortion in the time domain.

Comparing with the front-end circuits that can be designed to minimize the

performance fluctuations, the channel is not able to be controlled and its frequency dependence is decided by physics [34, 39]. It is important to investigate the effect on the overall system performance. In most cases, the antenna and channel should be considered as a whole. Therefore, this chapter studies the performance of realistic UWB system with realistic antennas in deterministic multipath channels.

Antenna is the most front stage transmitting and receiving UWB signals. The performance of antenna affects the overall UWB systems performance primarily. As the waveform distortion is inevitable for a given antenna [63], the effects of realistic antennas can be investigated through waveform distortion in either time or frequency domain.

Deterministic multipath channel models are also becoming more and more appropriate in modern wireless system design. Traditional channel models describe the electromagnetic wave propagation environments statistically; wireless systems employing such models will encounter a performance degradation in any specific environment because the statistic nature does not offer exact solution to any specific circumstance. Fortunately, it is now possible to use deterministic multipath channel models in the wireless systems with the development of computing technology. For instance, the specific propagation environment for a wireless system can be stored firstly; and deterministic channel model can be given by a first-time calculation. Subsequently, the wireless system can be designed or adapted to the specific environment so as to achieve the best performance. This idea is particularly useful for certain wireless applications such as accurate locating, imaging, ranging, etc..

Problems with regard to antennas and deterministic multipath channels

have been investigated to a certain extent [63, 64, 65, 66], whereas the investigation is often without considering other functional blocks. The multipath channels are modeled by assuming the antenna has a simple differentiation operation [65] or without considering antennas [66], while the antenna approximations and waveform optimizations are studied with a direct line-of-sight (LOS) path [63, 64]. Furthermore, the impact of realistic antenna and channel on overall system performance is still not quantitatively clear.

The performance of UWB impulse radio is evaluated in terms of error probability. The proposed method is firstly described in Section 3.2 for the evaluation of BER performance of realistic UWB systems. The approach employs waveform distortion analysis in the frequency domain to investigate the effect of the realistic block performance fluctuation and evaluate the overall system performance in the time domain. In Section 3.3, transfer function calculations of antennas and deterministic multipath channels are described. Section 3.4 presents the numerical results of a UWB system considering the realistic antennas and deterministic channels of Section 3.3. The optimal selection of realistic antennas in different propagation environment is also demonstrated from the point of view of system implementation.

3.2 Methodology of Analysis

Consider a typical UWB impulse radio system with time-hopping PPM and correlator-based receiver, as described in Section 2.2.1, when it is in single user operation with additive white Gaussian noise (AWGN), the BER performance

can be expressed as [9, 10, 67]

$$P_e = Q\left(\frac{a_0 - a_1}{2\sigma_0}\right), \quad (3.1)$$

where a_0 , a_1 and σ_0 are the integral products of the template signal with received signal and Gaussian noise, which are given by

$$a_0 = \int_0^{T_f} w(t)v(t)dt, \quad (3.2)$$

$$a_1 = \int_0^{T_f} w(t - \delta)v(t)dt, \quad (3.3)$$

$$\sigma_0 = \sqrt{\frac{N_0}{2} \int_0^{T_f} v^2(t)dt}. \quad (3.4)$$

Here N_0 is the noise PSD. The Q function is defined by

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp(-u^2/2)du. \quad (3.5)$$

It is clear that BER of this UWB system is a function of the received waveform $w(t)$, modulation parameter δ and noise PSD N_0 . Furthermore, δ can be optimized in single user environment by [9]

$$\delta_{opt} = \operatorname{argmin}_\delta \int_{-\infty}^\infty w(t)w(t - \delta)dt. \quad (3.6)$$

That is, δ is also a function of the pulse shape $w(t)$. Therefore, BER of this UWB system is eventually determined by the received waveform $w(t)$ and noise PSD N_0 .

To get the best BER performance, the pulse shapes should be carefully

Figure 3.1: Waveform of Gaussian second-order derivative in the time domain. The amplitude of the waveform is normalized to 1 V.

designed. Previous research indicates the UWB system employing Gaussian second-order derivative waveform, as used by Scholtz and colleagues, achieves better BER performance [35, 67]. The Gaussian second-order derivative pulse can be represented as

$$w(t) = A \left[1 - 4\pi \left(\frac{t}{\tau} \right)^2 \right] \exp \left[-2\pi \left(\frac{t}{\tau} \right)^2 \right], \quad (3.7)$$

where τ is the time constant which controls the pulse duration and resultant bandwidth. Figure 3.1 shows the waveform of a Gaussian second-order derivative whose duration in the time domain is about 0.37 ns and -10 dB bandwidth in the frequency domain is about 7 GHz.

Note that the analysis described above is obtained under the assumption of ideal transceiver characteristics, where the received waveform is identical to the

Figure 3.2: Flowchart for analyzing waveform distortion and system performance considering nonidealities in the individual blocks.

transmitted one. However, in the practical wireless transmission, the received waveform will definitely be distorted by the nonideal components, e.g. realistic antennas and multipath channels.

To investigate the nonideal effects in practical implementation, it is convenient to use frequency domain method since the performance of individual block is often evaluated in the frequency domain. The analysis methodology is shown in Figure 3.2. The block level characteristics are modeled as transfer functions in the frequency domain; while the performance of overall systems is considered in the time domain. Fourier transform and inverse Fourier transform are employed whenever needed. The idea of this approach is to investigate the waveform distortion along the signal transmission and processing path; this is because of the following two principles.

1. The performance fluctuation in the frequency domain exhibits a waveform distortion effect in the time domain. Nonidealities of the individual blocks can be considered through waveform distortion analysis.

2. The overall system performance is determined by the received waveform eventually.

Therefore, waveform distortion is the key point which contains information of both realistic blocks and overall system performance in terms of BER. By this approach, realistic block can be investigated from the system point of view as long as it can be described by a transfer function along the signal path, and interactive design between system level and component level is viable.

This chapter demonstrates the analysis of UWB impulse radio systems considering realistic antennas and deterministic multipath channels through the above method. A single-link PPM UWB transceiver is discussed and Gaussian second-order derivative pulse is selected as the transmitted waveform. It should be pointed out that multiple-access UWB impulse radio systems can also be investigated by this approach since the BER of these systems are also determined by received waveforms under certain assumptions [9]. Moreover, this method can also be employed to evaluate other building blocks such as low-noise amplifier, pulse generator, correlator, etc..

3.3 Antennas and Deterministic Channels

3.3.1 Antennas

Transfer function of short dipole antenna has been derived in [63]. However, to derive closed-form expressions of transfer functions is not always practical for antennas with complicated structures. Numerical method is often employed. In this chapter, the antenna transfer functions are obtained by method of moments

Figure 3.3: Structures and dimensions of 3 planar antennas for UWB transmission. The unit of scales is meter.

(MoM) [68]. The metal antenna under study is firstly divided into small triangles (mesh) and the Maxwell equations are applied to each mesh; the overall parameters are obtained by summing the contributions over all meshes. Note that frequency sweep is used to deal with the ultra-wide bandwidth. In the calculation, transmitting and receiving antennas need not to be identical; they can be treated separately as two port networks. In this chapter, however, we consider the same antennas used in both transmitter and receiver due to its advantage in higher-level transceiver integration [69].

Figure 3.3 shows three planar antennas for pulse-based or wideband applications, i.e. slot, diamond dipole and planar volcano-smoke slot antennas (PVSA) [70, 71, 72]. They are optimized for UWB systems and the input impedances are designed to be $50\ \Omega$. Putting each pair of the transmitting and receiving antennas in parallel with 1 meter distance, and considering only the LOS path, their voltage transfer functions can be obtained by numerical simulations. Here the transmitting antenna is driven by a source generator with $50\ \Omega$ series impedance and the receiving antenna load is also a series $50\ \Omega$. The input variable is defined as the transmitting source voltage; output variable is the voltage across the $50\ \Omega$ load. The amplitude and phase of the transfer functions are shown in Figure 3.4. It is clear that characteristics of the antennas ripple significantly over the ultra-wide bandwidth.

As discussed above, performance fluctuations in the frequency domain will certainly distort the transmitted waveform in the time domain. Figure 3.5 shows the received waveforms when a Gaussian second-order derivative pulse of Figure 3.1 is used to drive or excite the input antennas with the transfer functions of Figure 3.4. Although the above selected antennas are optimized to have ultra-wide bandwidth, it is still not practical to keep all frequency com-

Figure 3.4: Voltage transfer functions of the three antenna pairs. Each pair of transmitting and receiving antennas are identical and placed in parallel with 1 meter distance, only the LOS path is considered. The input variable is defined as the transmitting source voltage with $50\ \Omega$ series impedance, the output variable is the voltage across the $50\ \Omega$ load.

Figure 3.5: Received waveform by realistic antennas. The Gaussian second-order derivative pulse of Figure 3.1 is distorted by the antenna pairs with transfer functions of Figure 3.4. There is also a 3.3 ns time delay due to the 1 m LOS path.

ponents to scale and recover the signals without distortions. As the result, the UWB system performance will be degraded, which is discussed in Sec 3.4.

3.3.2 Deterministic Multipath Channels

The deterministic multipath channels are modeled with ray-tracing method [66]. For a given i^{th} ray, the transfer function can be represented as [39]

$$H_i(f) = \left(\frac{\lambda}{4\pi} \right) D_i(f) \frac{e^{-jkr_i}}{r_i}, \quad (3.8)$$

where r_i is the length of the ray, $k = 2\pi f \sqrt{\mu\epsilon}$ is the wavenumber, or propagation constant of the medium. μ and ϵ are the permeability and permittivity of the medium. $\lambda = v_p/f$ is the wavelength. v_p is the speed of wave in the medium. Obviously, k and λ are the functions of frequency f . $D_i(f)$ models the effects of reflections and diffractions in the propagation path, which is given by

$$D_i(f) = \prod_{m=1}^{R_i} D_{(i,m)}(f), \quad (3.9)$$

where R_i is the total number of reflections and diffractions in the i^{th} ray, $D_{(i,m)}(f)$ is the coefficient of the m^{th} reflection or diffraction. Note that $D_{(i,m)}(f)$ is also determined by the geometric dimensions and angles of reflections or diffractions.

Therefore, for a deterministic multipath channel with N_r rays, the transfer function can be calculated by tracing each ray, which is expressed as

$$H(f) = \left(\frac{\lambda}{4\pi} \right) \sum_{i=1}^{N_r} D_i(f) \frac{e^{-jkr_i}}{r_i}. \quad (3.10)$$

The characteristics of deterministic multipath channels can be employed to investigate the waveform distortion and performance degradation of practical UWB systems. Statistical models can not do so because they do not specially represent any realistic propagation environments. Here we choose four propagation environments which represent different applications of UWB radios, and their characteristics are discussed below.

LOS Path

The LOS path models the signal propagation in free space or in various spaces with the first Fresnel zone clearance that reflection and diffraction are not taken into account [73]. Transfer function of a LOS path is simply represented as

$$H(f) = \left(\frac{\lambda}{4\pi} \right) \frac{e^{-jkr}}{r}, \quad (3.11)$$

where r is the distance between the transmitting and receiving antenna. In the previous antenna discussion, we use $r=1\text{m}$. H is the function of frequency through k and λ . In this case, H curve is not flat even if no multipath components occurred.

Two-Ray Channel

The two-ray channel used here is simply a LOS path with a ground reflection path which models the outdoor open environment. The transfer function of a two-ray path is given by

$$H(f) = \left(\frac{\lambda}{4\pi} \right) \left(\frac{e^{-jkr_1}}{r_1} + \Gamma(\theta) \frac{e^{-jkr_2}}{r_2} \right), \quad (3.12)$$

where r_1 is the length of LOS path, r_2 is the length of ground reflection path. Γ is the reflection coefficient by the ground where the incident angle is θ .

In this chapter, the transmitting and receiving antennas are placed 6 meter away with same height of 1 meter. The reflected path is about 0.3 meter longer than the LOS path, thus, waveforms in this path are delayed by about 1 nanosecond.

Figure 3.6: Configuration of deterministic multipath channels. The unit of scales is meter.

Urban Environment

Figure 3.6(a) shows a typical urban propagation environment [66]. There is no direct path from transmitter to receiver. Each path contains reflection and/or diffraction by the buildings or other blocks. From (3.10), the transfer function can be calculated by ray-tracing method, which can be done in either time or frequency domain.

Inter-Floor Propagation

Figure 3.6(b) shows the propagation environment between different floors in a building [74]. In this case, it is important to consider the inter-floor diffraction paths. The diffraction rays are also shown in Figure 3.6(b). The transfer function can be obtained similarly from (3.10).

Multipath channels introduce both waveform distortion and time delay. The

latter effect results in the inter-symbol interferences (ISI), which further distorts the transmitted waveforms. For example, in the two-ray channel, 1 meter difference between the two paths leads to 3.3 ns time delay of the received waveform. In case the time interval of two contiguous transmitted pulse are 3.3 ns, the delayed waveform of the previous pulse, from the reflected path, arrives at the same time with the later transmitted pulse from LOS path. They are then overlapped and the later transmitted waveform is collapsed. As a result, the overall system performance is further degraded. In fact, the pulse-based UWB systems exhibit less overlapping than continuous-carrier systems, since the pulse width is very short, which prevents destructive interferences from multipath.

Precise model of ISI problem in a pulse based UWB system may not be practical with a general channel. In this chapter, however, the channels are deterministic, the ISI effect can be investigated in the particular environments.

3.4 Numerical Results and Discussions

Based on the analysis methodology described in Section 3.2 and individual blocks characterization discussed in Section 3.3, numerical simulation is performed which gives the results of waveform distortion and BER of overall realistic UWB impulse radio system.

As mentioned previously, a single-user UWB impulse radio system is considered without loss of generality, which includes a PPM modulator, correlation receiver, and realistic antennas in deterministic multipath channels. The Gaussian second-order derivative pulse is employed to drive the transmitting an-

Figure 3.7: Received waveforms by diamond dipole antennas in the two-ray channel. Comparing with the original transmitted waveform of Figure 3.1, the received waveform encounters a significant distortion because of the nonideal antennas and multipath channels. The 0.32 m longer reflected path leads to a 1.08 ns delay between the two rays. The pulse duration is too short comparing to the inter-path time delay.

tenna, whose amplitude is normalized to 1 V. Both source and load impedance of antenna are matched to 50Ω . Perfect synchronization is assumed. No error control coding is used which gives us the performance evaluation of the realistic transceiver front-end.

3.4.1 Received Waveform Distortion

Figure 3.7 and Figure 3.8 show the received waveforms considering both realistic antennas and various deterministic multipath channels. To make the waveforms clear, we only give the examples of diamond dipole antenna in each multipath channel. For better intelligibility, Figure 3.8 shows only the strongest waveforms received among all multipath components. Due to the different time

(a) Urban environment of Figure 3.6(a).

(b) Inter-floor environment of Figure 3.6(b).

Figure 3.8: Received waveforms by diamond dipole antennas in the urban and inter-floor deterministic multipath environments. Note that due to the visibility of the figures, only the strongest waveforms are displayed among all multipath components.

delay and reflection/diffraction in the propagation, the signal from each ray arrives at the different time point and exhibits different distortion in terms of both shape and amplitude. Note that ISI is not considered here. Since the UWB pulse duration is very short, the multipath components from same transmission normally do not interfere with each other. For instance, about 0.3 m distance will produce 1 ns time delay. If the difference of two rays is larger than 0.3 m and pulse duration is smaller than 1 ns, the signal from two rays will be separated clearly. In the two-ray channel, the reflected path is 0.32 m longer than the LOS path, resulting in 1.08 ns time delay, which could be observed clearly from Figure 3.7. The fine time resolution of UWB signal is particularly useful for applications such as radar, ranging and imaging systems.

Figure 3.7 and Figure 3.8 give also the actual arrival time of rays. The time zero is defined when the exciting Gaussian derivative pulse reaches the maximum value, corresponding to Figure 3.1.

3.4.2 ISI Distorted Waveform

As mentioned previously, the deterministic channel permits a more accurate ISI simulation under particular environment. Here we consider the system with diamond dipole antenna in the two-ray channel, the received waveform without ISI is Figure 3.7. Assume the symbol rate is 1 G symbols per second, that is, transmitting time interval between two symbols are 1 ns. The reflected waveform from the previous symbol (interferer) will overlap with the direct waveform from the later symbol. In this case, the waveform is distorted as shown in Figure 3.9. It is clear that overlap of symbols distorts the waveform significantly, leading to a serious performance degradation.

Figure 3.9: Waveform distortion considering ISI with diamond dipole antenna in the two-ray channel. The ISI occurs by assuming both symbol time interval and time delay due to reflection are about 1 ns. Comparing with Figure 3.7, the ISI further distorts the received waveform in both of the two rays.

3.4.3 BER Performance

With the received waveform obtained, the overall UWB system performance can be calculated from (3.1). Figure 3.10, 3.11, 3.12 and 3.13 show the performance of overall UWB impulse radio system in terms of BER considering realistic antennas in various deterministic multipath environments. Note that E_b in the figure represents the bit energy of the received waveform rather than transmitted waveform. It is clear that UWB system with slot antenna achieves better performance than with other two antennas in LOS channel. However, the difference in this case is slight. In other three cases, diamond dipole antenna performs much better.

A possible explanation is that transfer function of diamond dipole antenna is much flatter than slot antenna and PVSA. This does not further degrade the

Figure 3.10: BER of overall UWB impulse radio system considering realistic antennas in the LOS channel.

Figure 3.11: BER of overall UWB impulse radio system considering realistic antennas in the two-ray channel.

Figure 3.12: BER of overall UWB impulse radio system considering realistic antennas in the urban environment of Figure 3.6(a).

Figure 3.13: BER of overall UWB impulse radio system considering realistic antennas in the inter-floor environment of Figure 3.6(b).

waveform distortion resulting from channels. On the other hand, diamond dipole antenna has less voltage gain comparing to slot antenna and PVSA, which requires a driving pulse with larger amplitude. Note that this result is not a general conclusion since the channel and input pulse are both deterministic. In other environments, the result may be different and should be re-evaluated. This chapter demonstrates the optimal selection of antenna in the real system design and implementation, which can be easily adopted for other pulse based UWB systems.

Figure 3.11 gives also the BER result in the presence of ISI. It is clear that BER increases when symbols overlap. ISI is a severe problem in the real system design, the symbol rate and pulse shape should be carefully designed to minimize the ISI effect. In the particular environment that deterministic channel can be applied, it is possible to observe and evaluate the overlap with this method. Note that ISI demonstrated in this chapter is in a relatively worse case, since the interferer overlaps completely with the signal and its energy is comparable to LOS components.

Based on this result, the transfer function characterization should be as flat as possible. However, it may not be achieved in practical implementation. One approach is to compensate the performance fluctuation at following stages. For example, employing a stage with complementary fluctuation of transfer function can obtain a much flatter transfer function of the overall stage. These blocks should be co-designed to achieve a better performance of the overall system.

Chapter 4

UWB for Body-Area Networks

4.1 Introduction

A body area network (BAN) is a network with its nodes placed close to the body on or in everyday clothing [75]. A WBAN employs wireless technology to realize the connectivity among nodes [76], which is proposed for a wide range of lifestyle and medical applications. The emerging UWB impulse radio is a promising technology for WBAN due to its low power and wideband characteristics [7, 9, 10]. According to the FCC regulations that approved the use of UWB in USA, the mean transmit power of these devices must not exceed -41.3 dBm/MHz from 3.1 to 10.6 GHz [5]. 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. [77, 78], which basically require low transmit power together with high accuracy. All of these features make UWB an ideal candidate for WBAN applications.

Since the WBAN devices are attached and operate around the human body, the human body effect becomes a crucial part of radio propagation channel. Measurements and characterization of the on-human-body channel are essential for the investigation. Furthermore, the impacts of on-human-body channel on the performance of the overall WBAN transceiver are highly demanded in the WBAN system design and implementation.

The UWB antenna and channel characterization considering human body effects were conducted for wireless personal area network (WPAN) applications [79, 80, 81]. The effect of the body on UWB signal propagation was measured for one antenna placed on the body and the other antenna kept away from the body. There are only a few papers that report channel measurements and modeling with both transmit and receive antennas placed on the body for WBAN applications [27, 29, 82, 83, 84, 85, 86, 87, 88]. Review of these studies shows that no measurement and modeling that cover the whole UWB band from 3.1 to 10.6 GHz has been made. Also, the reported channels were all sparsely sounded on a few pre-defined points. The important statistical characteristics of the channels had to be extracted from the limited number of samples or FDTD simulations.

The impact of on-human-body channel on the overall UWB impulse radio WBAN system is not clear. As the operation band covers several gigahertz range, the ripple or fluctuation in the frequency response of the channel leads to a comprehensive effects on the overall WBAN system. However, there is few paper dealing with this kind of issues currently.

This chapter presents the study of on-human-body UWB channels for WBAN applications, which contains two parts of work: the characterization of the

channel itself, and the impact of the on-human-body channel on the whole UWB WBAN system.

Section 4.2 discusses a UWB impulse radio WBAN system with various candidate monocycle pulse shapes and modulation schemes. A realistic system BER evaluation method based on waveform distortion analysis is studied. Section 4.3 presents the measurement and characterization of on-human-body channel. Based on the method discussed in Section 4.2 and the measurement results obtained in Section 4.3, the overall UWB WBAN system performance with on-human-body channel is evaluated, the results are summarized and discussed in Section 4.4.

4.2 Impulse Radio WBAN

UWB impulse radio is basically a time-domain method employing discrete monocycle pulses for data transmission. Therefore, the pulse shape and modulation scheme affect the system performance directly. For binary modulation, there are three widely proposed schemes: pulse position modulation [9], phase-shift keying and on-off keying [26]. The binary antipodal (anti-phase) PSK and OOK are actually also special cases of pulse amplitude modulation. All of these three schemes require correlation-based demodulation, where the received monocycle is multiplied by the template monocycle and then integrated for a binary decision. If the pulse shape is not distorted by the channel, BER performance of BPSK and OOK will keep independent on the pulse shape employed. BER of PPM is always dependent to the pulse shape, thus pulse shape design is necessary for PPM [35].

However, the received pulse shape is inevitably distorted by the varied frequency responses of realistic UWB system. This can be clearly observed from Chapter 3. As a result, BER performance of these UWB modulations will be degraded and depends on the received (and thus transmitted) pulse shapes. In a WBAN scenario, the most crucial part along the signal path that results in waveform distortion is the on-human-body channel, where the fluctuation of wideband channel characteristics is inherently decided by physics.

Similar to the evaluation discussed in Section 3.2, frequency domain method of Figure 3.2 is employed for the analysis of real WBAN impulse radio system. Note that in this chapter, as the transfer functions used are from real-scenario measurements, the separation of WBAN channel and antennas is not practical. The transfer function $G(f)$ is the comprehensive effect including transmit antenna, on-human-body channel and receive antenna.

As the BER analysis is deterministic, it is desirable to find the most typical and representative channel characteristics for the evaluation. Various modulation schemes and pulse shapes for the transmission need to be compared under the given representative WBAN channels.

The impulse radio WBAN with three transmitted pulse shapes are evaluated: Gaussian first-order derivative, Gaussian second-order derivative and rectangular pulse. The Gaussian derivatives are proposed in terms of their ultra-wideband power spectrum density, especially for Gaussian second-order derivative, as discussed in Section 3.2. The rectangular pulse is easy to realize in terms of hardware complexity and power consumption, the latter is a basic requirement for WBAN device to reduce potential hazards.

In this chapter, binary PPM, PSK and OOK modulations are compared. From

the viewpoint of communication theory, OOK is always 3-dB less power efficient than binary PSK. But OOK demodulation can be very simple in hardware realization [27]; the received waveform can be used also as the template monocycle for demodulation. Thus, OOK demodulation becomes a process of energy detection. The BER to be obtained for OOK in this chapter is based on energy detection. It is predicted that BER result of OOK does not depend on the received waveform.

4.3 UWB On-Human-Body Channel

Radio propagation channel measurement and characterization have always been recognized as having an important part to play in the development of complex wireless networks. This is particularly true for UWB radio [81, 89, 90, 91].

The channel measurements were devised to generate the radiographs of path loss and delay spread for the first time. The channel parameters as path loss and delay spread are extracted and the statistical distributions of the channel variations are determined from the radiographs. Considering that a human body has a finite size, we conducted channel measurements on and in close proximity to the body at a much higher spatial resolution. The sampling distance was set to be 2 cm, which is shorter than the minimum wavelength within the frequency range. We believe that these measurements represent the most extensive set of publicly reported measurements taken to characterize on-human-body UWB radio channel.

4.3.1 Description of Measurements

The measurements were carried out using a 40-GHz HP ES-8510 network analyzer. The antennas used in the measurements were two planar monopoles [92]. The shape and geometry of the antennas are shown in Figure 4.1. The antennas are made in silver on a $\epsilon_r = 5.9$ substrate. Figure 4.2 shows the measured antenna return loss on and off the human body. As shown, the antenna return loss is above -10 dB from 3.1 to 3.5 GHz and is below -10 dB from 3.5 to 10.6 GHz for the off-body case. The antenna return loss is improved to below -10 dB for the on-body case over the entire UWB band. This is due to the lossy nature of the body. Figure 4.3 shows the measured antenna far-field radiation patterns for both E- and H-planes at the UWB central frequency 6.85 GHz. It is seen that the antenna has an omni directional radiation pattern in the E-plane and an ∞ shaped radiation pattern in the H-plane. It is known that the antenna radiation patterns will change when it is placed on the body [93]. Unfortunately, the change cannot be measured with our testing facilities. The WBAN device in the real scenario suffers also from the unstable antenna radiations, as a result, such effects can not be ignored in the measurement.

The setup can measure the transfer function of the channel. The amplitude loss and phase shift of each frequency component caused by the channel were recorded. With 4.6875 MHz steps, 1601 frequency points were recorded over the frequency range of 3.1–10.6-GHz. That is, a multipath with a time delay up to 213 ns can be detected, which is suitable for indoor environments [94]. The time domain resolution corresponds to the ability to resolve two closely spaced responses. There are abrupt transitions in a frequency domain measurement at the start and stop frequencies, which cause overshoot and ringing in

4.3 UWB On-Human-Body Channel

Figure 4.1: The antenna used in the measurement of on-human-body channel. It is made in silver on a 0.8mm-thick substrate with $\epsilon_r = 5.9$.

Figure 4.2: The antenna and its measured return loss. The antenna return loss is improved to below -10 dB for the on-body cases over the entire UWB band due to the lossy nature of the body.

Figure 4.3: The measured antenna radiation patterns at 6.85-GHz. It has an omni directional radiation pattern in the E-plane and an ∞ shaped radiation pattern in the H-plane.

the time domain response. The frequency domain windowing can smoothen the abruptness of the frequency domain transitions but at the cost of the time domain resolution. For UWB channel sounding, the time domain resolution is important and critical due to narrow pulses, consequently, a minimum window is desirable [91]. Using the minimum window, one can find the time domain resolution to be 133 ps.

The measurements were conducted in both anechoic chamber and a staff lounge room, as shown in Figure 4.4. The staff lounge room is a typical indoor environment. The main rationale of choosing the staff lounge room and the anechoic chamber was to investigate the influence of the environment and to expose the impact of the human body on the UWB signal propagation.

The measurements were made on or in close proximity to each body. During the measurements, the person under test was to stand straight and upright at a

Figure 4.4: The measurement environment of (a) anechoic chamber and (b) staff lounge room.

fixed location and the transmit antenna was placed on the right upper arm near the shoulder. The transmit antenna placed there was because a longer path and a better human safety are offered. Figure 4.5 shows the defined points to generate the radiographs of path loss and delay spread. The receive antenna was moved to the points to collect the data. The horizontal step is 8 cm and the vertical step is 2 cm. Note that not all the Rx points are placed on the body, a cylindrical distribution of test points are performed, with the perimeter of 1 m to make most test points in trunk placed on the body.

The measured result is severely related to the orientation of antennas. In this measurement campaign, both transmit and receive antennas are placed in parallel to the nearest skin of human body. This is because that WBAN devices need to be wearable in most scenarios. The transmit antenna is fixed on clothes of right upper arm. At the point of measurement, the receive antenna is carefully rotated within the parallel plane to the body to minimize polarization loss with respect to the transmit antenna.

Figure 4.5: Location of test points around human body. All test points form a cylinder with 1 m perimeter.

4.3.2 Statistical Parameters of the Channel

A total of 2930 frequency transfer functions were recorded of which 2730 were recorded in the staff lounge room and 200 in the anechoic chamber. Each frequency transfer function consists of 1601 frequency points.

Figure 4.6 plots the transfer functions measured at test point A in both anechoic chamber and staff lounge room. The loss is relatively small in the staff lounge room due to the reflection paths by the environment, this is particularly clear at frequencies lower than 6-GHz. However, the fluctuations will certainly affect the system performance.

The path loss in decibel can be directly calculated from the measured transfer function of the channel as [81]

$$PL(d) = 10 \log_{10} \left[\frac{1}{MN} \sum_{j=1}^M \sum_{i=1}^N |H_j^d(f_i)|^2 \right]. \quad (4.1)$$

$H_j^d(f_i)$ denotes the j^{th} transfer function of the channel at a frequency f_i at a distance d . M is the number of transfer functions for the distance d , and N is the number of frequency components in the transfer function of the channel [82, 84]. The average path loss can be expressed as

$$PL(d) = PL_0 + 10n \log_{10} \left(\frac{d}{d_0} \right) + X_\sigma, \quad (4.2)$$

where PL_0 is the path loss at the close-in reference distance d_0 , n is the path loss exponent, and X_σ is the shadowing fading in dB. We set d_0 to be 1 m for better comparison with available results and find the values for PL_0 , n , and X_σ from equation (4.1).

Figure 4.6: Transfer functions at test point A in Figure 4.5 in anechoic chamber and staff lounge room. The phase is unwrapped. The loss is relatively small and amplitude fluctuations is high in the staff lounge room due to environment reflections.

Figure 4.7: The radiograph of channel path loss on and in close proximity to the body. The unit of path loss is dB. The result is obtained by the measurements in staff lounge room. The position of Tx antenna is (0, 1.2m).

Figure 4.7 shows the radiograph of channel path loss on and in close proximity to the body in an un-folded format. Note the position of Tx antenna is (0, 1.2m). The radiograph is generated from equation (4.1) based on the channel transfer functions measured in the staff lounge room. It clearly shows the qualitative information on the channel. From Figure 4.7, the LOS region is located at the top two corners in the radiograph, which is mainly the side of the body near the transmit antenna. The NLOS region is located from the top to bottom middle part of the radiograph, which is mainly the other side of the body away from the transmit antenna.

The radiograph also contains quantitative information on the channel. The channel path loss model parameters such as PL_0 , n , and X_σ defined in equation (4.2) can be extracted from it by performing a least square fit computation. Table 4.1 shows the extracted PL_0 and n value for various routes. Path loss

Table 4.1: Extracted PL_0 and n from Radiograph

Routes	Front Side		Back Side	
	PL_0 (dB)	n	PL_0 (dB)	n
Vertical	-82	2.6	-82	2.9
Horizontal	-101	3.7	-93	2.8
Diagonal	-86	2.2	-84	2.7

exponents given in [82] are 4.1 and 2.7 for the horizontal and vertical routes, respectively. They are quite closed to the results here that are 3.7 and 2.6. Note that the larger path loss exponent for the horizontal route indicates more severe diffractions around the human body. The horizontal route can be considered as the most difficult propagation path for the WBAN channel.

The rms delay spread τ_{rms} can be calculated from channel impulse responses, which are obtained by inverse Fourier transform from the transfer functions. The mean excess delay $\bar{\tau}$ or the first moment of the power delay profile with respect to the first arriving wave is given by [34]

$$\bar{\tau} = \frac{\sum_{k=1}^K \tau_k h(p, \tau_k)}{\sum_{k=1}^K h(p, \tau_k)}. \quad (4.3)$$

where $h(p, \tau_k)$ is the power delay profile obtained from measurement point p and scaled such that the first wave arrives at $\tau = 0$ in the profile. The root mean square delay spread τ_{rms} is the square root of the second central moment of the power delay profile and expressed as

$$\tau_{rms} = \sqrt{\frac{\sum_{k=1}^K \tau_k^2 h(p, \tau_k)}{\sum_{k=1}^K h(p, \tau_k)} - \bar{\tau}^2}. \quad (4.4)$$

Figure 4.8: The radiograph of channel rms delay spread on and in close proximity to the body. The unit of rms delay spread is ns. The result is obtained by the measurements in staff lounge room. The position of Tx antenna is (0, 1.2m).

It is clear that strong echoes with long delays contribute significantly to τ_{rms} .

Figure 4.8 shows the radiograph of channel rms delay spread on and in close proximity to the body in an un-folded format. The radiograph is generated from the transfer functions measured in the staff lounge room. Note that the threshold level should be carefully considered. Usually, the threshold level is set to be 30 dB down from the peak in the studies of radio propagation channels for cellular radio or wireless local area network applications. Here another approach is adopted. We consider a time interval from 0 to 100 ns to calculate and τ_{rms} . This is because echoes from obstacles fade away after this time for WBAN applications.

The radiograph of channel rms delay spread contains both qualitative and quantitative channel information. As can be seen, the rms delay spread τ_{rms}

is larger in the NLOS region and smaller in the LOS region. Since the rms delay spread values were not normally distributed, it was necessary to employ nonparametric statistical analysis. The statistics of the rms delay spread were compiled. The maximum value of the rms delay spread is 12 ns. The rms delay spread values are less than 6 ns 50% of the time and less than 9 ns 90% of the time. It is a very important parameter since the performance of wireless communications systems operating in multipath environments is very sensitive to the values of rms delay spread.

4.4 UWB WBAN Performance

Based on the analysis in Section 4.2, numerical simulation can be employed to evaluate the performance of the UWB WBAN over on-human-body channels.

The transfer functions of on-human-body channel are obtained from previous measurement. Only are a few cases out of the 2930 transfer functions selected to demonstrate the impact of on-human-body channel. According to the radiograph shown in Figure 4.7 and Figure 4.8, three representative cases are selected according to their path loss and rms delay spread characteristics.

1. The LOS component is dominant, the path loss and τ_{rms} are small. This occurs along or near right arm where the transmitting antenna is placed. The transfer function is obtained from point A in Figure 4.5 (right side near hand).
2. Both LOS and NLOS components exist and they are comparable. The path loss and τ_{rms} are near average. This occurs at the front and back of the

trunk. The transfer function is obtained from point B in Figure 4.5 (front side near heart).

3. The NLOS component is dominant, the path loss and τ_{rms} are large. This occurs at the opposite (left) side of the body. The transfer function is obtained from point C in Figure 4.5 (left side below hand).

The above three cases of on-human-body channel are evaluated in both anechoic chamber and staff lounge room. This highlights the impact of human body in different environments. The simulation setting and parameters used are the same as described in Section 4.2. As rms delay spread values are less than 6 ns 50% of the time, 6 ns window in time domain is selected for monocycle duration, corresponding to a maximum data rate (chip rate) of 167 Mbps.

Figure 4.9, 4.10 and 4.11 show the BER performance of the UWB impulse radio WBAN system over on-human-body channels in both anechoic chamber and staff lounge room. Various transmitted pulse shapes and modulation schemes described in Section 4.2 are presented. For better distinction and comparison, these BER results are categorized by their channel scenarios. It is clear that the BER performance is more sensitive to its position on human body (A, B, and C) than to the environment (anechoic chamber and staff lounge room), which indicates that the impact of human body is more significant than that of the environment.

Note that in Figure 4.9, 4.10 and 4.11, the E_b represents the bit energy of received waveform rather than transmitted waveform. This actually normalized the energy of received waveform and neglected the absolute value of path loss. The intention of using received bit energy E_b is to expose the unique impacts of human body on the WBAN radio propagation. Otherwise, the path loss

(a) anechoic chamber

(b) staff lounge room

Figure 4.9: BER performance of UWB impulse radio when the receiver is placed at point A in (a) anechoic chamber and (b) staff lounge room.

(a) anechoic chamber

(b) staff lounge room

Figure 4.10: BER performance of UWB impulse radio when the receiver is placed at point B in (a) anechoic chamber and (b) staff lounge room.

(a) anechoic chamber

(b) staff lounge room

Figure 4.11: BER performance of UWB impulse radio when the receiver is placed at point C in (a) anechoic chamber and (b) staff lounge room.

dominates the channel characteristics, and the impact of human body is less distinctive. It is predicted in this case that BER of OOK (energy detection) will be independent to the pulse shape distortion as well as the channel environment if path loss is not considered. This is shown clearly in Figure 4.9–4.11.

The BER at each point in anechoic chamber and staff lounge room indicates the impacts of on-human-body channel on the overall UWB WBAN system. A clear trend is, the more propagation path involving body, the worse performance of the WBAN. From point A to point C, the LOS component decreases, more and more radio propagation relies on the human body. As a result, more and more performance degradation is observed in the whole WBAN system. The indoor propagation channel further degrades the BER comparing with result obtained for anechoic chamber.

It is notable that BER at point C is very poor even in the anechoic chamber where the indoor multipath is eliminated. Clearly, this is resultant from the human-body effects. At point C, the rms delay spread is approaching 10 ns. However, 6 ns window is selected as the monocycle duration. There is strong inter-symbol interference that results in destructive BER degradation.

Review of these results shows that PPM is more sensitive to the rms delay spread than PSK and OOK. It can be explained from its time-shift nature, which needs a longer chip time and thus higher probability of inter-symbol interference. PSK is a slightly better choice than PPM in most cases, especially when higher rms delay spread presents (e.g. at point B). The energy detection method of OOK is less sensitive to the channel environment, and can be the most power-efficient way under very high rms delay spread (e.g. at point C). But OOK is less superior when the LOS components dominate the channel.

The efficiency of modulation scheme also depends on the pulse shape. It seems that the result is case by case, there is no any pulse shape in certain modulation that can always perform best in different channel environments. However, if OOK is employed, rectangular pulse should be selected for its advantages in hardware implementation.

Chapter 5

Inductorless UWB Low-Noise Amplifier

5.1 Introduction

Driven by the rapid development of UWB technology, ultra-wideband low-noise amplifiers have received extensive research interests in recent years. The 3.1–10.6-GHz full band operation poses stringent requirements for the front-end LNA [21, 28], which has to provide an ultra-wide bandwidth with reasonable noise figure (NF) and impedance matching. Minimizing supply voltage and power consumption is always demanded for portable applications.

Traditionally, this kind of wideband amplifiers were implemented with balanced or distributed architectures that originally used in microwave circuit design [95, 96, 97, 98]. However, the large area occupation and high power dissipation of traveling-wave amplifier make it infeasible for low-power single-chip integration. Recently, a distributed amplifier is reported in [99], which

achieves comparable performance with lumped design in terms of power and area consumption.

With the development of advanced semiconductor technology, lumped implementation of LNA in CMOS and SiGe BiCMOS has been pushed up to tens of GHz [3, 100]. There are also a few of UWB LNAs that have been reported in literature. Various circuit techniques have been proposed to enhance the bandwidth.

A classical approach to widening bandwidth is negative feedback, which is normally realized in the form of resistive shunt feedback [101, 102, 103, 104]. A LNA with active feedback technique employing emitter-follower in the shunt path is present in [105]. Meanwhile, a stand-alone negative feedback can hardly achieve sufficient bandwidth, inductor peaking technique is often adopted in these feedback based designs. Further, the analysis presented in [106] shows that resistive feedback amplifiers can not provide required performance with low power consumption. In [106, 107, 108], UWB LNAs with multi-section impedance matching are reported, which expand the use of inductor-degenerated amplifier [109] with emphasis on the wideband impedance matching. This approach is efficient in terms of bandwidth and has been proved in both CMOS and SiGe BiCMOS technologies. However, the extra passive devices used for matching purpose increase design complexity and area occupation. Recently, a combined feedback architecture for UWB LNA design is reported [97, 110, 111]. Adopted from microwave circuits [112, 113], the resistive feedback is applied both globally and locally. This architecture achieves sufficient bandwidth and gain even without inductor. However, potential instability problem occurs due to the multiple feedback loops. High transconductance device is also required for bandwidth-gain tradeoff. Thus, this configuration has

not been realized in standard CMOS technology.

Comparing with narrowband LNA designs, severe tradeoff between noise figure and source impedance matching exists in wideband LNA. Most of reported UWB LNA designs are focused on bandwidth enhancement. As a result, few of them achieve comparable noise performance with narrowband LNAs. A CMOS UWB LNA employing noise-canceling technique is reported in [114], where thermal noise of input matching device can be sensed and canceled by the feedforward configuration [115]. This avoids the potential instability due to global negative feedbacks. However, the gain performance of such configuration is often less superior.

This chapter presents a UWB LNA with comprehensive considerations on noise figure and wideband gain performance. A modified noise sensing and canceling architecture is exploited, which employs a current-reuse stage for gain compensation. Inductorless design is also explored. The bandwidth enhancement is achieved by capacitive peaking.

Section 5.2 reviews the wideband noise canceling techniques. Section 5.3 presents a modified wideband noise canceling architecture capable of higher gain performance, the properties and considerations are discussed in detail. In Section 5.4, approaches to bandwidth enhancement are discussed and an inductorless approach using capacitive peaking is presented. Section 5.5 shows the experimental results of the demonstrated LNA circuit.

5.2 Review of Wideband Noise Canceling

The wideband LNA design exploiting thermal noise canceling was originally presented by F. Bruccoleri *et al.* [115, 116]. The idea is briefly introduced as follows. Consider the LNA circuit shown in Figure 5.1(a); the wideband input impedance matching is realized by a common-source stage with resistive shunt feedback. The noise current of $M1$ (and $M2$) flows out of the node B through feedback resistor R_f and source impedance to the ground, leading to two equal-signed noise voltages at node A and B . They are subsequently amplified by a combining stage, where opposite-signed noise voltages are produced by the source-follower (see from B) and common-source amplifier (see from A). By carefully designing the circuit parameters, the noise contribution of matching device $M1$ (and $M2$) can be canceled. The cancellation condition is derived as

$$A_{v,2} = 1 + \frac{R_f}{R_s}, \quad (5.1)$$

where R_s is the source impedance. $A_{v,2}$ denotes the gain of the output common-source stage ($M3$ – $M4$). This equation is determined by the noise sampling mechanism of the input stage. Note that $M2$ is not necessarily needed by the noise canceling principle; it increases the effective g_m of input matching stage where $Z_{IN} \approx 1/g_{mi} = 1/(g_{m1} + g_{m2})$. On the other hand, the signal voltages at node A and B have opposite sign as long as $g_{mi}R_f > 1$, thus, the combining stage produces equal-signed output signal voltages and adds them up. The noise contribution of R_f and combining stage can not be canceled in this configuration, they dominate the noise performance of overall LNA. An inductorless LNA is demonstrated in [115], which achieves 13.7-dB voltage gain, 0.002–1.6-GHz –3-dB bandwidth, and less than 2.4-dB NF in 0.25- μm CMOS.

Figure 5.1: Simplified wideband LNA architectures exploiting thermal noise canceling with (a) resistive shunt-feedback and (b) common-gate input.

An alternative noise canceling configuration, as shown in Figure 5.1(b), is also proposed in [115] and has been successfully implemented for UWB in [114]. In this configuration, a common-gate stage (M1) is employed for wideband input matching. Designed in $0.18\text{-}\mu\text{m}$ CMOS, 9.7-dB gain, 1.2–11.9-GHz -3-dB bandwidth, and 4.5–5.1-dB NF (over 3.1–10.6-GHz) are obtained in [114].

Review of above results shows that there is less freedom in controlling the gain performance in noise canceling architectures, especially in the GHz range. For the matching stage using resistive shunt feedback (Figure 5.1(a)), although the noise restraints are relaxed, the input matching requirement limits the voltage gain A_v of such stages [107]. A common-gate input stage (Figure 5.1(b)) can not provide high gain as well. For the output combining stage, the gain is constrained by the noise canceling principle, where the value of gain in two paths should produce two equal-amplitude noise voltages so that they can be canceled [114, 115]. As a result, both designs encounter architectural diffi-

culties in achieving higher gain. Comparing to SiGe BiCMOS where higher g_m is available, the gain issue becomes more severe when it applies to CMOS. In fact, the design in [114] has incorporated an additional common-source output stage (M4) to enhance the gain to 9.7-dB. However, this stage contributes extra noise to the output. This can not be canceled, resulting a well above 3-dB overall NF. In this case, there is little noise cancellation observed.

5.3 Gain-Enhanced Wideband Noise Canceling Architecture

A gain-enhanced noise canceling architecture is presented in this section. The purpose of gain enhancement is not only on improving the gain of noise canceling architectures, but also on creating more freedom in tradeoffs involving gain. Since the inductorless approach will be explored in this chapter, parasitic capacitances can not be driven with inductor-peaking method. To achieve ultra-wideband, extensive tradeoff between bandwidth and gain may be required. Thus, a higher gain configuration offers more flexibilities for such tradeoff.

5.3.1 Architecture and Analysis

Figure 5.2 shows the modified architecture of wideband noise-canceling LNA. A current-reused amplification stage including $M2-R2$, $M3-R3$ and $C2$ is adopted between the noise sensing and combining stages. $M2-R2$ and $M3-R3$ are stacked gain blocks within a same DC current branch. $C2$ is a large decoupling capacitor that creates a ground node at RF. This stage provides two feedforward

Figure 5.2: Architecture of gain-enhanced wideband LNA with thermal noise canceling.

paths concurrently. Meanwhile, the current reuse permits higher efficiency of power and available voltage headroom [117]. $R8$ is employed to provide extra current to the gain stack $M2$ – $R2$; this will be discussed later. Note that the input parasitic capacitance degrades both input matching and noise sensing [107, 115]. As the target frequency of the design is up to 10-GHz, it is important to keep the input parasitics as small as possible. Thus, the input stage in Figure 5.1(a) is replaced by a simple NMOS common-source stage. As a result, this stage would consume slightly more current for impedance matching.

Intuitively, the gain ratio of the two feedforward paths can be kept unchanged if the parameters of $M2$ – $R2$ and $M3$ – $R3$ are given the same values (and without $R8$). Therefore, the noise canceling condition presented in [115] is retained, with improved absolute gain values available. However, the gain-enhancement stage contributes additional noise as well. Using same parameters

for both $M2-R2$ and $M3-R3$ is not always the optimal choice in terms of overall noise performance. This can be analyzed as follows.

At node C , the noise voltage due to $M3$ and $R3$ can be expressed as

$$\overline{v_{n,c}^2} = 4kT \cdot \left(\text{NEF} \cdot g_{m3} + \frac{1}{R_3} \right) R_3^2 \cdot \Delta f, \quad (5.2)$$

where $\text{NEF} = \gamma \cdot g_{ds0}/g_m$ is the noise excess factor. g_{ds0} is the channel conductance when $V_{DS} = 0$. For submicron MOSFETs, NEF is well above 1 since $g_{ds0}/g_m > 1$ and $\gamma > 1$. Δf is the calculation bandwidth. Note that in (5.2), flicker noise is not considered due to the high-frequency operation.

Practically, we have $\text{NEF} \cdot g_{m3} \gg 1/R_3$. Also we consider the noise per unit bandwidth. Thus, (5.2) can be simplified as

$$\overline{V_{n,c}^2} = 4kT \cdot \text{NEF} \cdot g_{m3} R_3^2. \quad (5.3)$$

Similarly, at node D , the noise voltage due to $M2$ and $R2$ can be expressed as

$$\overline{V_{n,d}^2} = 4kT \cdot \text{NEF} \cdot g_{m2} R_2^2. \quad (5.4)$$

It is clear that the two noise voltages are uncorrelated. They could not be canceled by subsequent combining stage. Refer to the output node, the noise due to gain-enhancement stage is

$$\overline{V_{n,out}^2} = 4kT \cdot \text{NEF} \cdot \left(g_{m2} R_2^2 \frac{g_{m4}^2}{g_{m5}^2} + g_{m3} R_3^2 \right). \quad (5.5)$$

Equation (5.5) shows the noise contribution of $M2-R2$ is of higher weight than that of $M3-R3$.

5.3 Gain-Enhanced Wideband Noise Canceling Architecture

77

The output noise of matching device $M1$ can be canceled by properly designing of the gain in the two feedforward paths. In Figure 5.2, the noise voltage sampled at node A is amplified by $M2$ and $M4$; while the noise voltage at node B is amplified by $M3$ and $M5$. The cancellation condition can be written as

$$R_s \cdot g_{m2} R_2 \cdot \frac{g_{m4}}{g_{m5}} = (R_f + R_s) \cdot g_{m3} R_3. \quad (5.6)$$

Therefore, (5.5) becomes

$$\overline{V_{n,out}^2} = \alpha \cdot (g_{m3} R_3)^2 \left[\frac{\left(1 + \frac{R_f}{R_s}\right)^2}{g_{m2}} + \frac{1}{g_{m3}} \right], \quad (5.7)$$

where $\alpha = 4kT \cdot \text{NEF}$. From (5.7), the noise contribution of gain-enhancement stage can be minimized when

$$\frac{g_{m2}}{g_{m3}} = \left(1 + \frac{R_f}{R_s}\right)^2. \quad (5.8)$$

And the minimum output noise voltage is

$$\overline{V_{n,out}^2} = 2 \cdot \alpha \cdot g_{m3} R_3^2. \quad (5.9)$$

It is interesting to note that these results are independent of the following noise combining stage. Note that (5.7)–(5.9) are obtained under the noise cancellation condition for the input matching device. From (5.8), the condition (5.6) can be simplified as

$$\frac{g_{m5}}{g_{m4}} = \frac{R_2}{R_3} \left(1 + \frac{R_f}{R_s}\right). \quad (5.10)$$

Consider the overall small-signal gain, we have

$$\begin{aligned} A_V &= -A_{v,1} \cdot g_{m3}R_3 + g_{m2}R_2 \cdot \frac{g_{m4}}{g_{m5}} \\ &= \left(-A_{v,1} + 1 + \frac{R_f}{R_s} \right) \cdot g_{m3}R_3, \end{aligned} \quad (5.11)$$

where $A_{v,1}$ is the gain of input matching stage,

$$A_{v,1} = -\frac{(g_{m1}R_f - 1)R_1}{R_f + R_1}. \quad (5.12)$$

The result is intuitive, a gain-enhancement factor $g_{m3}R_3$ is achieved comparing with the original architecture (Figure 5.1(a)) whose gain is $-A_{v,1} + 1 + R_f/R_s$.

5.3.2 Properties and Limitations

Design Flexibility

The additional parameters in the noise cancellation equations (5.8)–(5.11) permit more design flexibilities. Generally, the original noise cancellation technique is a two-step operation: noise sensing and combining/canceling. Besides the system requirements, the design parameters of the two stages are further limited by each other (to fulfill the noise cancellation condition). The proposed technique provides a three-step operation: noise sensing, amplification, and combining. The noise cancellation requirement can be fulfilled with more design flexibilities. From this point of view, the amplification stage acts as a “buffer” which reduces the coupling between the noise sensing and combining stages.

A good example is the design of the output combining stage, which dom-

inates the noise performance of the original architecture. In Figure 5.1(a), $g_{m3}/g_{m4} = A_{v,2} = (1 + R_f/R_s)$ should be fulfilled. This poses a rather restrained headroom in minimizing the output noise due to combining stage. However, from (5.6) or (5.10), the additional parameters in the proposed architecture (Figure 5.2), e.g. R_2 and R_3 , provide a more flexible requirement of g_{m4} and g_{m5} . This can be used to minimize the noise of combining stage. As a result, the additional noise contributed by the gain-enhancement stage can be partially compensated by the improved combining stage.

The design flexibilities offered in the proposed circuit can also be understood from the point of view of tradeoff. The fundamental tradeoff between noise figure and source impedance matching exists in wideband LNA. To relax this tradeoff, the original noise cancellation technique introduces one more parameter of gain. The simultaneous noise and matching performance is achieved by the tradeoff with gain through the noise cancellation equations. By inserting the amplification stage as proposed here, the tradeoff among noise, matching and gain is relaxed, while the cost is the increased noise and distortion contributed by the additional amplification stage.

From (5.9) and (5.11), the amplification stage determines the tradeoff between the gain enhancement and the additional noise contribution. Although the gain can be improved by increasing $g_{m3}R_3$, the noise figure of overall LNA also increases, and vice versa. To maintain a reasonable noise figure, the enhanced gain can not be arbitrarily high. The optimal design of circuit parameters depends on the target specification.

Distortion

As the additional amplification stage does not affect the noise sensing mechanism, the canceling property and robustness of the original architecture is maintained. Any small signal that can be modeled as a current source along the channel of matching device is canceled as well. To the first order, the distortion of the matching device is also canceled [115], and this property is kept in the proposed design.

Inevitably, the gain-enhancement stage contributes extra distortion and thus degrades the linearity of the overall LNA. To have a easy comparison with [115], the similar assumption is made here: the distortion originates only from the nonlinear memoryless voltage to current conversion of the transistors. In Figure 5.2, the nonlinear voltage generated by the additional amplification stage can be written as

$$\begin{aligned} V_C &= (g_{m3}V_B + I_{NL3}) R_3, \\ V_D &= (g_{m2}V_A + I_{NL2}) R_2, \end{aligned} \quad (5.13)$$

where I_{NL2} and I_{NL3} denote the weakly nonlinear current generated by $M2$ and $M3$, which are also uncorrelated. The expressions of these terms normally employ Volterra series. Therefore, it is difficult to estimate the effect of the additional distortion. In fact, the distortion originates from various mechanisms and more elements. Considering the high-frequency effects, the distortion analysis is only able to be performed to a limited extent [118, 119]. In this chapter, the distortion is evaluated by its linearity performance, which will be discussed in Section 5.5.

Mismatch

Another issue introduced by the gain-enhancement stage is mismatch. In principle, the noise canceling architecture employs two feedforward paths, device mismatch between them will degrade the noise canceling. When the amplification stage is not used, it is shown that the two-step noise canceling is robust to device parameter variations [115]. With the amplification stage in Figure 5.2, the excess noise factor contributed by the matching device can be expressed as

$$EF_{MD} = NEF \frac{[(R_f + R_s) A_{v,3} - R_s A_{v,2} A_{v,4}]^2}{R_s A_V^2}, \quad (5.14)$$

where $A_{v,2} = g_{m2}R_2$, $A_{v,3} = g_{m3}R_3$ denote the gain of the two paths, $A_{v,4} = g_{m4}/g_{m5}$ denotes the gain of noise combining stage. A_V is the overall gain as discussed in (5.11).

Note that (5.14) is obtained by assuming equal NEF and $Z_{IN} = R_s$. Since the nominal values of the parameters in (5.14) are determined by (5.5)–(5.12), mathematical expressions of the mismatch effect may not be intuitional. Here the numerical simulation is performed assuming $NEF=1.5$ and $R_f/R_s = 3$. Figure 5.3 shows the noise contribution of the matching device (EF_{MD}) resulting from the variations of $A_{v,2}$ and $A_{v,3}$. It is shown that EF_{MD} is kept at 0 as long as the variations of $A_{v,2}$ and $A_{v,3}$ are equal to each other. That is, it is not the absolute variation but the relative variation that degrades the noise canceling. In other words, the mismatch between the two amplification paths degrades noise canceling. With regard to EF_{MD} margin of 0.1, $\pm 6\%$ mismatch between $A_{v,2}$ and $A_{v,3}$ is allowed.

The result obtained in [115] shows that $\pm 20\%$ variations in the noise sens-

Figure 5.3: Numerical simulation of EF_{MD} with regard to the variations of $A_{v,2}$ and $A_{v,3}$.

ing and combining stages are allowed for EF_{MD} margin of 0.1. However, the additional gain stage is relatively sensitive to device parameter mismatch. Therefore, the mismatch robustness of the overall LNA is limited by this stage, which is another price that is paid for the increased gain and design flexibilities.

5.3.3 Circuit Design

Based on the design equations (5.5)–(5.12), the circuit parameters of Figure 5.2 can be optimized. A standard 1.2-V 0.13- μm CMOS process is used in this design. The target specifications are: 1) 20-dB voltage gain; 2) < 3 -dB noise figure; 3) input impedance matching to 50Ω over 3.1–10.6-GHz. The -3 -dB bandwidth of voltage gain is also of high importance. For this architecture, -3 -

dB bandwidth is mainly decided by parasitic capacitances. In this subsection, minimizing parasitics will be taken into considerations, but no bandwidth target is set as the current design phase is dedicated to gain and noise performances.

R_f can not be very large although large R_f offers high gain. From (5.8), a large R_f/R_s results in a huge ratio of g_{m2} and g_{m3} . Considering that $M2$ and $M3$ share a same current branch, it is quite difficult to obtain a huge g_{m2}/g_{m3} . Secondly, the input matching requires a small R_f as compared to the input parasitics $C_{in} = C_{gs1} + C_{gs2}$ at several GHz frequencies. As a result, the gain of input stage, $A_{v,1}$, is a small value. Furthermore, C_{in} should be kept small for precise noise sensing, otherwise noise cancellation degrades. Thus, large $M1$ and $M2$ should be avoided. High g_{m1} and g_{m2} have to be obtained by increasing the respective drain current. Here $R_f/R_s = 3$ and $A_{v,1} \approx 0.97$. The input stage draws 3-mA with 1.2-V supply voltage.

Equation (5.8) then becomes $g_{m2}/g_{m3} = 16$. It is still a large ratio in practical circuit. On the other hand, g_{m2} and g_{m3} can not be selected according only to this condition. A large g_{m2} requires high current and large size of $M2$, which is not favorable for the sake of power and C_{in} , respectively; while a low g_{m3} degrades the overall gain. In fact, (5.8) considers only the minimum noise contribution of gain-enhancement stage. Thus, there are tradeoffs in the design of g_{m2} and g_{m3} . Simulation shows a reasonable ratio $g_{m2}/g_{m3} = 50\text{mS}/10\text{mS} = 5$ provides most balanced performances. To maintain a minimum C_{in} , large g_{m2} can not be realized only by increasing the size of $M2$. Thus, $R8$ is added in parallel with $M3$ – $R3$ to provide more current to $M2$. Meanwhile, as $M3$ draws only part of the drain current of $M2$, a larger $R3$ within the same voltage headroom is available, leading to a larger available gain-enhancement factor. Note that $R8$'s noise contribution is little due to the large $C2$.

Figure 5.4: Simulated performance of the gain-enhanced noise canceling circuit in Figure 5.2.

The parameters of combining stage depends on (5.6). In ideal case, the output impedance matching could be achieved [115]. However, the stringent noise budget permits little freedom in doing so. In this design, the output is not intentionally matched, which would certainly affect the overall power gain.

Figure 5.4 shows the simulated performance of a gain-enhanced noise canceling LNA. Note that 2-pF capacitors are used for input and output coupling. 2.4–2.8-dB noise figure, 15.6-dB maximum gain, and $S_{11} < -10$ -dB input matching are achieved. The whole circuit draws around 10.8-mA with 1.2-V supply.

5.4 Capacitive Peaking

On-chip inductor is widely used to improve the bandwidth performance of broadband LNA. For the proposed circuit in Figure 5.2, inductors can also be

Figure 5.5: Example of a gain-enhanced noise canceling LNA with inductor peaking.

used for bandwidth extension. A possible example is shown in Figure 5.5, where L_1 and L_2 are employed in the input stage. Figure 5.6 shows the simulation result with $L_1=0.5\text{nH}$, $L_2=1.7\text{nH}$. Comparing with Figure 5.4, the bandwidth is increased and very good gain flatness is obtained. Furthermore, the input matching and noise performance are improved significantly. This is because the input capacitance is the dominate limitation for both impedance matching and noise sensing, the inductors in the input stage decrease such effect at high frequencies.

In [114], the design employs an extra stage after the noise canceling LNA to improve the gain, where 5 inductors are used and the simulated noise figure is well above 3.5-dB. The comparison shows the proposed gain-enhancement technique permits better noise performance with less inductors. Meanwhile, the inductors used here can possibly be implemented with bond wire inductance

Figure 5.6: Simulation result of Figure 5.5 with $L_1=0.6\text{nH}$, $L_2=1.5\text{nH}$.

[53].

However, the on-chip inductor occupies large silicon area. For applications where the area requirement is stringent, e.g. multiple-input multiple-output (MIMO) transceiver, the inductor based design is not favorable. While the use of bond wire inductance depends on the circuit configuration and is severely limited by the packaging and modeling technologies. Hence, in this chapter, the capacitive peaking technique is employed for bandwidth extension, which has been successfully implemented in transimpedance amplifiers [120].

Figure 5.7 shows the schematic of a capacitive peaking common-source stage. It is adopted from the source degeneration architecture. The gain of this stage can be written as

$$A_v = -\frac{R_L}{\frac{1}{g_m} + R_p \parallel C_p} = -\frac{R_L}{\frac{1}{g_m} + \frac{1}{\frac{1}{R_p} + j\omega C_p}}. \quad (5.15)$$

Figure 5.7: Schematic of a capacitive peaking gain stage.

For the first order, C_p provides an increasing A_v regarding frequency. This effect can be understood conceptually from the resistive degeneration architecture. It is well known that a source-degenerative resistor degrades gain and improves linearity. While C_p provides a signal path in parallel with R_p to the ground. At high frequencies, the path through C_p has a low impedance. Thus, the resistive degeneration effect is degraded by C_p with the increase of frequency, resulting in an increasing gain characteristic. However, the gain can not exceed the value of a common-source stage without source degeneration, which actually sets the upper bound of the gain of capacitive peaking stage. In other words, capacitive peaking is rather an approach to controlling the gain flatness at the cost of absolute gain value. Note that a negative impedance is created, therefore, the stability of the capacitive peaking LNA should be carefully examined.

Besides gain degradation, the degeneration resistor and capacitor also affect the impedance matching and noise-canceling conditions. Thus, capacitive peaking should be carefully verified in the circuit. For the gain-enhanced noise canceling architecture in Figure 5.2, the input transistor $M1$ is of the highest impact on the noise and matching performance, thus, source degeneration is not used in $M1$. The gain of $M3$ – $R3$ stage determines the gain-enhancement

Figure 5.8: Complete schematic of gain-enhanced noise canceling LNA with capacitive peaking.

factor directly; it is then efficient to apply capacitive peaking technique in $M3$ – $R3$. $M2$ – $R2$ and $M4$ – $M5$ also adopt the capacitive degeneration.

The use of source degeneration depletes the available voltage headroom, especially for the gain-enhancement stage which actually contains two stacked common-source amplifiers with source degeneration. Although the presence of R_8 redistributes the current flow in the branch, a 1.2-V supply can not safely provide sufficient voltage headroom for M_2 and M_3 . Hence, 1.5-V supply voltage is used. Meanwhile, a tail resistor R_5 is added to control the DC current of M_4 and M_5 .

Figure 5.9: Simulation result of the proposed noise canceling LNA with gain-enhancement and capacitive peaking.

Figure 5.8 shows the final circuit of the proposed LNA. The simulation result is shown in Figure 5.9. With 1.5-V supply voltage, the circuit parameters are redesigned to preserve a similar condition as presented in Section 5.3.3. At the price of degraded gain of 12-dB and minimum NF of 2.8-dB, simulation shows the capacitive peaking technique enhances gain flatness within 3-dB over 2.6–12-GHz. The stability is carefully verified, which shows the LNA is unconditionally stable.

5.5 Measurement Results and Discussions

The LNA circuit was fabricated in a two-poly eight-metal 0.13- μm triple-well CMOS technology. The cut-off frequency f_T of NMOS transistor is over 90-

Figure 5.10: Die microphotograph of the fabricated LNA chip.

GHz. The triple well process employs buried deep *N*-well in the *P*-substrate to form an isolated *P*-well, which reduces the noise coupling through substrate. Figure 5.10 shows the die microphotograph of the fabricated LNA. As the passive devices used are only resistors and capacitors, the LNA chip with test pads occupies only $415\mu\text{m} \times 415\mu\text{m}$ area, which is among the smallest designs.

The measurement of the LNA chip is performed on wafer using Cascade G-S-G RF probes. Figure 5.11 shows the measured performance of the fabricated LNA. The maximum gain is 11-dB and the -3 -dB bandwidth is 2–9.6-GHz. The frequency dependent difference comparing to simulated power gain is most likely due to the extra parasitic capacitances in the real chip. At the same time, the gain degradation may also result from the device parameter variations. From Figure 5.8, the whole circuit is self-biased; the current in each branch is directly determined by the supply voltage and resistor accuracy. The

Figure 5.11: Measured performance of the fabricated LNA.

latter is process dependent and can not be controlled manually. Measurement shows the LNA draws totally 12.65-mA instead of the simulated 12-mA, which is a clear evidence of the existence of inaccurate resistors. Using mirrored current sources can improve the current accuracy; however, the transistors are not comparable with resistors in terms of wideband performance.

The S_{11} is less than -8.3 -dB over 2–9.6-GHz, which shows the resistive shunt feedback can effectively offer a 50Ω matching resistance over the UWB frequencies. The reverse transmission gain S_{12} is below -50 -dB (not shown here).

The noise figure is within 3.6–4.8-dB over 2–9.6-GHz, which is around 1-dB higher than the simulated result. As discussed previously, this is probably due to the mismatch between the two gain-enhancement paths. The inaccuracy of noise model can also create discrepancy. The average noise figure over 2–9.6-GHz is 4-dB.

Figure 5.12: The input 1-dB compression point (ICP) and input third-order intercept point (IIP3) at different frequencies.

The linearity is evaluated by both input 1-dB compression point (ICP) and input third-order intercept point (IIP3). During the two-tone test, 100-MHz tone spacing is used. Figure 5.12 shows the test result at different frequencies over 3–11-GHz. It is shown that the ICP is within $[-18.8, -15.2]$ -dBm and the IIP3 is within $[-9.1, -6.5]$ -dBm. At 6-GHz, the ICP is -16.5 -dBm and the IIP3 is -7.2 -dBm. Both ICP and IIP3 increase slightly at higher frequencies, which may result from the increasing capacitive loss and decreasing power gain at higher frequencies. Similar observation can be found in [106]. Furthermore, the UWB system also suffers from the second-order intermodulation (IM2) interferences [48]. A further two-tone test is performed with $f_1 = 6$ GHz and $f_2 = 2.4$ GHz, since their second-order intermodulation product falls in the signal band. The input second-order intercept point (IIP2) in this case is 21-dBm. The linearity of this LNA is comparable to the reported inductor based designs.

Note that since the output is not intentionally matched to 50Ω , and there is no additional buffer stage employed, the power gain and noise figure performances achieved here have been degraded by the 50Ω termination of test equipment. The maximum voltage gain calculated in this design is around 14-dB, which is relatively low since the capacitive peaking degrades the gain severely. When the area requirement is not stringent, the inductor based approach as shown in Figure 5.5 is much more preferred, the bandwidth and noise performance should be improved significantly as compared to this design.

Table 5.1 shows a summary of measured performances of the inductorless LNA. The comparison with other reported state-of-the-art amplifier designs is also presented. It is shown that LNA designed without inductor achieves comparable performances with inductor-based LNA designs over the ultra-wide bandwidth. The presented design is suitable for UWB wireless receiver front-ends.

Table 5.1: Summary of the LNA Performances and Comparison with Reported State-of-the-Art Designs

	Techology	Gain (dB)	BW (GHz) ¹	NF (dB)	S ₁₁ (dB)	ICP (dBm)	IIP3 (dBm)	Area (mm ²)	Power (mW)
This Work	0.13-μm CMOS	11	2-9.6	3.6-4.8	< -8.3	-16.5 *	-7.2 *	0.17	19
[107] ²	0.18-μm CMOS	10.4	2.4-9.5	4.2-8	< -9.4	-18 *	-8.8 *	1.1	9
[106]	0.18-μm SiGe	21	3-10	2.5-4.2	< -10	-11.8 †	-5.5 †	1.8	30
[115]	0.25-μm CMOS	13.7	0.002-1.6	1.9-2.2	< -8	-9 †	0 †	0.075	35
[114]	0.18-μm CMOS	9.7	1.2-11.9	4.5-5.1	< -14	-16 *	-6.2 *	0.59	20
[103]	0.13-μm CMOS	16	2-5.9	4.7-5.7	< -9	-8 ³	—	0.24	38 ⁴
[102]	0.18-μm CMOS	9.8	2-4.6	2.3-4	< -14	—	-7 ⁴	0.9	12.6
[95]	0.18-μm CMOS	10.6	0.5-14	3.4-5.4	< -11	+10	+20	1.6	52
[105]	0.5-μm SiGe	12	0-15	2.8-4	< -8	-7.6 [▷]	+1.9 [▷]	0.25	24
[112]	InGaP-GaAs HBT	16	0-11.6	—	< -10	—	—	0.23	96
* at 6-GHz † at 3.4-GHz ‡ at 0.9-GHz ◁ at 4-GHz ▷ at 2-GHz									

¹ -3-dB bandwidth.

² Performance of the LNA designed in triple-well process.

³ Output 1-dB compression point.

⁴ Total power including output buffer.

Chapter 6

Differential T/R Switch

6.1 Introduction

With the development of modern silicon technology, more and more high-frequency circuits can be implemented in standard CMOS process. Radio-frequency integrated circuits in standard CMOS technology have proven feasible [43]. The trend of SoC/SiP requires further integration of transmit/receive antenna switch.

For years, RF switch has been dominated by discrete components using PIN diodes and III-V MESFETs. Recently, CMOS T/R switch design has been explored to a certain extent. Regarding the performance of insertion loss and isolation, the effect of substrate resistance is studied in [121, 122], where low insertion loss was obtained by minimizing the substrate resistance and DC biasing the T/R nodes. A high isolation was achieved using CMOS-SOI technology [123]. In both cases, however, the linearity was limited due to the parasitic capacitance and source/drain junction diodes. Thus, techniques of body float-

ing are developed for higher linearity. A LC-tuned substrate bias technique is firstly reported in [124], where the bulk is not separated from substrate. Using on-chip inductor can tune the bulk of switching transistor to be floating at certain frequencies. At 5.2-GHz, 28-dBm 1-dB compression point (P_{1dB}) in the transmit mode was obtained. The disadvantages of this approach are the design complexity and large silicon area consumed. Taking advantage of modern triple-well CMOS process, the idea of body floating can be simply realized by using a large resistor to bias the bulk [125]. As resistors are intrinsically wideband, the linearity improvement of this approach is also wideband. 20-dBm P_{1dB} was achieved at 5.8-GHz. Another approach to linearity improvement is using stacked transistors [126], however, insertion loss will be degraded and has to be compensated, e.g. using the special DET process. A 15-GHz T/R switch is reported in [127]; the impedance transformation network was employed to improve the linearity, while the isolation performance is degraded.

Differential T/R switch designed in commercial CMOS technology has not been reported to date. Comparing with single-ended architecture, the differential nature permits higher linearity, lower offset, makes it immune to power supply variations and substrate noise. Therefore, differential architecture is normally preferred in applications requiring higher signal quality. Exploring the design of integrated differential T/R switch is essential for transceiver front-ends with fully differential architecture.

In this chapter, the preliminary design and implementation of differential T/R switches are explored. The design is targeting on the 3.1–6-GHz lower-band UWB applications. Design issues on the transistor sizing, crosscoupling and matching are considered. A fully-differential T/R switch is demonstrated in a 0.18- μm standard CMOS technology. Measurement results exhibit less than 2-

dB insertion loss, higher than 15-dB isolation. 15-dBm P_{1dB} is achieved without using additional techniques to enhance the linearity. The silicon area occupied is only $60\mu\text{m} \times 40\mu\text{m}$. These results show that the proposed switch is suitable for multi-standard and broadband wireless front-ends with moderate power level.

6.2 Architecture and Considerations

6.2.1 Architecture

The schematic of the proposed differential T/R switch is shown in Figure 6.1. Transistors M1, M2, M3 and M4 perform the main switching function. A high control voltage V_{ctrl} turns M1 and M3 on, which enables the differential path between antenna and receiver. Similarly, the differential transmit path is turned on when the control voltage is low. The control voltage is biased through a resistance R_G to reduce the effect due to capacitive coupling around the gate of the OFF transistors [123].

The differential nature results in an improved power handling capability comparing with single-ended configurations. From the power point of view, differential output scheme is able to handle twice over the single-ended output power, that is, 3-dB higher P_{1dB} could be achieved in the proposed differential switch. As the power handling capability is the bottleneck of CMOS T/R switches, differential architecture is of great advantage in current silicon technology.

Figure 6.1: Architecture of the proposed differential T/R switch.

6.2.2 Design Considerations

Theoretically, transistor sizing in differential T/R switches is the same with that in single-ended switches. In the case where the MOS transistor is turned *ON*, the equivalent circuit is shown in Figure 6.2 [121]. R_{ON} is the on-resistance of the transistor which is operating in the linear region. At low frequencies, the insertion loss is determined by R_{ON} . R_B is the substrate resistance. C_{DB} , C_{SB} , C_{GB} , C_{GS} and C_{GD} are parasitic capacitances of the transistor. The parameters are eventually frequency dependent. The capacitive coupling effect increases with the increase of operating frequency, resulting in the increase of the power loss on the substrate resistance R_B .

The insertion loss can be calculated as

$$IL = \frac{\text{Power Available from Source}}{\text{Power Delivered to Load}} = \frac{P_{AVS}}{P_L}$$

Figure 6.2: Small signal equivalent circuit of a MOS transistor switch.

$$\begin{aligned}
 &= \frac{1}{|S_{21}|^2} \\
 &= \frac{\omega^2 C_T^2 [(R_{ON} + 2Z_0)R_B + (R_{ON} + 2Z_0)Z_0]^2}{(2Z_0)^2 (1 + \omega^2 C_T^2 R_B^2)} \\
 &\quad + \frac{(R_{ON} + 2Z_0)^2}{(2Z_0)^2 (1 + \omega^2 C_T^2 R_B^2)}, \tag{6.1}
 \end{aligned}$$

where Z_0 is the characteristic impedance the source and load are terminated with. ω is the operating radian frequency. The total effective parasitic capacitance C_T makes the insertion loss frequency dependent. The first term in (6.1) increases as frequency increases, which can be minimized by optimizing the values of R_B and R_{ON} . Previous research shows that low substrate resistance is preferred in the high frequency switch design [122]. The second term in (6.1) is actually the intrinsic insertion loss in DC due to the on-resistance R_{ON} of the transistor, which decreases as frequency increases. This term can also be min-

imized by minimizing R_{ON} . In the linear/triode region, R_{ON} can be expressed as [128]

$$R_{ON} = \frac{1}{\mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{th})}. \quad (6.2)$$

Thus, the second term in (6.1) can be minimized by increasing the W/L ratio of the switch transistors, where tradeoff occurs since a large transistor leads to a large parasitics, resulting in a large C_T and thus frequency dependent first term in (6.1) will be increased. An optimal W/L ratio exists to minimize the insertion loss.

When the MOS transistor is turned *OFF*, the undesired signal couples from drain-source path and substrate parasitics. The isolation of a T/R switch severely depends on technology [123]. In a bulk CMOS process, the isolation is determined by the parasitics around drain and source [129], increasing of the gate width leads to the degradation of isolation performance. Previous research [129] and simulation show the trend is monotonous, transistor size can be optimized to provide the minimum insertion loss while maintaining a reasonable isolation.

The linearity, or power handling capability, is directly related to the bias condition of the MOS transistors. A large signal at the drain/source may cause the junction diodes forward biased and clip the signal. The unintentional turn on of the *OFF* transistors can also distort the signal. As a result, the DC bias of TX/RX nodes affects the linearity significantly [121, 122, 123]. High DC bias and high control voltages are often used in the CMOS T/R switch for better linearity. However, the reliability problems may exist potentially with a large gate-source voltage. In this design, the differential architecture is employed that improves the linearity fundamentally rather than overstress the CMOS transistors, the

control voltage is provided by the build-in inverter with 1.8-V standard supply voltage.

Shunt transistors are commonly used to improve the isolation [121, 122, 123, 129], which provides a low-impedance path for the undesired signal to the RF ground. However, the additional transistors increase the possibility of unintentional turning on and thus degrade the linearity. Considering the stringent voltage limitation that gives little space to safely bias the shunt transistors, they are not used in this design. Therefore, the total number of components in this differential switch is exactly equal to that of the single-end switch, but the linearity or voltage requirement is expected to be better. The price is that isolation performance will be degraded without shunt transistors.

Note that in (6.2), R_{ON} can also be reduced simply by increasing the term $V_{GS} - V_{th}$ with the transistor size unchanged, which improves the insertion loss intuitively. In fact, the performance of T/R switch depends significantly on the working condition. Assuming the TX/RX nodes are biased at 0.8-V, and control voltage is 1.8/0-V, numerical calculation shows a $105\mu\text{m}/0.18\mu\text{m}$ transistor should be chosen to achieve the lowest insertion loss at 4-GHz. In this design, however, $100\mu\text{m}/0.18\mu\text{m}$ is used due to the scale limit of RF CMOS transistors provided by the foundry.

$R_{G1}-R_{G4}$ should be chosen large enough to create a floating-gate terminal at RF. Considering that large bias resistance will lower the switch speed, $3.9\text{ k}\Omega$ bias resistance is chosen.

Figure 6.3 shows the simulated differential-mode insertion loss and isolation performance of the T/R switch. It is shown that the insertion loss is less than 1.8-dB and the isolation is larger than 25-dB up to 6-GHz. Note that the sim-

Figure 6.3: Simulated insertion loss and isolation performance.

ulation is performed under 0.8-V bias voltage at each port and 1.8/0-V control voltage.

6.2.3 Layout Considerations

Practical problems in the single-ended T/R switch include substrate resistance, source/drain parasitics, DC biasing, and transistor sizing [121, 123, 124]. In differential T/R switch design, issues on crosscoupling and transistor matching between the two signal paths should also be considered carefully. Tradeoff exists among these issues, e.g. good matching of transistors requires a close placement, whereas crosstalk increases when transistors are placed near to others.

It depends on the application and system requirement to choose the layout

Figure 6.4: Layout of the proposed differential T/R switch.

strategy. In general, unintentional crosstalk and coupling increase the nonlinearity in the circuits, resulting in degradation on the insertion loss performance and power handling capability. On the other hand, mismatch is a signal independent process, which does not produce signal dependent problems as crosscoupling does, e.g. crosscoupling may distort signal waveform, which mismatch only results in a DC offset in most cases. Therefore, considerations are focused on reducing the effect of crosscoupling, at the price of matching degradation.

Figure 6.4 is the layout of the differential T/R switch. The matching transistors in the same signal path (e.g. M1 and M3, M2 and M4) are placed together without special arrangement (e.g. interdigitize) to improve the matching. The transistors in the different on/off status are placed far away to avoid crosscoupling. In addition, substrate contacts are placed densely to reduce the substrate resistance R_B .

Figure 6.5: Die photograph of the fabricated differential T/R switch.

6.3 Measurement Results and Discussions

The differential switch was fabricated in a 1.8-V one-poly six-metal 0.18- μm standard CMOS technology. The active area of the switch is $60\mu\text{m} \times 40\mu\text{m}$. Figure 6.5 shows the die microphotograph of the fabricated switch. Four G-S-S-G pads were designed for on-chip measurement purpose. The measurements were carried out using Cascade Microtech's 100- μm differential G-S-S-G probes. A four-port network analyzer was used in the experiment, which avoids the complicated on-chip balun design for testability. The control voltage is 1.8/0-V and the TX/RX nodes are biased at 0.8-V. Note that the differential impedance at each port is 100 Ω , and the common-mode impedance is 25 Ω .

In the differential T/R switch circuits, performances need to be considered include not only the differential-mode parameters, but also common-mode pa-

rameters and common-mode rejection performance. For a T/R switch that working in two status (*on* and *off*), these performances should be evaluated in both cases.

6.3.1 Differential-Mode Performance

Consider only the small differential signal applied to each port, the insertion loss and isolation are shown in Figure 6.6. An extremely low insertion loss 0.53-dB is obtained at 0.9-GHz. In the frequency range up to 2.5-GHz, the insertion loss is lower than 0.8-dB. 0.72-dB and 0.79-dB insertion loss are obtained at 1.8-GHz and 2.4-GHz, respectively. At 5.2-GHz, the insertion loss is 1.7-dB. The measured insertion loss is very close to the simulation result in Figure 6.3. The results obtained indicate that on-resistance R_{ON} of switch transistor is extremely small with large W/L ratio, whereas C_T is increased and the frequency-dependent component in (6.1) has obvious effect on the insertion loss when the operating frequency is increased. The isolation at 0.9-GHz, 1.8-GHz, 2.4-GHz and 5.2-GHz are 30-dB, 23-dB, 20-dB, 16-dB, respectively. The discrepancy from simulation may result from the inaccurate model of transistors in the cutoff region as well as the extra parasitics in the real chip.

In the frequency range from 40-MHz up to 6-GHz, the differential-mode insertion loss is within 2-dB and differential-mode isolation is above 15-dB. Although in the differential case crosscoupling and mismatch further degrade the insertion loss and isolation, these results are comparable with other single-ended CMOS T/R switches [122, 124].

Figure 6.6: Measured insertion loss and isolation parameters for differential-mode small-signals.

6.3.2 Common-Mode Performance

Figure 6.7 shows the common-mode insertion loss and isolation parameters. This is obtained by applying the common-mode small signal to each port. Comparing with the differential-mode performance, both insertion loss and isolation are degraded at high frequencies. 0.83-dB, 1.1-dB, 1.3-dB, 3.3-dB insertion loss and 28-dB, 22-dB, 20-dB, 14-dB isolation are obtained at 0.9-GHz, 1.8-GHz, 2.4-GHz and 5.2-GHz, respectively. This is reasonable since the common-mode signal with same amplitude and phase result in a more severe crosscoupling between the two signal paths. However, the common-mode signal is not of the same importance as the differential signal which affects the signal quality fatally, thus a slightly inferior performance in the common-mode is sustainable.

Figure 6.7: Measured insertion loss and isolation parameters for common-mode small-signals.

6.3.3 Common-Mode Rejection

The common-mode rejection ratio (CMRR) is measured by the forward transmission coefficient from the transmitted common-mode signal to the received differential-mode signal. Obviously, the CMRR is related to the *ON-OFF* conditions of differential switch. In the *ON* and *OFF* state, the common-mode rejection ratio are shown in Figure 6.8. It is clear that CMRR is almost constant when the switch is *ON*, and varies significantly when the switch is *OFF*. 28-dB, 27-dB, 27-dB and 27-dB CMRR in the *ON*-state and 53-dB, 45-dB, 42-dB and 33-dB CMRR in the *OFF*-state are obtained at 0.9-GHz, 1.8-GHz, 2.4-GHz and 5.2-GHz, respectively. In the *OFF* state, the CMRR is much higher, which is actually part of the isolation parameters.

Figure 6.8: Measured common-mode rejection ratio in the *ON* and *OFF* state of the differential T/R switch.

6.3.4 Power Handling Capability

The linearity of the switch determines the maximum power it can handle. Power 1-dB compression point (P_{1dB}) is used to measure the linearity. Since the switch is fully symmetric, the linearity in the transmitted mode and receive mode is identical. Figure 6.9 shows the large-signal result at 5.2-GHz. 15.2-dBm P_{1dB} is obtained. As mentioned above, the differential switch should have 3-dB higher P_{1dB} than the single-ended switch in theory. This is proved by other single-ended CMOS T/R switches where a maximum 11–12-dBm power handling was reported without tuning the substrate bias [121, 123, 129]. In the case that reliability is not a big concern, higher control voltages can be used and the resultant P_{1dB} is about 17–18-dBm [121, 122]. Under similar condition where 3.6/0-V control voltage and 1.6-V drain/source bias are used, 20-dBm P_{1dB} is obtained for this differential switch.

Figure 6.9: Linearity of the differential T/R switch at 5.2-GHz. The control voltage is 1.8/0-V and the TX/RX nodes are biased at 0.8-V.

Figure 6.10: Linearity of the differential T/R switch at different frequencies. The TX/RX nodes are biased at 0.8-V with 1.8/0-V control voltage, and 1.6-V with 3.6/0-V control voltage.

The P_{1dB} results obtained at different frequencies up to 6-GHz are shown in Figure 6.10. At the frequency of 0.9-GHz, 1.8-GHz, 2.4-GHz and 5.2-GHz, the P_{1dB} obtained are 14-dBm, 15-dBm, 15-dBm and 15-dBm. With 3.6/0-V V_{ctrl} and 1.6-V drain/source bias, the P_{1dB} are 19-dBm, 20-dBm, 20-dBm and 20-dBm, respectively. The results shown here provide possible design flexibility in the power handling and reliability.

Table 6.1 shows a summary of the measured performance of the differential T/R switch at 0.9-GHz, 1.8-GHz, 2.4-GHz and 5.2-GHz, which are the most widely used bands by current wireless systems. The results are also compared with other state-of-the-art designs. These results show that the proposed switch is suitable for the low band UWB as well as other wireless applications with moderate peak power level.

Table 6.1: Summary of Differential Switch Performance

Specifications		This work				[124] (T/R)		[129]	[122] (H/L R_B)
		0.9-GHz	1.8-GHz	2.4-GHz	5.2-GHz	[121]	2.4-GHz	5.2-GHz	5.8-GHz
Insertion Loss (dB)	DM *	0.53	0.72	0.79	1.7	0.7	1.5/1.6	1.5/1.4	1.0/0.8
	CM *	0.83	1.1	1.3	3.3				
Isolation (dB)	DM *	30	23	20	16	42	32/17	30/15	29/27
	CM *	28	22	20	14				
CMRR (dB)	ON	28	27	27	27	-	-	-	-
	OFF	53	45	42	33				
P_{1dB} (dBm)	HV †	19	20	20	20	17	-	-	18/17
	NV †	14	15	15	15	11.5	28.5/12.5	28/11.5	11 ‡

* DM—Differential Mode, CM—Common Mode.
† HV—High DC bias and control voltage, NV—Normal DC bias and control voltage.
‡ 0.2-dB gain compression point ($P_{-0.2dB}$).

Chapter 7

High-Frequency T/R Switch

7.1 Introduction

Succeeding to the preliminary differential T/R switch design discussed in Chapter 6, this chapter studies the T/R switch design towards full UWB bandwidth and higher frequency operations. Comparing with other RF IC circuits that have been pushed up to 60-GHz [2, 3, 4], CMOS T/R switches for higher frequency operations are explored only to a limited extent. Most of reported RF switches use a series/shunt architecture. At higher frequencies, the loss due to shunt-arm severely degrades the insertion loss. While the lack of shunt arm results in a low isolation, as shown in Chapter 6. This chapter presents a comprehensive consideration on insertion loss, isolation and linearity performances of CMOS T/R switch. A series only CMOS T/R switch is proposed, which is capable of ultra-wideband operations up to 20-GHz.

Section 7.2 discusses high frequency considerations of the switch transistor in cutoff region. A customized layout technique for minimizing drain-source

Figure 7.1: Schematic of series/shunt type single-ended T/R switch.

interconnect coupling of switch transistor is explored. Section 7.3 presents an improved resistive double-well body-floating technique for linearity enhancement; its impacts on insertion loss and isolation are discussed. Section 7.4 shows the experimental results and discussions.

7.2 Switch Transistor in Cutoff Region

Conventionally, the T/R switch design follows a series/shunt architecture [121, 122, 123, 125, 127, 129], as shown in Figure 7.1. The series transistors, $M1$ and $M2$, perform switch functions for TX and RX paths. The shunt transistors, $M3$ and $M4$, turn on when $M1$ and $M2$ are off, respectively, so that the undesired signal in each mode can be grounded by the shunt transistor. Previous analysis has proved that a shunt transistor improves the isolation effectively [129]. As the tradeoff, degradation on the insertion loss is observed, which is resulted from the parasitic capacitances of shunt transistors in the cutoff region. The impedances of these parasitic capacitances are large at relatively low fre-

Figure 7.2: Parasitic capacitances of a MOS transistor that is used as a switch and turned off.

quencies, and the design effort is always put into the analysis of transistors in the triode region, acting as a voltage controlled resistance. However, at high frequencies, the impedances of parasitic capacitances due to *off* transistors become small and more comparable to impedances of *on* transistors, leading to a severe impact of transistors in the cutoff region. Furthermore, observation of the series/shunt configuration shows that there is not only the shunt transistor but also another series transistor that is in the *off* state. Each *on* transistor is connected by two *off* transistors, e.g. both $M2$ and $M3$ are *off* when $M1$ is *on*. Therefore, it is essential to investigate the high frequency behavior of a switch transistor in the cutoff region.

Figure 7.2 shows a model of high-frequency parasitic capacitances for a MOS that is used as a switch and turned off. The *off* resistance between drain and source is very large and not considered here. The parasitic capacitances, couple part of the signal to ground and lead to insertion loss degradation. As the chan-

Figure 7.3: Layout sketch of drain-source interconnection for an interdigitized MOS transistor that is used as a switch.

nel is not formed, C_{gs} and C_{gd} are due to only overlap and fringing capacitances [128],

$$C_{gs} = C_{gd} = WL_{ov}C_{ox}, \quad (7.1)$$

where L_{ov} denotes the overlap distance between gate and source/drain, C_{ox} denotes the gate capacitance per unit area, W is the width of transistor. Obviously, they are very small comparing to that of a MOS in saturation or triode region. C_{sb} and C_{db} become smaller as well when the channel is not present [128]. They depend mainly on the area of source and drain, respectively.

Comparing with the small-signal model presented in [128], an extra capacitance C_{ds} is added in Figure 7.2. For a stand-alone MOS transistor, C_{ds} is very small and normally not considered. However, when the transistor is used as a switch, the metal connection style exhibits severe coupling between drain and source, and thus C_{ds} can not be ignored. This can be explained by Figure 7.3, where an interdigitized transistor layout sketch is shown. The metal connections of drain and source are in parallel and next to each other. When these

Table 7.1: Parasitic Capacitances of NMOS Transistor in Cutoff Region with $W = 108\mu m$, $L = 0.13\mu m$, fingers = 6

Parasitic Capacitances	Foundry Layout (Figure 7.4(a))	Customized Layout (Figure 7.4(b))
C_{gs} (fF)	19	19
C_{gd} (fF)	19	19
C_{gb} (fF)	3.7	7.1
C_{sb} (fF)	9.3	37
C_{db} (fF)	8.3	33
C_{ds} (fF)	36	0

metals are connected as a switch, they actually form a lateral metal capacitor, which is of high capacity [130]. This effect becomes more significant with the scale down of technologies, as the metal distance between drain and source decreases with smaller channel length.

The accurate values of these parasitic capacitances can be extracted from post layout simulations. Table 7.1 shows the extracted value of an NMOS transistor with $W = 108\mu m$, $L = 0.13\mu m$, fingers = 6. Note that some of capacitances are voltage dependent, the results shown in Table 7.1 are obtained with $V_{GS} = V_{GD} = V_{DS} = V_{SB} = V_{DB} = 0$. The values in the column *Foundry Layout* show the capacitances extracted from foundry provided layout (p-cell). A 36-fF C_{ds} is exhibited, which is much larger than any other capacitances and dominates the drain-source coupling effect in the cutoff region.

Therefore, a straightforward technique to improve isolation is reducing C_{ds} , which can be realized by increasing the distance between drain and source fingers. As shown in Figure 7.4, the drain-source distance of a customized

Figure 7.4: Foundry provided p-cell (a) and customized (b) layout of an NMOS transistor operating as a switch. $W = 108\mu m$, $L = 0.13\mu m$, fingers = 6. The drain-source distance in the customized layout is 4 times of that in the standard p-cell.

NMOS layout (Figure 7.4(b)) is 4 times of the p-cell default (Figure 7.4(a)). Table 7.1 gives also the parasitic capacitances extracted from the customized layout under the exactly same condition. It is shown that C_{ds} has been reduced to a value that can be omitted by the simulation tools. The drawback of this decoupling technique is due to the increased drain and source area, where C_{sb} and C_{db} also increase by a factor around 4. For the gate parasitics, C_{gs} and C_{gd} , there is no difference according to (7.1). C_{gb} is nearly doubled due to the increased C_{sb} and C_{db} .

The impacts of customized layout on the T/R switch performances are discussed as follows. Based on the configuration of Figure 7.1, the bulk is connected to ground, and the gate is biased through a large resistor to create a floating node at RF. When the transistor is used as the series transistor and is *on*, the increased drain and source area increases C_{sb} and C_{db} , insertion loss

will be certainly degraded. It is a typical case of operation and agrees to the well-known principle that drain and source area should be minimized in high frequency designs. However, when the transistor is *off*, the customized layout provides better isolation. The overall drain-source coupling capacitance can be written as

$$C_{OFF} = C_{ds} + \frac{C_{gs} \cdot C_{gd}}{C_{gs} + C_{gd}}. \quad (7.2)$$

It is shown that C_{OFF} is reduced significantly from 45.5-fF for the standard layout to 9.5-fF for the customized layout, which provides directly better isolation when the switch is used as the series transistor and is *off*. Furthermore, when this transistor is used as the shunt transistor, the smaller C_{OFF} leads to smaller loss, which poses positive effect on the overall insertion loss. Again, the increase of C_{sb} and C_{db} increases the loss due to shunt transistor and degrades the overall insertion loss.

Briefly, the overall performance depends severely on the transistors operating in the *off* state, especially for switch targeting on high frequency applications. The customized layout provides better isolation due to the drain-source decoupling, and degrades the insertion loss due to the increase of parasitics regarding the bulk (C_{sb} and C_{db}).

The increase of C_{gb} slightly affects the *on/off* transition speed and is not considered to affect steady-state performances. Note that the increased drain/source area also increases the ohmic loss. However, the overall resistance of drain/source region is still very small comparing to that of the *on* resistance of switch transistors. It is expected that a slight degradation of insertion loss can be observed at DC and low frequencies.

Further observing the series/shunt configuration in Figure 7.1, the shunt

Figure 7.5: Simulated insertion loss and isolation of T/R switch with and without shunt arm. The switch without shunt arm employs customized layout of Figure 7.4(b).

arm has to provide sufficient large impedance when it is turned off, so that insertion loss can be prevented from severe degradation. At high frequencies, the capacitive coupling effect becomes significant, leading to a shunt path with lower impedance and thus higher loss. At the same time, the presence of shunt arm degrades the power handling capability as the unintentional turn on of the shunt transistor increases loss significantly [124, 125]. Thus, when the isolation can be maintained with other method, the shunt arm becomes not necessary and can be removed to improve the insertion loss and linearity performances.

Figure 7.5 shows the simulated performances of T/R switches with and without shunt arms. Both results are obtained from post-layout simulations. For the switch with shunt arms, the sizes of series transistors and shunt transistors are $108\mu\text{m}/0.13\mu\text{m}/6\text{-fingers}$ and $21\mu\text{m}/0.13\mu\text{m}/3\text{-fingers}$, respectively. The layout employs standard p-cells. For the switch without shunt arm, the sizes of

series transistors are $108\mu\text{m}/0.13\mu\text{m}/6$ -fingers, and customized layout of Figure 7.4(b) is employed. In the simulations, the signal is biased at 0.5-V and the high and low control voltages are 2-V and 0-V, respectively. It is shown that the customized switch layout achieves 2-dB better isolation comparing with standard transistor layout, and this result is obtained without shunt arms. Thus, the insertion loss and isolation tradeoffs due to shunt transistors are relaxed. Benefiting from the absence of shunt arms, significant improvement of insertion loss at high frequencies can be observed. Note that at frequencies below 3-GHz, the insertion loss of customized layout is slightly higher than that of the standard layout, which is due to the increased drain/source ohmic loss and has been predicted previously. The linearity performance is also improved without shunt arms. At 10-GHz, 19-dBm $P_{1\text{dB}}$ is obtained for the switch without shunt arms, while only 15-dBm $P_{1\text{dB}}$ is obtained for the switch with shunt arms.

7.3 Double-Well Body Floating

The linearity of a T/R switch determines its power handling capability. The problems of linearity can be considered by different points of view. One is from the viewpoint of unintentional turn-on of junction diodes [121, 124]. Normally, the gate-source and gate-drain can be kept around the proper value as the gate is RF floating and its voltage is bootstrapped. But a strong signal still turns on the source-bulk and/or drain-bulk diodes, which clips the signal itself. Note that this happens in both series transistor and shunt transistor, and it is the reason that a shunt arm degrades the power handling capability. Based on this point of view, the linearity performance can be improved by either giving a large voltage headroom for these diodes or bootstrapping the bulk voltages. The former can

Figure 7.6: Simplified cross-sectional view of an NMOS transistor in triple-well technology.

be realized by large DC bias at sensitive nodes, at the price of reliability. While the latter can be done by floating the bulk at RF, which can be implemented using LC-tuned substrate bias when the bulk is not separated from the substrate [124]. Another view of linearity problem is from the DC I - V characteristics [125]. The similar body-floating approach to linearity improvement can be reached. The use of impedance transformation network can also improve the power handling capability [127].

Nowadays the triple-well process is becoming commonly available for analog and mixed-signal designs, which offers better performance in terms of noise, isolation, bulk control, etc.. Figure 7.6 shows the simplified cross-sectional view of an NMOS transistor in triple-well CMOS technology. The buried deep N -well separates the body of NMOS from the common substrate. Thus, the body can be easily biased through a large resistor. In a similar manner with gate bias, the bulk becomes RF floating and its voltage is bootstrapped to the source and drain voltages.

The resistive body-floating technique has been realized in 2.4- and 5.8-GHz

and reported in [125], the result shows that 20–21-dBm P_{1dB} can be obtained by floating the P -well through resistors. Although the area is saved comparing with LC -tuned approach, the linearity improvement is not comparable to that of the LC -tuned body-floating designs in [124], where more than 28-dBm P_{1dB} was achieved.

The comparison result leads to a further study of the diode models in triple-well process. As can be seen in Figure 7.6, the adding of the deep N -well layer creates two more diodes: the diode between P -well and deep N -well, and the diode between deep N -well and P -substrate. When the body is floated by a large resistor, the transient voltage of P -well is actually bootstrapped to the signal voltage. This prevents the source-bulk and drain-bulk diodes from being turned on by large signals and improves the linearity. However, the diode between P -well and deep N -well becomes unprotected and can be turned on by large P -well voltages. This will not happen if the P -well is biased directly by a DC voltage (real ground), but the bootstrapped P -well voltage depends on signal voltage and thus may turn on the diode between P -well and deep N -well. Once the turn-on happens, the RF-floating state of the body (P -well) is broken and the linearity will get degraded immediately.

To overcome the body-floating limitation presented above, the deep N -well should also be designed floating at RF. With both terminals floating, the diode between P -well and deep N -well is then made safe. In a similar manner, the diode between deep N -well and P -substrate is then unprotected and may be turned on by large signals. However, as the bootstrapping effect of these voltages is degressive, the signals that can turn on the outer diode need to be very large.

Figure 7.7: Simulated linearity performance for T/R switch with and without shunt arms, and with different types of body-floating.

Figure 7.7 gives the linearity simulation results with different switch configurations at 10-GHz. As discussed previously, the presence of shunt arms decreases the linearity, only 15-dBm P_{1dB} is obtained for a switch with shunt arms and 19-dBm is obtained for that without shunt arms. Both results are simulated without body floating. With a P -well-only floating resistor of 5-k Ω , the P_{1dB} is improved to 22-dBm. Dramatic improvement of linearity is observed when a second 5-k Ω resistor is used to further float the deep N -well. With resistive body-floating for both P -well and deep N -well, 29-dBm P_{1dB} is obtained. These results exhibit clearly the influence of junction diodes on the linearity performance and the efficiency of double-well body floating.

Furthermore, the body-floating will certainly affect the insertion loss and isolation performances, especially when it is combined with the proposed customized layout scheme. For a switch transistor that is turned on, the parasitic

Figure 7.8: Simulated insertion loss and isolation with and without body-floating. Both switches use the proposed customized layout and have no shunt transistors. Double-well body-floating is used.

capacitances regarding body, C_{sb} and C_{db} , will no longer affect the insertion loss as they are now floated. Thus, the negative effect of increased drain and source areas for *on*-transistor is eliminated. On the other hand, when the transistor is turned off and the body is floating, from Figure 7.2, C_{sb} and C_{db} become in series between drain and source. The increased drain-source coupling occurs, which will degrade the isolation. For the transistor parameters of Table 7.1, the coupling capacitance, C_{OFF} , increases again from 9.5-fF to 27-fF. Although it is still smaller than that from a standard p-cell, the increased coupling degrades the efficiency of the customized layout on the isolation performance. Figure 7.8 shows the simulated insertion loss and isolation for a switch with and without body-floating. Double-well body-floating is used in the comparison. Both circuits employ the proposed customized layout and have no shunt transistor. It is clear that the insertion loss is improved and isolation is degraded with body-

Figure 7.9: Final schematic and circuit parameters of the differential T/R switch.

floating technique. Note that at frequencies below 7-GHz, the insertion loss of the switch with body-floating is slightly higher than that of the switch without body-floating. This is interesting and may result from the body effect due to the large body-floating resistance.

7.4 Measurement Results and Discussions

The final circuit is shown in Figure 7.9, the dashed line denotes the deep N-well isolation. Only four switch transistors are employed, the transistor count of the proposed differential T/R switch is exactly equal to that of a single-ended switch with shunt arms. The customized layout is employed for all switch transistors.

The circuit was fabricated in a two-poly eight-metal 0.13- μm triple-well

Figure 7.10: Die microphotograph of the fabricated differential T/R switch.

CMOS technology. The cut-off frequency f_T of NMOS transistor is over 90-GHz. Figure 7.10 shows the die microphotograph of the fabricated differential T/R switch. The active area of the switch is only $180\mu\text{m} \times 150\mu\text{m}$. With test pads, the whole T/R switch chip occupies $415\mu\text{m} \times 415\mu\text{m}$ die area.

The measurements are carried out on wafer, using Cascade Microtech's differential G-S-S-G probes. A four-port network analyzer was employed in the experiment, which avoids the complicated on-chip balun design for testability. The control voltage is 2/0-V and the TX/RX nodes are biased at 0.5-V. Note that the differential impedance at each port is $100\ \Omega$, and the common-mode impedance is $25\ \Omega$.

7.4.1 Differential-Mode Performance

Figure 7.11 shows the differential-mode insertion loss and isolation performances. The insertion loss is within 2.0-dB over DC to 20-GHz. At the frequency of 0.9-, 5.8-, 10-, 15- and 20-GHz, the insertion loss is 0.7-, 1.5-, 1.7-, 1.7- and 2.0-dB, respectively. The isolation is below 21-dB at frequencies up to 20-GHz.

The results are close to the simulation results in Figure 7.8, however, the trends in Figure 7.11 are different from that in Figure 7.8. It is shown that the degradation effect of insertion loss at higher frequencies becomes slower at frequencies above 8-GHz. The differential T/R switch reported in [131] does not exhibit such trends up to 6-GHz. Review of previous publications shows this may result from the parasitic inductances. In [127], the switch designed with *LC* network shows a rather improved insertion loss at frequencies above 7-GHz, while the switch without *LC* network does not show that. The *LC* network changes the input impedance and may offer higher available input power at certain frequencies, therefore, the insertion loss may be improved at that frequency. In this design, the parasitic inductances of the input metal lines are relatively small, the *LC* resonance occurs at much higher frequencies, and the effect is not large enough to exhibit an improved insertion loss characteristic.

7.4.2 Common-Mode Performance

The common-mode performance is of little importance, it is presented here mainly for comparisons with differential-mode performance. Figure 7.12 shows the common-mode insertion loss and isolation performances. The insertion loss

Figure 7.11: Measured insertion loss and isolation parameters for differential-mode small-signals.

Figure 7.12: Measured insertion loss and isolation parameters for common-mode small-signals.

is within 7.9-dB over DC to 20-GHz. At 0.9-, 5.8-, 10-, 15- and 20-GHz, the insertion loss is 1.1-, 1.7-, 3.2-, 6.5- and 7.9-dB, respectively. The isolation is below 22-dB at frequencies up to 20-GHz.

It is shown the common-mode insertion loss is severely degraded comparing to the differential-mode signals. Similar observation can also be found in [131], which is caused by the severe coupling and loss of common-mode signals. The parasitic inductances have little impact on the common-mode insertion loss due to the large value of the loss.

Observing Figure 7.11 and Figure 7.12 shows that the *LC* resonance probably occurs at around 12-GHz. It is interesting to note the effects of such resonance: for insertion loss, the effect is positive for differential-mode signals and negative for common-mode signals; while for isolation, the effect is negative for differential-mode signals and positive for common-mode signals.

7.4.3 Common-Mode Rejection

The common-mode rejection ratio is measured by the forward transmission coefficient from the transmitted common-mode signal to the received differential-mode signal. Obviously, the CMRR is related to the *ON-OFF* conditions of differential switch. In the *ON* and *OFF* state, the common-mode rejection performances are shown in Figure 7.13. In the *ON* mode, the common-mode rejection is better than 28.7-dB. In the *OFF* mode, it is better than 40-dB.

From Figure 7.13, it is more clear that the *LC* resonance occurs at about 12-GHz. Its effects are also opposite in *ON* and *OFF* state of the switches.

Figure 7.13: Measured common-mode rejection ratio in the *ON* and *OFF* state of the differential T/R switch.

Figure 7.14: Power handling capability in terms of P_{1dB} .

7.4.4 Power Handling Capability

Figure 7.14 shows the 1-dB compression point of the differential T/R switch. The P_{1dB} increases from 24.6-dBm at 1-GHz till 30.2-dBm at 8-GHz, and it remains at the level around 30-dBm at frequencies above 8-GHz. This is reasonable considering the insertion loss degradation is reduced above 7-GHz due to the parasitic *LC* network. Similar improvement on linearity performance can be found in [127], where *LC* network is designed intentionally for higher power handling capability.

Note that 3-dB improvement in P_{1dB} is benefited from the differential architecture, without which the P_{1dB} is around 27-dBm. It is slightly lower than the simulated value of 29-dBm. Comparing with the 21.3-dBm and 20-dBm P_{1dB} obtained by single-well resistive body-floating [125], the power handling capability is improved significantly using double-well body floating technique.

Table 7.2 shows a summary of the measured performance of the differential T/R switch. The comparison with other reported state-of-the-art CMOS T/R switch designs is also presented. The T/R switch presented in this work achieves 20-GHz operating frequency, which is the highest among reported CMOS T/R switch designs. The bandwidth and linearity is suitable for UWB wireless transceiver front-ends. It can also be used in other potential wireless applications with up to 30-dBm transmit power level and 20-GHz bandwidth.

Table 7.2: Summary of the T/R Switch Performances and Comparison with Reported State-of-the-Art Designs

	Insertion Loss (dB)					Isolation (dB)					P _{1dB} (dBm)
	2.4-GHz	5.8-GHz	10-GHz	15-GHz	20-GHz	2.4-GHz	5.8-GHz	10-GHz	15-GHz	20-GHz	
This Work *	0.9	1.5	1.7	1.7	2.0	43	34	32	32	21	30
[127] †	—	—	1.1	1.8	—	35	29	24	18	—	21.5
[125] †	0.7	1.1	—	—	—	35	27	—	—	—	21.3
[124] †	1.5	1.4 ‡	—	—	—	32	30 ‡	—	—	—	28.5
[122] †	—	0.8	—	—	—	—	29	—	—	—	18
[129]	1.5	—	—	—	—	24	—	—	—	—	11 [¶]
[131] *	0.79	1.7 ‡	—	—	—	20	16 ‡	—	—	—	20

* Performances in differential-mode.
† Best available insertion loss and isolation in case of multiple designs, or highest P_{1dB} among different frequencies.
‡ Performances at 5.2-GHz.
[¶] 0.2-dB gain compression point (P_{-0.2dB}).

Chapter 8

Conclusions and Recommendations

8.1 Conclusions

The thesis studied several implementation-oriented issues for CMOS integrated UWB impulse radio transceiver. The performance of realistic UWB impulse radio was evaluated considering realistic antennas and deterministic multipath channels. The RF integrated circuit design for UWB transceiver front-end was also studied. UWB low-noise amplifier and T/R switch were designed and silicon verified in 0.18- and 0.13- μm CMOS technologies. In detail, the work is summarized as follows.

A general framework for realistic system design and evaluation with non-ideal block performance was proposed based on waveform distortion analysis. A PPM and correlation based UWB transceiver with three widely proposed UWB antennas was studied in four deterministic multipath channels which demonstrate various UWB applications. The waveform distortion effect of realistic front-end components and resultant BER performance degradation of the over-

all system were discussed. The results indicate that slot antenna is optimal in LOS channel while diamond dipole antenna performs better in multipath environments. The ISI effects on waveform distortion and system performance degradation were also demonstrated. To achieve better performance of the overall system, different components and circuits in the system should be co-designed to compensate the performance fluctuations of individual blocks.

The ultra-wideband on-human-body channel for wireless body area networks was studied, including the characterization of the on-human-body UWB channel and its impact on the UWB WBAN system performance in terms of bit-error rate. The channel characterization was based on the measurement in indoor and anechoic chamber environments. The radiographs of path loss and delay spread were generated for the first time, which contain both qualitative and quantitative channel information. The results show that the path loss exponent is around $2.7 \sim 3.7$ and the maximum value of the rms delay spread is 12 ns. The rms delay spread values are less than 6 ns 50% of the time and less than 9 ns 90% of the time. The impact of on-human-body channel on the overall UWB WBAN system was analyzed through the waveform distortion. Regarding the overall system performance, the results indicate that the human body effect is more significant than the environment effect, especially when the propagation channel contains no LOS path. Various candidate monocycle pulse shapes and modulation schemes were compared, showing that PPM is very sensitive to rms delay spread and OOK is less sensitive to channel environments. The most suitable pulse shape depends on the modulation schemes employed and should be studied case by case.

An inductorless LNA capable of UWB applications was designed and fabricated. Without on-chip inductors, the ultra-wide -3 -dB bandwidth is achieved

by a syncretic adoption of thermal noise canceling, capacitor peaking, and current reuse. A gain-enhanced architecture based on noise canceling principle was proposed and detailed analysis of the stage is performed. Modified noise canceling equations were presented. Capacitive peaking is not able to enhance the absolute bandwidth, however, it is an effective way of controlling the gain flatness. Fabricated in a 0.13- μm triple-well CMOS process, the LNA achieves a small signal gain of 11-dB and a -3-dB bandwidth of 2–9.6-GHz with $S_{11} < -8.3$ -dB in-band impedance matching. The noise figure is 3.6–4.8-dB over 2–9.6-GHz. The LNA consumes 19-mW from a low supply voltage of 1.5-V. The performance achieved in this inductorless design is comparable and sometimes better than inductor-based designs. The silicon area is reduced significantly. The LNA circuit with test pads occupies only 0.17 mm², which is among the smallest designs reported in literature.

The feasibility of differential T/R switch in CMOS was demonstrated, which improves the power handling capability without performance degradation of insertion loss and isolation. A fully differential T/R switch was implemented in a 0.18- μm standard CMOS technology. The design consideration of the differential switch was discussed. Measurement results exhibit less than 2-dB insertion loss, higher than 15-dB isolation with reasonable common-mode rejection performance at frequencies up to 6-GHz. 15-dBm power 1-dB compression point was obtained which is theoretically 3-dB superior to the single-end switches. Effects of crosscoupling and mismatch can be reduced by careful layout. The broadband characteristics make it suitable for multi-standard and low band UWB wireless applications with moderate power level.

At frequencies above 6-GHz, the widely used series-shunt architecture of CMOS T/R switch encounters difficulties in achieving good insertion loss with

reasonable isolation and linearity performances. The bottleneck lies in the operation of transistors in the cutoff region. This issue was investigated and a customized transistor layout was proposed to minimize the parasitic capacitances due to drain-source interconnections, which permits a series-only architecture with better insertion loss and reasonable isolation. Benefitting from the triple-well process, the resistive body-floating technique can be used to improve the wideband power handling capability. A double-well resistive body floating was explored and the body-floating effects on the insertion loss and isolation performances were studied. The switch employs also the differential architectures for better linearity and other performances. The circuit analysis and design approaches were verified by experimental results. Fabricated in 0.13- μm triple-well CMOS, the T/R switch exhibits less than 2-dB insertion loss and higher than 21-dB isolation up to 20-GHz. 30-dBm $P_{1\text{dB}}$ is obtained. The results shows that CMOS T/R switch is capable of ultra-wideband and even higher frequency operations up to 20-GHz.

8.2 Recommendations for further research

The performance analysis presented in the thesis was based on the deterministic transfer functions. Although the evaluation framework can be adopted in various cases as long as the transfer functions are available, specific antennas and deterministic channel models have to be employed. This prevents the general utilization of the method. To develop a system evaluation methodology with application generality is of great interest. At the same time, not only the gain that will affect the system performance. As a time-domain approach, precise timing is crucial for the impulse radio UWB system. The impact of timing

jitter on the system performance is of great interest. In the WBAN part of the work, only one type of antenna was used, to investigate the effects considering different antenna types may get more general conclusions.

For the circuit part of work presented in the thesis, the design of both LNA and T/R switch can be potentially improved. For the LNA, the capacitive degeneration was employed for inductorless bandwidth enhancement, which requires severe gain-bandwidth tradeoff and degrades the NF as well. Novel technique for inductorless peaking is highly demanded. RF IC Design for packaged chip with electrostatic discharge (ESD) protection is also quite challenging and has much room of improvement. In the high-frequency T/R switch studies, it is found that body-floating technique and customized layout strategy are not always positive to isolation performance. Circuit techniques to relax the tradeoffs and improve the performance can be further investigated.

Finally, to design and implement a monolithic UWB transceiver is of ultimate interest and is very challenging, a lot of design issues should be further studied.

Author's Publications

Refereed Journal Papers

1. Qiang Li, and Y. P. Zhang, "CMOS T/R switch design: towards ultra-wideband and higher frequency," *IEEE Journal of Solid-State Circuits*, vol. 42, no. 3, pp. 563–570, Mar. 2007.
2. Qiang Li, Y. P. Zhang, K. S. Yeo and W. M. Lim, "16.6- and 28-GHz fully integrated CMOS RF switches with improved body-floating," *IEEE Journal of Solid-State Circuits*, submitted and under review.
3. Y. P. Zhang, Qiang Li, F. Wei, C. H. Ang, and H. Li, "A differential CMOS T/R switch for multistandard applications," *IEEE Transactions on Circuits and Systems–II*, vol. 53, no. 8, pp. 782–786, Aug. 2006.
4. Qiang Li, and Y. P. Zhang, "A 1.5-V 2–9.6-GHz inductorless low-noise amplifier in 0.13- μm CMOS," *IEEE Transactions on Microwave Theory and Techniques*, submitted and under review.
5. Qiang Li, and Y. P. Zhang, "Alternative approach to low-noise amplifier design for ultra-wideband applications," *International Journal of RF and Microwave Computer-Aided Engineering*, vol. 17, no. 2, pp. 153–159, Mar.

2007.

6. Y. P. Zhang, and Qiang Li, "Performance of UWB impulse radio with planar monopoles over on-human-body propagation channel for wireless body area networks," *IEEE Transactions on Antennas and Propagation*, revised and under review.
7. Qiang Li, and Y. P. Zhang, "Waveform distortion and performance of ultra-wideband radio with realistic antennas in deterministic multipath channels," *International Journal of Antennas and Propagation*, submitted and under review.
8. Y. P. Zhang, J. J. Wang, Qiang Li, and X. J. Li, "Antenna and transmit/receive switch for single-chip radio transceivers of differential architecture," *IEEE Transactions on Antennas and Propagation*, submitted and under review.
9. X. J. Li, Y. P. Zhang, and Qiang Li, "Design and analysis of a single-pole double-throw transmit/receive switch in 0.18- μm CMOS," *International Journal of RF and Microwave Computer-Aided Engineering*, submitted and under review.

Refereed Conference Papers

1. Qiang Li, Y. P. Zhang, and J. S. Chang, "An inductorless low-noise amplifier with noise cancellation for UWB receiver front-end," *2006 IEEE Asian Solid-State Circuits Conference (A-SSCC)*, Hangzhou, China, Nov. 13–15, 2006, pp. 267–270.

2. Qiang Li, Y. P. Zhang, and G. X. Zheng, "Gain-flatness considerations on the ultra-wideband low-noise amplifier design," *IEEE 17th Asia Pacific Microwave Conference (APMC)*, Suzhou, China, Dec. 4–7, 2005.
3. Qiang Li, and Y. P. Zhang, "Implementation oriented analysis of ultra-wideband front-end," *10th International Symposium on Integrated Circuits, Devices and Systems (ISIC)*, Singapore, Sept. 8–10, 2004.

Filed Patents

1. *Switch Transistors*, US patent, filed Sept. 21, 2006.
2. *A Low Noise Amplifier Circuit with Noise Cancellation and Increased Gain*, US patent, filed Oct. 04, 2006.

Bibliography

- [1] P. Smulders, "Exploiting the 60 GHz band for local wireless multimedia access: prospects and future directions," *IEEE Commun. Mag.*, vol. 40, no. 1, pp. 140–147, Jan. 2002.
- [2] C. H. Doan, S. Emami, A. M. Niknejad, and R. W. Brodersen, "Design of CMOS for 60GHz applications," in *2004 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, vol. 1, San Francisco, CA, Feb. 15–19, 2004, pp. 440–441.
- [3] B. Razavi, "A 60-GHz CMOS receiver front-end," *IEEE J. Solid-State Circuits*, vol. 41, no. 1, pp. 17–22, Jan. 2006.
- [4] C. H. Doan, S. Emami, A. M. Niknejad, and R. W. Brodersen, "A 60GHz receiver and transmitter chipset for broadband communications in silicon," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 184–185.
- [5] Federal Communications Commission, *Revision of Part 15 of the Commission's Rules Regarding Ultra-Wideband Transmission Systems*, Mar. 2003.
- [6] J. Foerster, E. Green, S. Somayazulu, and D. Leeper, "Ultra-wideband technology for short- or medium-range wireless communications," *Intel Technol. Journal Q2*, 2001.
- [7] D. Porcino and W. Hirt, "Ultra-wideband radio technology: potential and challenges ahead," *IEEE Commun. Mag.*, vol. 41, no. 7, pp. 66–74, July 2003.
- [8] S. Roy, J. Foerster, V. Somayazulu, and D. Leeper, "Ultrawideband radio design: the promise of high-speed, short-range wireless connectivity," *Proc. IEEE*, vol. 92, no. 2, pp. 295–311, Feb. 2004.

- [9] R. A. Scholtz, "Multiple access with time-hopping impulse modulation," in *IEEE 1993 Military Commun. Conf. (MILCOM)*, vol. 2, Oct. 1993, pp. 447–450.
- [10] M. Z. Win and R. A. Scholtz, "Ultra-wide bandwidth time-hopping spread-spectrum impulse radio for wireless multiple-access communications," *IEEE Trans. Commun.*, vol. 48, no. 4, pp. 679–691, Apr. 2000.
- [11] R. Aiello, J. Ellis, U. Kareev, K. Siwiak, and L. Taylor. Understanding UWB – principles & implications for low-power communications. IEEE 802.15 Working Group. Doc.: IEEE 802.15-03/157r1. [Online]. Available: <http://www.ieee802.org/15/pub/Download.html>
- [12] T. W. Barrett, "History of UltraWideband (UWB) radar & communications: Pioneers and innovators," in *Proc. Progress in Electromagn. Symp. (PIERS) 2000*, Cambridge, MA, July 2000, pp. 1–29.
- [13] D. C. Laney, "Modulation, coding and RF components for ultra-wideband impulse radio," Ph.D. dissertation, University of California, San Diego, 2003.
- [14] MIT Ultra-Wideband Group. [Online]. Available: <http://www-mtl.mit.edu/researchgroups/icsystems/uwb/>
- [15] L. Larson, D. Laney, and J. Jamp, "An overview of hardware requirements for UWB systems: interference issues and transceiver design implications," in *IEEE 2003 Military Commun. Conf. (MILCOM)*, vol. 2, Oct. 13–16, 2003, pp. 863–867.
- [16] L. Yang and G. Giannakis, "A general model and SINR analysis of low duty-cycle UWB access through multipath with narrowband interference and rake reception," *IEEE Trans. Wireless Commun.*, vol. 4, no. 4, pp. 1818–1833, July 2005.
- [17] J. R. Foerster, "Interference modeling of pulse-based UWB waveforms on narrowband systems," in *IEEE 55th Veh. Technol. Conf. (VTC) Spring*, vol. 4, May 6–9, 2002, pp. 1931–1935.

- [18] M. Hamalainen, V. Hovinen, R. Tesi, J. H. J. Linatti, and M. Latva-aho, "On the UWB system coexistence with GSM900, UMTS/WCDMA, and GPS," *IEEE J. Select. Areas Commun.*, vol. 20, no. 9, pp. 1712–1721, Dec. 2002.
- [19] J. Bellorado, S. S. Ghassemzadeh, L. J. Greenstein, T. Sveinsson, and V. Tarokh, "Coexistence of ultra-wideband systems with IEEE-802.11 a wireless LANs," in *IEEE 2003 Global Telecommun. Conf. (GLOBECOM '03)*, vol. 1, Dec. 1–5, 2003, pp. 410–414.
- [20] R. A. Scholtz and M. Z. Win, "Impulse radio: how it works," *IEEE Commun. Lett.*, vol. 2, pp. 36–38, Feb. 1998.
- [21] R. A. Scholtz, R. Weaver, E. Homier, J. Lee, P. Hilmes, A. Taha, and R. Wilson, "UWB deployment challenges," in *Proc. 2000 IEEE PIMRC*, vol. 1, Sept. 2000, pp. 620–625.
- [22] G. D. Weeks, J. K. Townsend, and J. A. Freebersyser, "Performance of hard decision detection for impulse radio," in *IEEE 1999 Military Commun. Conf. (MILCOM)*, vol. 2, Oct. 31 –Nov. 3, 1999, pp. 1201–1206.
- [23] R. Fisher, R. Kohno, H. Ogawa, H. Zhang, and K. Takizawa. (Nov. 10, 2003) Merger#2 proposal DS-CDMA. IEEE 802.15 Working Group. Doc.: IEEE 802.15-03/334r5. [Online]. Available: <http://www.ieee802.org/15/pub/Download.html>
- [24] R. W. Brodersen, "Issues in CMOS implementation of UWB radios," in *2003 IEEE RFIC Symp. UWB Workshop*, June 2003.
- [25] U.C. Berkeley UWB Group. [Online]. Available: <http://bwrc.eecs.berkeley.edu/Research/UWB/>
- [26] *Mono-Phase and Bi-Phase Ultra-Wideband WHITE PAPER*, XtremeSpectrum, Inc., May 2002.
- [27] D 4.1.1 - design and performance analysis of UWB communication system for low data rate WPAN applications. [Online]. Available: <http://www.imec.be/pacwoman/deliverables.shtml>
- [28] [Online]. Available: <http://www.ieee802.org/15/pub/TG3.html>

- [29] [Online]. Available: <http://grouper.ieee.org/groups/802/15/pub/04/>
- [30] A. Batra *et al.* (Sept. 17, 2003) Multi-band OFDM physical layer proposal. IEEE 802.15 Working Group. Doc.: IEEE 802.15-02/372r8. [Online]. Available: <http://www.ieee802.org/15/pub/Download.html>
- [31] ——. (Nov. 10, 2003) Multi-band OFDM physical layer proposal for IEEE 802.15 task group 3a. IEEE 802.15 Working Group. Doc.: IEEE 802.15-03/268r2. [Online]. Available: <http://www.ieee802.org/15/pub/Download.html>
- [32] Freescale's Ultra-Wideband. [Online]. Available: <http://www.freescale.com/webapp/sps/site/overview.jsp?nodeId=02XPgQhHPR0220>
- [33] WiMedia Alliance. [Online]. Available: <http://www.wimedia.org/>
- [34] T. S. Rappaport, *Wireless Communications: Principles and Practice*, 2nd ed. Prentice Hall PTR, 2001.
- [35] X. Chen and S. Kiaei, "Monocycle shapes for ultra wideband system," in *Proc. 2002 IEEE Int. Symp. Circuits and Systems*, vol. 1, May 2002, pp. I-597–I-600.
- [36] Y. Wu, A. F. Molisch, S.-Y. Kung, and J. Zhang, "Impulse radio pulse shaping for ultra-wide bandwidth (UWB) systems," in *Proc. 2003 IEEE PIMRC*, vol. 1, Sept. 7–10, 2003, pp. 877–881.
- [37] K. P. Wallace, A. B. Parr, B. L. Cho, and Z. Ding, "Performance analysis of a spectrally compliant ultra-wideband pulse design," *IEEE Trans. Wireless Commun.*, vol. 4, no. 5, pp. 2172–2181, Sept. 2005.
- [38] G. Lu, P. Spasojevic, and L. Greenstein, "Antenna and pulse designs for meeting UWB spectrum density requirements," in *Proc. 2003 IEEE Conf. Ultra-Wideband Syst. Technol. (UWBST)*, Nov. 16–19, 2003, pp. 162–166.
- [39] H. L. Bertoni, *Radio Propagation for Modern Wireless Systems*. Upper Saddle River, NJ: Prentice Hall PTR, 1999.
- [40] M. M. Pietrzyk and J. H. Weber, "Performance of UWB-IR with polarity randomization and interleaved coding-modulation on multipath fading

- channels,” in *IEEE 61th Veh. Technol. Conf. (VTC) Spring*, vol. 2, May 30–June 1, 2005, pp. 1365–1369.
- [41] E. Baccarelli, M. Biagi, C. Pelizzoni, and P. Bellotti, “A novel multi-antenna impulse radio UWB transceiver for broadband high-throughput 4G WLANs,” *IEEE Commun. Lett.*, vol. 8, no. 7, pp. 419–421, July 2004.
- [42] S. sheu Tan, B. Kannan, and A. Nallanathan, “Performance of UWB multiple access impulse radio systems in multipath environment with antenna array,” in *IEEE 2003 Global Telecommun. Conf. (GLOBECOM '03)*, vol. 4, Dec. 1–5, 2003, pp. 2182–2186.
- [43] A. A. Abidi, “RF CMOS comes of age,” *IEEE J. Solid-State Circuits*, vol. 39, no. 4, pp. 549–561, Apr. 2004.
- [44] R. Singh, M. M. Oprysko, and D. Harame, *Silicon Germanium: Technology, Modeling, and Design*. Wiley-IEEE Press, 2003.
- [45] A. Bevilacqua and A. M. Niknejad, “An ultra-wideband CMOS LNA for 3.1 to 10.6 GHz wireless receivers,” in *2004 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 15–19, 2004, pp. 382–383.
- [46] A. Ismail and A. Abidi, “A 3 to 10 GHz LNA using a wideband LC-ladder matching network,” in *2004 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 15–19, 2004, pp. 384–385.
- [47] J. Bergervoet, K. Harish, G. van der Weide, D. Leenaerts, R. van de Beek, H. Waite, Y. Zhang, S. Aggarwal, C. Razzell, and R. Roovers, “An interference robust receive chain for UWB radio in SiGe BiCMOS,” in *2005 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 6–10, 2005, pp. 200–201.
- [48] R. Roovers, D. M. W. Leenaerts, J. Bergervoet, K. S. Harish, R. C. H. van de Beek, G. van der Weide, H. Waite, Y. Zhang, S. Aggarwal, and C. Razzell, “An interference-robust receiver for ultra-wideband radio in SiGe BiCMOS technology,” *IEEE J. Solid-State Circuits*, vol. 40, no. 12, pp. 2563–2572, Dec. 2005.

- [49] A. Ismail and A. Abidi, "A 3.1 to 8.2GHz direct conversion receiver for MB-OFDM UWB communications," in *2005 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 6–10, 2005, pp. 208–209.
- [50] —, "A 3.1- to 8.2-GHz zero-IF receiver and direct frequency synthesizer in 0.18- μ m SiGe BiCMOS for mode-2 MB-OFDM UWB communication," *IEEE J. Solid-State Circuits*, vol. 40, no. 12, pp. 2573–2582, Dec. 2005.
- [51] B. Razavi, T. Aytur, F.-R. Yang, R.-H. Yan, H.-C. Kang, C.-C. Hsu, and C.-C. Lee, "A 0.13 μ m CMOS UWB transceiver," in *2005 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 6–10, 2005, pp. 216–217.
- [52] B. Razavi, T. Aytur, C. Lam, F.-R. Yang, K.-Y. Li, R.-H. Yan, H.-C. Kang, C.-C. Hsu, and C.-C. Lee, "A UWB CMOS transceiver," *IEEE J. Solid-State Circuits*, vol. 40, no. 12, pp. 2555–2562, Dec. 2005.
- [53] F. Lee and A. Chandrakasan, "A BiCMOS ultra-wideband 3.1-10.6GHz front-end," in *Proc. IEEE 2005 Custom Integrated Circuits Conf. (CICC)*, Sept. 18–21, 2005, pp. 153–156.
- [54] S. Iida, K. Tanaka, H. Suzuki, N. Yoshikawa, N. Shoji, B. Griffiths, D. Mellor, F. Hayden, I. Butler, and J. Chatwin, "A 3.1 to 5 GHz CMOS DSSS UWB transceiver for WPANs," in *2005 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 6–10, 2005, pp. 214–215.
- [55] T. Terada, S. Yoshizumi, Y. Sanada, and T. Kuroda, "A CMOS impulse radio ultra-wideband transceiver for 1Mb/s data communications and ± 2.5 cm range findings," in *IEEE 2005 Symp. VLSI Circuits Dig. of Tech. Papers*, June 16–18, 2005, pp. 30–33.
- [56] T. Terada, S. Yoshizumi, M. Muqsith, Y. Sanada, and T. Kuroda, "A CMOS ultra-wideband impulse radio transceiver for 1-Mb/s data communications and ± 2.5 -cm range finding," *IEEE J. Solid-State Circuits*, vol. 41, no. 4, pp. 891–898, Apr. 2006.

- [57] J. Ryckaert, M. Badaroglu, V. De Heyn, G. Van der Plas, P. Nuzzo, A. Baschiroto, S. D'Amico, C. Desset, H. Suys, M. Libois, B. Van Poucke, P. Wambacq, and B. Gyselinckx, "A 16mA UWB 3-to-5GHz 20Mpulses/s quadrature analog correlation receiver in 0.18- μ m CMOS," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 114–115.
- [58] Y. Zheng, Y. Tong, C. W. Ang, Y.-P. Xu, W. G. Yeoh, F. Lin, and R. Singh, "A CMOS carrier-less UWB transceiver for WPAN applications," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 116–117.
- [59] A. Tanaka, H. Okada, H. Kodama, and H. Ishikawai, "A 1.1V 3.1-to-9.5GHz MB-OFDM UWB transceiver in 90nm CMOS," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 120–121.
- [60] C. Sandner, S. Derksen, D. Draxelmayer, S. Ek, V. Filimon, G. Leach, S. Marsili, D. Matveev, K. Mertens, F. Michl, H. Paule, M. Punzenberger, C. Reindl, R. Salerno, M. Tiebout, A. Wiesbauer, I. Winter, and Z. Zhang, "A WiMedia/MBOA-compliant CMOS RF transceiver for UWB," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 122–123.
- [61] S. Lo, I. Sever, S.-P. Ma, P. Jang, A. Zou, C. Arnott, K. Ghatak, A. Schwartz, L. Huynh, and T. Nguyen, "A dual-antenna phased-array UWB transceiver in 0.18 μ m CMOS," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 118–119.
- [62] T. Aytur, H.-C. Kang, R. Mahadevappa, M. Altintas, S. ten Brink, T. Diep, C.-C. Hsu, F. Shi, F.-R. Yang, C.-C. Lee, R.-H. Yan, and B. Razavi, "A fully integrated UWB PHY in 0.13 μ m CMOS," in *2006 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, San Francisco, CA, Feb. 5–9, 2006, pp. 124–125.

- [63] D. M. Pozar, "Waveform optimizations for ultra-wideband radio systems," *IEEE Trans. Antennas Propagat.*, vol. 51, no. 9, pp. 2335–2345, Sept. 2003.
- [64] —, "Closed-form approximations for link loss in a UWB radio system using small antennas," *IEEE Trans. Antennas Propagat.*, vol. 51, no. 9, pp. 2346–2352, Sept. 2003.
- [65] B. Uguen, E. Plouhinec, Y. Lostanlen, and G. Chassay, "A deterministic ultra wideband channel modeling," in *Proc. 2002 IEEE Conf. Ultra-Wideband Syst. Technol. (UWBST)*, May 21–23, 2002, pp. 1–5.
- [66] J. W. Schuster and R. J. Luebbers, "Using FDTD to evaluate ray-tracing models for propagation prediction in urban microcells," in *Proc. 1998 IEEE AP-S Int. Symp.*, vol. 3, June 1998, pp. 1676–1679.
- [67] Qiang Li, "Design of a CMOS Ultra-Wideband Receiver," PhD First Year Report, Nanyang Technological University, Singapore, Jan. 2004.
- [68] S. N. Makarov, *Antenna and EM Modeling with MATLAB*. New York: Wiley-Interscience, 2002.
- [69] Y. P. Zhang, "Integrated circuit ceramic ball grid array package antenna," *IEEE Trans. Antennas Propagat.*, vol. 52, no. 10, pp. 2538–2544, Oct. 2004.
- [70] H. Schantz and M. Barnes, "The COTAB UWB magnetic slot antenna," in *Proc. 2001 IEEE AP-S Int. Symp.*, vol. 4, July 8–13, 2001, pp. 104–107.
- [71] H. Schantz and L. Fullerton, "The diamond dipole: a Gaussian impulse antenna," in *Proc. 2001 IEEE AP-S Int. Symp.*, vol. 4, July 8–13, 2001, pp. 100–103.
- [72] J. Yeo, Y. Lee, and R. Mittra, "Design of a wideband planar volcano-smoke slot antenna (PVSA) for wireless communications," in *Proc. 2003 IEEE AP-S Int. Symp.*, vol. 2, June 22–27, 2003, pp. 655–658.
- [73] Y. P. Zhang, "Novel model for propagation loss prediction in tunnels," *IEEE Trans. Veh. Technol.*, vol. 52, no. 5, pp. 1308–1314, Sept. 2003.

- [74] W. Honcharenko, H. Bertoni, and J. Dailing, "Mechanisms governing propagation between different floors in buildings," *IEEE Trans. Antennas Propagat.*, vol. 41, no. 6, pp. 787–790, June 1993.
- [75] "Wireless world research forum: Book of visions 2001," Dec. 2001.
- [76] D. Marculescu, R. Marculescu, S. Park, and S. Jayaraman, "Ready to ware," *IEEE Spectrum*, vol. 40, no. 10, pp. 28–32, Oct. 2003.
- [77] E. Staderini, "UWB radars in medicine," *IEEE Aerosp. Electron. Syst. Mag.*, vol. 17, no. 1, pp. 13–18, Jan. 2002.
- [78] A. Tan and M. Chia, "UWB radar transceiver and measurement for medical imaging," in *2004 IEEE Int. Workshop Biomed. Circuits Syst.*, Singapore, Dec. 1–3, 2004, pp. S3.1–9–S3.1–12.
- [79] S. Promwong, W. Hachitani, G. Ching, and J. Takada, "Characterization of ultra-wideband antenna with human body," in *2004 IEEE Int. Symp. Commun. Inform. Theory*, vol. 2, Oct. 26–29, 2004, pp. 1213–1217.
- [80] T. Welch, R. Musselman, B. Emessiene, P. Gift, D. Choudhury, D. Casadine, and S. Yano, "The effects of the human body on UWB signal propagation in an indoor environment," *IEEE J. Select. Areas Commun.*, vol. 20, no. 9, pp. 1778–1782, Dec. 2002.
- [81] S. S. Ghassemzadeh, R. Jana, C. W. Rice, W. Turin, and V. Tarokh, "A statistical path loss model for in-home UWB channels," in *Proc. 2002 IEEE Conf. Ultra-Wideband Syst. Technol. (UWBST)*, May 16–23, 2002, pp. 59–64.
- [82] T. Zasowski, F. Althaus, M. Stager, A. Wittneben, and G. Troster, "UWB for noninvasive wireless body area networks: channel measurements and results," in *Proc. 2003 IEEE Conf. Ultra-Wideband Syst. Technol. (UWBST)*, Nov. 16–19, 2003, pp. 285–289.
- [83] A. Alomainy, Y. Hao, C. G. Parini, and P. S. Hall, "On-body propagation channel characterisation for UWB wireless body-centric network," in *Proc. 2005 IEEE AP-S Int. Symp.*, Washington, USA, July 3–8, 2005.

- [84] A. Fort, C. Desset, J. Ryckaert, P. De Doncker, L. Van Biesen, and S. Donnay, "Ultra wide-band body area channel model," in *2005 IEEE Int. Conf. Commun.*, vol. 4, Korea, May 16–20, 2005, pp. 2840–2844.
- [85] Y. Hao, A. Alomainy, W. Song, C. G. Parini, P. Hall, Y. Nechayev, and C. Constantinou, "Numerical modeling of on-body radio propagation channel," in *Proc. 2005 IEEE AP-S Int. Symp.*, Washington, USA, July 3–8, 2005.
- [86] A. Alomainy, Y. Hao, C. Parini, and P. Hall, "Comparison between two different antennas for UWB on-body propagation measurements," *IEEE Antennas Wireless Propagat. Lett.*, vol. 4, pp. 31–34, 2005.
- [87] M. Kamarudin, Y. Nechayev, and P. Hall, "Antennas for on-body communication systems," in *2005 IEEE Int. Workshop Antenna Technol.*, Singapore, Mar. 7–9, 2005, pp. 17–20.
- [88] I. Kovacs, G. Pedersen, P. Eggers, and K. Olesen, "Ultra wideband radio propagation in body area network scenarios," in *2004 IEEE 8th Int. Symp. Spread Spectr. Tech. Applicat.*, Aug. 30 –Sept. 2, 2004, pp. 102–106.
- [89] J. Keignart and N. Daniele, "Subnanosecond UWB channel sounding in frequency and temporal domain," in *Proc. 2002 IEEE Conf. Ultra-Wideband Syst. Technol. (UWBST)*, May 21–23, 2002, pp. 25–30.
- [90] A. Molisch, J. Foerster, and M. Pendergrass, "Channel models for ultrawideband personal area networks," *IEEE Trans. Wireless Commun.*, vol. 10, no. 6, pp. 14–21, Dec. 2003.
- [91] L. Rusch, C. Prettie, D. Cheung, Q. Li, and M. Ho. Characterization of UWB propagation from 2 to 8 GHz in a residential environment. Intel Research Labs. [Online]. Available: <http://www.intel.com/technology/comms/uwb/docs.htm>
- [92] Z. Low, J. Cheong, and C. Law, "Low-cost PCB antenna for UWB applications," *IEEE Antennas Wireless Propagat. Lett.*, vol. 4, pp. 237–239, 2005.
- [93] P. Salonen, Y. Rahmat-Samii, and M. Kivikoski, "Wearable antennas in the vicinity of human body," in *Proc. 2004 IEEE AP-S Int. Symp.*, June 20–25, 2004.

- [94] Y. P. Zhang and F. Wang, "Time-delay characteristics of in-room UHF radio propagation channels," *Microwave Optical Technol. Lett.*, vol. 33, no. 2, pp. 115–119, Apr. 2002.
- [95] R.-C. Liu, C.-S. Lin, K.-L. Deng, and H. Wang, "A 0.5-14-GHz 10.6-dB CMOS cascode distributed amplifier," in *IEEE Symp. VLSI Circuits Technol. Dig.*, Kyoto, Japan, June 12–14, 2003, pp. 139–140.
- [96] C. Grewing, M. Friedrich, G. Li Puma, C. Sandner, S. van Waasen, A. Wiesbauer, and K. Winterberg, "Fully integrated ultra wide band CMOS low noise amplifier," in *Proc. IEEE 30th Europ. Solid-State Circuits Conf. (ESSCIRC)*, Sept. 21–23, 2004, pp. 435–438.
- [97] Q. He and M. Feng, "Low-power, high-gain, and high-linearity SiGe BiCMOS wide-band low-noise amplifier," *IEEE J. Solid-State Circuits*, vol. 39, no. 6, pp. 956–959, June 2004.
- [98] F. Ellinger, "60-GHz SOI CMOS traveling-wave amplifier with NF below 3.8 dB from 0.1 to 40 GHz," *IEEE J. Solid-State Circuits*, vol. 40, no. 2, pp. 553–558, Feb. 2005.
- [99] M.-D. Tsai, K.-Y. Lin, and H. Wang, "A 5.4-mW LNA using 0.35- μ m SiGe BiCMOS technology for 3.1-10.6-GHz UWB wireless receivers," in *IEEE Radio Frequency Integrated Circuits (RFIC) Symp. Dig. Papers*, June 12–14, 2005, pp. 335–338.
- [100] F. Ellinger, "26-42 GHz SOI CMOS low noise amplifier," *IEEE J. Solid-State Circuits*, vol. 39, no. 3, pp. 522–528, Mar. 2004.
- [101] D. Barras, F. Ellinger, H. Jackel, and W. Hirt, "A low supply voltage SiGe LNA for ultra-wideband frontends," *IEEE Microwave Wireless Compon. Lett.*, vol. 14, no. 10, pp. 469–471, Oct. 2004.
- [102] C.-W. Kim, M.-S. Kang, P. T. Anh, H.-T. Kim, and S.-G. Lee, "An ultra-wideband CMOS low-noise amplifier for 3–5-GHz UWB system," *IEEE J. Solid-State Circuits*, vol. 40, no. 2, pp. 544–547, Feb. 2005.
- [103] R. Gharpurey, "A broadband low-noise front-end amplifier for ultra wide-band in 0.13 μ m CMOS," in *Proc. IEEE 2004 Custom Integrated Circuits Conf. (CICC)*, Oct. 3–6, 2004, pp. 605–608.

- [104] Y. Park, C.-H. Lee, J. Cressler, J. Laskar, and A. Joseph, "A very low power SiGe LNA for UWB application," in *2005 IEEE MTT-S Int. Microwave Symp. Dig.*, June 12–17, 2005, pp. 1041–1044.
- [105] H. Knapp, D. Zoschg, T. Meister, K. Aufinger, S. Boguth, and L. Treitinger, "15 GHz wideband amplifier with 2.8 dB noise figure in SiGe bipolar technology," in *IEEE Radio Frequency Integrated Circuits (RFIC) Symp. Dig. Papers*, May 20–22, 2001, pp. 287–290.
- [106] A. Ismail and A. A. Abidi, "A 3–10-GHz low-noise amplifier with wide-band LC-ladder matching network," *IEEE J. Solid-State Circuits*, vol. 39, no. 12, pp. 2259–2268, Dec. 2004.
- [107] A. Bevilacqua and A. M. Niknejad, "An ultrawideband CMOS low-noise amplifier for 3.1–10.6-GHz wireless receivers," *IEEE J. Solid-State Circuits*, vol. 39, no. 12, pp. 2259–2268, Dec. 2004.
- [108] B. Shi and M. Chia, "A 3.1–10.6 GHz RF front-end for multiband UWB wireless receivers," in *IEEE Radio Frequency Integrated Circuits (RFIC) Symp. Dig. Papers*, June 12–14, 2005, pp. 343–346.
- [109] D. K. Shaeffer and T. H. Lee, "A 1.5-V, 1.5-GHz CMOS low noise amplifier," *IEEE J. Solid-State Circuits*, vol. 32, no. 5, pp. 745–759, May 1997.
- [110] N. Shiramizu, T. Masuda, M. Tanabe, and K. Washio, "A 3–10 GHz bandwidth low-noise and low-power amplifier for full-band UWB communications in 0.25- μm SiGe BiCMOS technology," in *IEEE Radio Frequency Integrated Circuits (RFIC) Symp. Dig. Papers*, June 12–14, 2005, pp. 39–42.
- [111] J. Lee and J. Cressler, "A 3–10 GHz SiGe resistive feedback low noise amplifier for UWB applications," in *IEEE Radio Frequency Integrated Circuits (RFIC) Symp. Dig. Papers*, June 12–14, 2005, pp. 545–548.
- [112] M.-C. Chiang, S.-S. Lu, C.-C. Meng, S.-A. Yu, S.-C. Yang, and Y.-J. Chan, "Analysis, design, and optimization of InGaP-GaAs HBT matched-impedance wide-band amplifiers with multiple feedback loops," *IEEE J. Solid-State Circuits*, vol. 37, no. 6, pp. 694–701, June 2002.

- [113] K. Kobayashi and A. Oki, "A DC-10 GHz high gain-low noise GaAs HBT direct-coupled amplifier," *IEEE Microwave Guided Wave Lett.*, vol. 5, no. 9, pp. 308–310, Sept. 1995.
- [114] C.-F. Liao and S.-I. Liu, "A broadband noise-canceling CMOS LNA for 3.1–10.6-GHz UWB receiver," in *Proc. IEEE 2005 Custom Integrated Circuits Conf. (CICC)*, Sept. 18–21, 2005, pp. 160–163.
- [115] F. Bruccoleri, E. A. M. Klumperink, and B. Nauta, "Wide-band CMOS low-noise amplifier exploiting thermal noise canceling," *IEEE J. Solid-State Circuits*, vol. 39, no. 2, pp. 275–282, Feb. 2004.
- [116] F. Bruccoleri, E. Klumperink, and B. Nauta, "Noise cancelling in wide-band CMOS LNAs," in *2002 IEEE Int. Solid-State Circuits Conf. (ISSCC) Dig. Tech. Papers*, vol. 1, San Francisco, CA, Feb. 3–7, 2002, pp. 406–407.
- [117] J. Janssens, J. Crols, and M. Steyaert, "A 10 mW inductorless, broadband CMOS low noise amplifier for 900 MHz wireless communications," in *Proc. IEEE 1998 Custom Integrated Circuits Conf. (CICC)*, May 11–14, 1998, pp. 75–78.
- [118] P. Wambacq, G. G. E. Gielen, P. R. Kinget, and W. Sansen, "High-frequency distortion analysis of analog integrated circuits," *IEEE Trans. Circuits Syst.-I*, vol. 46, no. 3, pp. 335–345, Mar. 1999.
- [119] S. O. Cannizzaro, G. Palumbo, and S. Pennisi, "Distortion analysis of Miller-compensated three-stage amplifiers," *IEEE Trans. Circuits Syst.-I*, vol. 53, no. 5, pp. 961–976, May 2006.
- [120] F.-T. Chien and Y.-J. Chan, "Bandwidth enhancement of transimpedance amplifier by a capacitive-peaking design," *IEEE J. Solid-State Circuits*, vol. 34, no. 8, pp. 1167–1170, Aug. 1999.
- [121] F.-J. Huang and K. O, "A 0.5- μm CMOS T/R switch for 900-MHz wireless applications," *IEEE J. Solid-State Circuits*, vol. 36, no. 3, pp. 486–492, Mar. 2001.
- [122] Z. Li, H. Yoon, F.-J. Huang, and K. K. O, "5.8-GHz CMOS T/R switches with high and low substrate resistance in a 0.18- μm CMOS process," *IEEE Microwave Wireless Compon. Lett.*, vol. 13, no. 1, pp. 1–3, Jan. 2003.

- [123] C. Tinella, J. M. Fournier, D. Belot, and V. Knopik, "A high-performance CMOS-SOI antenna switch for the 2.5–5-GHz band," *IEEE J. Solid-State Circuits*, vol. 38, no. 7, pp. 1279–1283, July 2003.
- [124] N. A. Talwalkar, C. P. Yue, H. Gan, and S. S. Wong, "Integrated CMOS transmit-receive switch using LC-tuned substrate bias for 2.4-GHz and 5.2-GHz applications," *IEEE J. Solid-State Circuits*, vol. 39, no. 6, pp. 863–870, June 2004.
- [125] M.-C. Yeh, Z.-M. Tsai, R.-C. Liu, K.-Y. Lin, Y.-T. Chang, and H. Wang, "Design and analysis for a miniature CMOS SPDT switch using body-floating technique to improve power performance," *IEEE Trans. Microwave Theory Tech.*, vol. 54, no. 1, pp. 31–39, Jan. 2006.
- [126] T. Ohnakado, S. Yamakawa, T. Murakami, A. Furukawa, E. Taniguchi, H. Ueda, N. Suematsu, and T. Oomori, "21.5-dBm power-handling 5-GHz transmit/receive CMOS switch realized by voltage division effect of stacked transistor configuration with depletion-layer-extended transistors (DETs)," *IEEE J. Solid-State Circuits*, vol. 39, no. 4, pp. 577–584, Apr. 2004.
- [127] Z. Li and K. O, "15-GHz fully integrated nMOS switches in a 0.13- μ m CMOS process," *IEEE J. Solid-State Circuits*, vol. 40, no. 11, pp. 2323–2328, Nov. 2005.
- [128] D. A. Johns and K. Martin, *Analog Integrated Circuit Design*. John Wiley & Sons, 1997.
- [129] K. Yamamoto, T. Heima, A. Furukawa, M. Ono, Y. Hashizume, H. Komurasaki, S. Maeda, H. Sato, and N. Kato, "A 2.4-GHz-Band 1.8-V operation single-chip Si-CMOS T/R-MMIC front-end with a low insertion loss switch," *IEEE J. Solid-State Circuits*, vol. 36, no. 8, pp. 1186–1197, Aug. 2001.
- [130] R. Aparicio and A. Hajimiri, "Capacity limits and matching properties of integrated capacitors," *IEEE J. Solid-State Circuits*, vol. 37, no. 3, pp. 384–393, Mar. 2002.

BIBLIOGRAPHY

157

- [131] Y. P. Zhang, Qiang Li, W. Fan, C.-H. Ang, and H. Li, "A differential CMOS T/R switch for multi-standard applications," *IEEE Trans. Circuits Syst.-II: Express Briefs*, vol. 53, no. 8, pp. 782–786, Aug. 2006.

Colophon

This document was typeset with Charter, Palatino, and Helvetica, using $\text{\LaTeX} 2_{\epsilon}$ (pdf $\text{\LaTeX}$ -1.21a) on an Intel x86 PC (NEC desktop of CICS & IBM Thinkpad T43 of mine) with SuSE Linux 9.3 operating system. Text editing was done using Emacs-21.3.1 and AU $\text{\LaTeX}$ -11.82. Major figures were drawn using XCircuit, Xfig, and MATLAB. The $\text{\LaTeX} 2_{\epsilon}$ typesetting system was with the $\text{\TeX}$ -3.0 distribution. The $\text{\LaTeX}$ class file for NTU thesis format, `lqthesis.cls`, was written by myself. The Makefile for $\text{\LaTeX}$ compile was developed by myself based on the version from *Michael Forman*.