Metamaterial based CMOS terahertz integrated circuits

Shang, Yang

2015

Shang, Y. (2015). Metamaterial based CMOS terahertz integrated circuits. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/62937>

<https://doi.org/10.32657/10356/62937>

NANYANG
TECHNOLOGICAL
UNIVERSITY

**METAMATERIAL BASED CMOS TERAHERTZ
INTEGRATED CIRCUITS**

Thesis

Submitted to the School of Electrical and Electronic Engineering
of the Nanyang Technological University

by

SHANG YANG

for the Degree of Doctor of Philosophy

February 24, 2015

Acknowledgement

Firstly, I would like to express my sincerest gratitude to my supervisor Prof. Yu Hao for his countless support and help. His in-depth knowledge and profound view have enlighten me throughout my research time; his praise, criticism and tireless enthusiasm encourage me to deeply explore my research area; his generous and attentive set a good example for the rest of my life.

Secondly, I want to thank School of Electrical and Electronic Engineering and VIRTUS IC Design Centre of Excellence at Nanyang Technological University for giving me an opportunity to commence this degree. Prof. Yeo Kiat Seng, Prof. Liter Siek, Prof. Tan Chuan Seng, Prof. Tan Cher Ming, Dr. Ma Kaixue, Dr. Yan Mei, Dr. Wang Keping, Mr. Nagarajan Mahalingam, Mr. Zhang Shuai and Mr. Ye Wanxin have given me a lot of guidance during the research period; the Laboratory Executive Mr. Lim Weimeng, Ms. Tan Gek Eng, Mr. Sia Liang Poo, Ms. Yang Wanlan, Mr. Seow Yong Hing, Ms. Guee Geok Lian, Mr. Lee Chew Siong, Ms. Gan Siew Kim and Mr. David Robert Neubronner patiently supported me of using the lab facilities; and many other staffs in school kindly helped me. As my wonderful and talented partners in the NTU CMOS Emerging Technology Group, Mr. Fei Wei, Ms. Cai Deyun, Dr. Fu Haipeng, Dr. Zhang Chun, Mr. Liang Yuan, Mr. Li Peng, Mr. Ma shunli, Mr. Yang Chang, Mr. Revanth Nadipalli and many other people have given me a lot of valuable suggestions and supports in the process of my research work; I wish to thank them for all their help, support, interest and valuable hints. Moreover, I appreciate the fruitful comments and measurement supports from Dr. Hu Sanming, Dr. Bi Xiaojun, Dr. Yu Mingbin, Dr. Je Minkyu and Mr. Muthukumaraswamy Annamalai in Institute of Microelectronics, A*STAR (Agency for Science, Technology and Research) in Singapore.

Finally, I should give great appreciation to my parents for their love and endless support. Without their devotion and sacrifice, I could not be where I am today. I would like to give my special thanks to my parents in law, who scarified their retirement leisure time to help take good care of my son so that I can focus on the research work.

Especially, I would like to give a lot special thanks to my loving wife Zhang Rui. Her devoted love makes me stronger to conquer all the challenges.

Contents

Acknowledgement	i
Summary	vii
List of Figures	ix
List of Tables	xvii
1 Introduction	1
1.1 Overview of Terahertz Technology	1
1.1.1 Terahertz Applications	1
1.1.2 Optics based Terahertz System	4
1.1.3 Electronics based Terahertz System	5
1.2 Design challenges in THz CMOS Imaging Systems	6
1.3 Research Objective and Major Contributions	7
1.4 Thesis Organization	11
2 CMOS Metamaterial	13
2.1 Introduction	13
2.2 Non-resonant-type Metamaterial	14
2.2.1 Magnetic Plasmon Waveguide	15
2.2.2 Composite Right/Left Handed T-line	17
2.3 Resonant-type Metamaterial	19
2.3.1 T-line Loaded with Split Ring Resonator	20
2.3.2 T-line Loaded with Complementary Split Ring Resonator	22
2.4 Metamaterial based THz Transceiver	23
3 CMOS THz Modeling	24
3.1 Introduction	24
3.2 Fractional-order T-line Model	26
3.2.1 Fractional Calculus	26

3.2.2	Fractional-order Capacitance and Inductance	27
3.2.3	Fractional-order T-line Model	27
3.2.4	Fractional-order CRLH T-line Model	29
3.3	Model Extraction and Causality Analysis	32
3.3.1	Fractional-order Model Extraction	32
3.3.2	Causal LTI System and Causality Enforcement	33
3.3.3	Causality of T-line Model	35
3.4	Measurements	37
3.4.1	T-line Fractional-order Model Verification	37
3.4.2	CRLH T-line Fractional-order Model Verification	38
3.4.3	Causality Verification and Comparison	43
3.5	Conclusion	46
4	CMOS THz Signal Generation	49
4.1	Introduction	49
4.2	MPW with capacitance loaded coupled-line	51
4.3	Zero-phase CON in Close-looped	54
4.4	60GHz VCO Design by ZPC-Based CON	55
4.4.1	Differential ZPC Unit-cell	55
4.4.2	ZPC-based Oscillator with Tuning	56
4.4.3	60-GHz Zero-phase-coupled Oscillator Network	57
4.4.4	Measurements	58
4.5	140GHz Injection-locked Signal Source by ZPC-based CON	62
4.5.1	Zero-phase Oscillator Unit-cell at 70GHz	63
4.5.2	70GHz Zero-phase Coupled Oscillator Network	64
4.5.3	70GHz to 140GHz Output Frequency Doublers	65
4.5.4	35-GHz to 70-GHz Input Reference Frequency Doubler	66
4.5.5	Measurements	67
4.6	Conclusion	72
5	CMOS THz Signal Detection	73
5.1	Introduction	73
5.2	Oscillators Design with SRR and CSRR	75
5.2.1	Differential TL-SRR Resonator Design	76
5.2.2	Differential TL-CSRR Resonator Design	81
5.2.3	76GHz Oscillator Design with Differential TL-SRR Resonator	84
5.2.4	96GHz Oscillator Design with Differential TL-CSRR Resonator	84

5.2.5	Measurements	87
5.3	High Sensitivity THz Receivers with Resonant-type Metamaterial . . .	93
5.3.1	Fundamentals of Super-regenerative Amplification	93
5.3.2	Super-regenerative Receivers Design by Quench-controlled Os- cillators	96
5.3.3	Measurements	103
5.4	Conclusion	108
6	THz On-chip SIW and LWA Antennas	111
6.1	Introduction	111
6.2	On-chip Antenna Design with SIW	113
6.2.1	Circularly Polarized SIW Antenna with Corner Slots	113
6.2.2	On-chip Implementation in 65-nm CMOS	115
6.2.3	EM Simulation	116
6.3	On-chip Antenna Design with CRLH T-line	117
6.3.1	CRLH T-line Based Leaky Wave Antenna Design	117
6.3.2	280-GHz LWA Design by CRLH T-line	118
6.3.3	EM Simulation	119
6.4	Conclusion	121
7	CMOS THz Transceiver for Imaging	123
7.1	Introduction	123
7.2	135-GHz Narrow Band Imager by DTL-SRR-based SRX	125
7.2.1	THz Imaging by SRA Detection	125
7.2.2	Narrow band imaging at 135 GHz	126
7.3	240 ~ 280-GHz Wide Band Imager with Heterodyne Receiver	130
7.3.1	Architecture and System Specification	130
7.3.2	THz Down-conversion Mixer Design	131
7.3.3	Power Gain Amplifier at 3 GHz	132
7.3.4	Measurements	134
7.3.5	Wide band THz imaging	139
7.4	280-GHz Reflective Imaging System	140
7.4.1	280-GHz Injection-locked CON Signal Source	141
7.4.2	Differential Down-conversion Receiver	150
7.4.3	2D On-chip Leaky Wave Antenna Array	155
7.4.4	Transceiver Integration	156
7.5	Conclusion	156

8 Conclusion and Future Work	161
8.1 Conclusion	161
8.2 Recommendation for Future Work	162
A Author's Publication	164
A.1 Journal	164
A.2 Conference	165
A.3 Book and Book Chapter	167
Bibliography	168

Summary

Terahertz (THz) radiation (0.1-30 THz) fills the gap between electronics and photonics with unique spectroscopic properties. A great deal of attention has been paid to THz imaging system with application of biomedical and security due to the moderate wavelength of THz signal to leverage advantages of both microwave and optics, such as high spatial resolution, good penetration depth to dielectric material or human tissue with no harmful ionization. However, the current optics based THz imaging systems are bulky, expensive, lack of portability with low detection resolution by electro-optic sampling techniques.

With the rapid scaling of CMOS technology, it has become feasible to realize integrated circuits with standard CMOS process in THz regime towards low cost, portable and large-arrayed THz imaging system on a chip. However, it is challenging to deal with the generation, transmission and detection of THz signal by single CMOS transistor due to substrate loss and low device gain with huge path propagation loss. One needs to figure out solutions from all perspectives such as high output power transmitters, high gain antennas, and high sensitivity receivers.

This PhD thesis has explored the use of metamaterial to integrate a number or array of CMOS transistors with significantly improved performance of THz signal generation, detection and transmission. New metamaterial based THz imaging systems have been demonstrated at 140GHz and 280GHz, respectively.

For CMOS THz signal generation, the target is to improve the output power and power efficiency with wide frequency tuning range (FTR) as well as compact size. By coupling N oscillators in-phase, coupled oscillator network (CON) can effectively achieve an N times higher output power but also an N times less phase noise. The conventional on-chip coupling network by length of $\lambda/2$ or λ transmission-line is too bulky and lossy with difficulty for phase synchronization. Non-resonant-type metamaterial such as magnetic plasmon waveguide (MPW) with zero-phase-shift property is applied to achieve in-phase low loss coupling with compact size, which enables the design of signal source with high power density and high efficiency.

For CMOS THz signal detection, the target is to improve the receiver sensitivity with compact size. The use of resonant-type metamaterial: transmission line (T-line) loaded with split ring resonator (TL-SRR) or complementary split ring resonator (TL-CSRR) can significantly improve both high-Q oscillation and oscillatory amplification within compact area. As such, one can achieve low phase noise oscillator for the design of high sensitivity super-regenerative receiver (SRX) with quench control.

For CMOS THz signal transmission, the target is to design wide band, high gain on-chip antennas with compact chip area as well as high efficiency. Substrate integrated waveguide (SIW) has been recently explored for the design of high quality factor (Q) passive devices from mm-wave to THz, which enables an on-chip antenna design that can leverage the advantages of both planar transmission line and non-planar waveguide with lower loss and wide band performance in a miniaturized cavity. Moreover, non-resonant-type metamaterial such as composite-right-left-handed (CRLH) T-line with nonlinear phase-to-length relationships enables more compact antenna design with even higher gain and efficiency.

Finally, with the proposed transmitter and receiver designs, both narrow and wide band THz imager can be demonstrated at 135GHz and 280GHz, respectively. In summary, a metamaterial based CMOS transceiver for THz imaging is proposed with significantly improved performance in THz signal generation, transmission and detection. For THz signal generation, non-resonant-type metamaterial such as MPW can be applied for high power signal source designs; for THz signal transmission, non-resonant-type metamaterial such as CRLH T-line can be applied for the high gain antenna designs; for THz signal detection, resonant-type metamaterial such as DTL-SRR or DTL-CSRR can be applied in the high sensitivity receiver designs. The component designs are supported by chip demonstration with measurement results. The system performance is also evaluated after integration by post-layout simulations.

List of Figures

1.1	Application examples of THz spectroscopy and imaging: (a) non-destructive detection of crack initiation in a film-coated layer on a swelling tablet [1], (b) hydration state characterization in solution [2], (c) in-vitro breast cancer diagnosis [3], and (d) in-vivo skin cancer diagnosis [4]. . .	3
1.2	Application examples of THz communication: (a) high-definition multimedia interface (HDMI) provided by WiGiga and Wireless-HD [5], and (b) THz intra-chip high speed data link between core and memory, and THz inter-chip high speed data link between core and core. . . .	4
1.3	Schematic of a THz spectroscopy and imaging system by electro-optic sampling technique in transmission type	5
1.4	(a) Comparing cut-off frequencies for different FETs [6]; (b) ITRS projected f_{MAX} and f_t of NMOS transistors.	6
1.5	Design challenges of THz imaging system to overcome the huge path. .	7
2.1	Definition of metamaterial.	14
2.2	(a) Equivalent circuit of magnetic plasmon waveguide (MPW); (b) dispersion diagram of MPW.	16
2.3	(a) Equivalent circuit model for the unit-cell of CRLH T-line; (b) dispersion diagram of CRLH T-line	18
2.4	Layout topologies of SRR in (a) and CSRR in (b), and equivalent circuits of SRR in (c) and CSRR in (d).	19
2.5	(a) Standing-wave formed by perfect reflection at DTL-SRR; (b) equivalent circuit of DTL-SRR with condition to form perfect reflection. . .	20
2.6	(a) Standing-wave formed by perfect reflection at DTL-CSRR; (b) equivalent circuit of DTL-CSRR with condition to form perfect reflection.	21

2.7	Metamaterial based THz transceiver design including: (a) high power THz signal source by MPW based zero-phase coupled oscillator network (CON), (b) high gain THz antenna by CRLH T-line, and (c) high sensitivity THz super-regenerative receiver by TL-SRR/TL-CSRR based quench-controlled oscillator.	22
3.1	RLGC unit-cell equivalent circuits of T-line: (a) integer-order model; and (b) fractional-order model.	25
3.2	Equivalent circuits of CRLH unit-cell: (a) integer-order model; and (b) fractional-order model.	30
3.3	Fractional-order T-line modeling parameters extraction flow.	33
3.4	Causality verification and enforcement flow.	35
3.5	T-line testing structure, (a) die photo, and (b) detailed dimensions. . .	37
3.6	Measurement setup of on-wafer S-parameter testing up to 110GHz. . .	38
3.7	Verification of fractional-order T-line model with measurement results for magnitude of S11 and S21 in dB.	39
3.8	Verification of fractional-order T-line model with measurement results for phase delay of S21 and characteristic impedance.	39
3.9	Chip micrograph of fabricated CRLH T-line in 65nm CMOS process. .	40
3.10	Dimension of each unit cell and layer configurations of LWA.	40
3.11	Measurement setup of on-wafer S-parameter testing from 220GHz to 325GHz.	41
3.12	Measurement, EM, integer-order and fractional-order circuits simulation results: magnitude of S21 and S11 in dB.	42
3.13	Measurement, EM, integer-order and fractional-order circuits simulation results: phase of S21 in degree and the absolute value of extracted phase constant β	43
3.14	Measurement, EM, integer-order and fractional-order circuits simulation results: characteristic impedance of CRLH T-line (Z_0).	44
3.15	Causality verification by Hilbert transformation: integer-order T-line model S11.	45
3.16	Causality verification by Hilbert transformation: fractional-order T-line model S11.	45
3.17	Causality verification by Hilbert transformation: error magnitude comparison of S11	46

3.18	Causality verification by Hilbert transformation: integer-order T-line model S21.	47
3.19	Causality verification by Hilbert transformation: fractional-order T-line model S21.	47
3.20	Causality verification by Hilbert transformation: error magnitude comparison of S21	48
4.1	(a) Equivalent circuit of differential ZPC loaded with parasitic capacitance (cross-coupled pairs); (b) on-chip realization of inter-digital coupled T-lines.	51
4.2	The calculated l for zero coupling phase vs. loaded capacitance and even-mode characteristic admittance at 60GHz.	53
4.3	Closed loop coupled oscillator network with center combined output. .	54
4.4	Layout of inter-digital differential ZPC with effective electrical length controlled by MOS switches.	56
4.5	Simulated magnitude of S11 and S21 in dB and phase of S21 of proposed the proposed coupler structure with comparison of S21 magnitude to the conventional one by coupled T-line.	57
4.6	The extracted dispersion diagram and relative permittivity (ϵ_r).	58
4.7	Schematic of proposed VCO with differential outputs combined at the center.	59
4.8	Die micrograph of the fabricated 60GHz VCO chip in 65nm CMOS. . .	60
4.9	Measured and simulated output power and power efficiency of proposed VCO over entire 60GHz band.	61
4.10	Measured and simulated VCO FTR under various switches configurations to cover 58.3-64.8GHz continuously.	61
4.11	Measured and simulated VCO phase noise at 60GHz.	62
4.12	Block diagram of proposed 140GHz signal source with center combined output from four 70GHz zero-phase-coupled oscillator unit-cells.	64
4.13	Schematic of on-chip ZPC-based oscillator unit-cell at 70-GHz band with inter-digital coupled T-line and switch-controlled inductive loadings. .	65
4.14	EM simulation results of 70-GHz ZPC with inter-digital coupled T-line. .	66
4.15	Extracted dispersion diagram of 70 GHz ZPC with inter-digital coupled T-line.	67
4.16	Schematic of injection-locked 70-GHz CON with 4 oscillator unit-cells. .	67

4.17 Schematic of four 70-GHz to 140-GHz push-push frequency doublers with center combined output.	68
4.18 Schematic of 35GHz to 70GHz frequency doubler with inter-digital Marchand Balun at 70 GHz.	68
4.19 EM simulation results of the proposed Marchant balun at 70GHz. . . .	69
4.20 Post-layout simulation results of the 35-GHz to 70-GHz frequency doubler.	70
4.21 Die micrograph of the fabricated THz source in CMOS.	70
4.22 Measured and simulated output power of signal source with three different inductive loading modes.	71
4.23 Measured and simulated output spectrum of the proposed signal source at 132GHz.	72
5.1 CMOS THz imager array.	74
5.2 (a) Block diagram of super-regenerative receiver; (b) impact of resonator Q-factor to receiver sensitivity.	75
5.3 (a) Stacked SRR unit-cell designed by metal layers of M7 and M6; (b) S21 simulation results with different stacking methods.	76
5.4 Geometries of resonators with slow-wave shielding: (a) differential T-line loaded with stacked SRR, (b) T-line based standing-wave resonator, and (c) cross-section of BEOL.	77
5.5 T-line based SRR excitation: (a) single-ended approach, (b) differential approach, (c) magnetic field distribution of single-ended approach, and (d) magnetic field distribution of differential approach.	79
5.6 Simulated Γ plot in smith chart of SRR/T-line unit cell at resonance. .	80
5.7 Simulated Γ and quality factor (Q) of SRR/T-line unit cell at resonance.	81
5.8 On-chip differential T-line loaded with CSRR.	82
5.9 EM characterization of the proposed differential CSRR resonator. . . .	83
5.10 Voltage distribution of the DTL-SRR based resonator.	83
5.11 Schematic of the DTL-SRR based oscillator.	85
5.12 Circuit diagram of the 96GHz CMOS oscillator with the use of the proposed metamaterial resonator by CSRR structure.	86
5.13 Equipment setup for W-band 75~110GHz phase noise measurement. . .	87
5.14 Chip photo of the DTL-SRR based oscillator and SWO.	88
5.15 Spectrum of the 76GHz DTL-SRR oscillator.	89
5.16 Spectrum of the 76GHz SWO.	89

5.17	Phase noise measurement and simulation results of the 76GHz DTL-SRR oscillator and SWO.	90
5.18	Chip photo of the DTL-CSRR based oscillator.	91
5.19	Spectrum of the 96-GHz oscillator with DTL-CSRR.	92
5.20	Phase noise measurement and simulation results of 96-GHz oscillator with DTL-CSRR.	93
5.21	Simplified equivalent circuit model of super-regenerative amplifier. . . .	94
5.22	Reflection coefficient of T-line loaded with CSRR unit-cells.	97
5.23	Reflection loss compensation by cross-coupled NMOS pair with controlled tail current.	98
5.24	Layout for CMOS on-chip implementation of DTL-CSRR for 96GHz SRX.	99
5.25	EM-simulation based comparison of DTL-CSRR and LC-tank resonator for CMOS 96GHz SRX design.	100
5.26	Schematic of CMOS 96GHz SRX with DTL-CSRR.	101
5.27	Layout for CMOS on-chip implementation of DTL-SRR for 135GHz SRX. . . .	102
5.28	EM-simulation based comparison of DTL-SRR and LC-tank resonator for CMOS 135GHz SRX design.	102
5.29	Impedance Diagram of DTL-SRR and LC-Tank in Globalfoundries 65-nm CMOS process, (a) real and imaginary parts of Z_{Diff} , (b) phase of Z_{Diff} , (c) phase of Z_{Diff} when the ideal 16-fF capacitor is included. . .	104
5.30	Schematic of CMOS 135GHz SRX with DTL-SRR.	105
5.31	Die micrographs: (a) CMOS 96GHz SRX with DTL-CSRR, and (b) CMOS 135GHz SRX with DTL-SRR.	105
5.32	Measurement and simulation results of CMOS 96GHz SRX, (a) gain and input S11, and (b) output voltage (V_{out}) and responsivity vs. input power (P_{in}).	107
5.33	Measurement setup of CMOS 135GHz SRX with DTL-SRR.	108
5.34	Measurement and simulation results of CMOS 135GHz SRX, (a) gain, and (b) output voltage (V_{out}) and responsivity vs. input power (P_{in}). .	109
6.1	Geometrical configuration of the proposed SIW antenna with four corner slots	113
6.2	Design of on-chip integrated circular-polarized SIW antenna in CMOS 65nm process.	116

6.3	HFSS Simulation results of proposed SIW antenna directivity at 270 GHz.	117
6.4	HFSS simulation results of proposed SIW antenna radiation pattern at 270 GHz.	118
6.5	HFSS simulation results of polarized vector electric field in the slots of SIW cavity at 270 GHz.	119
6.6	HFSS simulation results of input S11 and antenna axial ratio on broadside radiation direction.	120
6.7	Operation diagram of leaky wave antenna: (a) $\beta > 0$, (b) $\beta = 0$, and (c) $\beta < 0$	120
6.8	Stacking of high resistivity Si layer on-top of the LWA.	121
6.9	Radiation efficiency enhancement by stacking high resistivity Si layer. .	121
6.10	Gain radiation patterns for the proposed antenna at three frequencies: $f=250\text{GHz}$ ($\beta < 0$, backward radiation), $f=280\text{GHz}$ ($\beta = 0$, broadside radiation), and $f = 290\text{GHz}$ ($\beta > 0$, forward radiation).	122
7.1	Metamaterial based THz imaging system	124
7.2	THz image system with heterodyne receiver for high spectrum resolution detection and circular-polarized antenna for the tolerance of depolarization effect.	124
7.3	(a) PCB integration of CMOS 135-GHz SRX with antenna; (b) THz imaging measurement setup with the proposed receiver chip integrated on PCB and object under test fixed on an X-Y moving stage.	127
7.4	Images captured by imaging system with the proposed 135-GHz SRX receiver: knife, perfume, and coin in hand-bag.	128
7.5	Images captured by imaging system with the proposed 135-GHz SRX receiver: moisturized and normal Panadol pills.	128
7.6	Images captured by imaging system with the proposed 135-GHz SRX receiver: various types of oil.	129
7.7	Absorption ratio of various types of oil detected at 135 GHz.	129
7.8	Schematic of THz down-conversion mixer at 280GHz.	131
7.9	Simulation results of proposed mixer: (a) S-parameters and conversion gain when sweeping RF and LO frequencies with $F_{LO} = F_{RF} + 3\text{GHz}$; (b) conversion gain at different LO power level when sweeping RF frequency with $F_{IF} = F_{RF} - 280\text{GHz}$	133
7.10	Schematic of the three-stage power gain amplifier and the output buffer.	134

7.11	Simulation results of the three-stage power gain amplifier with output buffer.	134
7.12	Chip micro photograph of the proposed CMOS 280-GHz heterodyne receiver with on-chip integrated circular polarized SIW antenna.	135
7.13	Equipment setup for 280-GHz receiver measurement.	135
7.14	Gain and sensitivity measurement results when sweeping RF and LO frequencies with $F_{LO} = F_{RF} + 3GHz$	136
7.15	Gain measurement results when sweeping RF and LO frequencies with $F_{IF} = F_{LO} + 280GHz$	138
7.16	Receiver sensitivity at 250 GHz verse receiver resolution bandwidth. . .	139
7.17	Measurement setup of THz image system.	140
7.18	Captured THz image results of Panadol pills and skin samples under 240 GHz and 280 GHz radiation.	141
7.19	The block diagram of CMOS based THz reflective imaging system. . .	142
7.20	Block diagram of proposed 280GHz signal source.	143
7.21	Schematic of on-chip MPW base oscillator unit-cell at 140-GHz band with inter-digital coupled T-line.	144
7.22	EM simulation results of 140-GHz on-chip MPW unit-cell with inter-digital coupled T-line.	144
7.23	(a) Schematic of injection-locked 140GHz CON with 4 MPW unit-cells; (b) 2nd harmonic outputs power combining network.	145
7.24	Schematic of injection-locked 70GHz CON with 2 MPW unit-cells. . . .	146
7.25	140GHz power splitter from one single-ended input to two differential outputs: (a) On-chip layout, (b) EM-simulation results.	147
7.26	Input reference frequency quadrupler from 17.5 GHz to 70 GHz. . . .	148
7.27	Post-layout simulation results of 17.5 GHz to 70 GHz quadrupler. . . .	149
7.28	Cadence layout of the proposed 280GHz source in CMOS.	150
7.29	Simulated output power of the proposed 280GHz source in CMOS. . . .	151
7.30	Schematic of proposed differential down-conversion mixer with a input network of 90° high- Z_0 T-line.	152
7.31	Post-layout simulation results of the high- Z_0 T-line network with 90° phase delay.	153
7.32	Post-layout simulation results: (a) the conversion gain and output phase mismatch of the 280-GHz differential mixer; (b) the conversion gain and NF of the 280-GHz receiver with differential mixer and VGA from [7]. .	154
7.33	2D LWA array with 2×13 unit-cells.	155

7.34	HFSS simulated radiation pattern of the 2D LWA array at 280 GHz in: (a) 3D plot, and (b) polar plots in ZOX and ZOY planes.	159
7.35	Simulated antenna input S11 and gain at broadside direction (Z-axis). .	160
7.36	Cadence layout of the proposed 280GHz transceiver in CMOS	160

List of Tables

1.1	Comparison of general spectroscopy and imaging technologies	2
3.1	Modeling parameters of integer-order and fractional-order RLGC model for T-line	40
3.2	Modeling parameters of integer-order and fractional-order models of CRLH T-line unit-cell	41
4.1	Performance comparison of state-of-the-art VCO designs around 60 GHz	63
4.2	Performance comparison with recently published THz signal source . .	69
5.1	Performance comparison of state-of-the-art oscillator designs around 70 GHz	90
5.2	Performance comparison of state-of-the-art oscillator designs around 100 GHz	92
5.3	Performance comparison of state-of-the-art receivers for imaging appli- cation	110
7.1	State-of-the-art CMOS THz image receivers performance comparison .	138
7.2	Performance comparison with recently published THz signal sources . .	149
7.3	Performance comparison with recently published THz transmitters and receivers	157

Chapter 1

Introduction

1.1 Overview of Terahertz Technology

The terahertz (THz) radiation ($0.1 \sim 30$ THz) is categorized between millimeter-wave (mm-wave) and infrared light wave [8]. Recently, a great deal of attention has been paid to the THz spectroscopy and imaging system due to the moderate wavelength of THz wave that can leverage the advantages of both millimeter-waves (mm-waves) and light waves [8–10]. Like mm-wave, THz wave has deep penetration to dielectric substances such as ceramics, plastics, powders and food; like light wave, THz images with high spatial resolution can be obtained by a 2-dimensional (2D) detection with THz sensor array [11]. On the other hand, high-data-rate THz wireless communications have also come into view due to the abundance of undeveloped bandwidth resources [12].

1.1.1 Terahertz Applications

1.1.1.1 Terahertz Spectroscopy and Imaging

Spectroscopy is a very mature method to identify substance composition by fingerprints analysis in physical and analytical chemistry. A comparison of general spectroscopy and imaging technologies is shown in TABLE 1.1. The commonly used spectroscopy methods are fourier transform infrared spectroscopy (FTIR), non-dispersive infrared spectroscopy (NDIR), Raman spectroscopy, and X-ray spectroscopy. THz spectroscopy does not only show the unique spectral fingerprints for many substances [13], but also has several distinct advantages when compared to these conventional spectroscopy methods. Firstly, unlike X-ray, THz radiation is safe to the tissue under test as well as the people who are conducting the measurement due to its longer

wavelength with non-ionizing nature [8]. Secondly, many materials that cannot be penetrated by infrared are transparent to THz, which enables a non-destructive analysis to the coated substance. For example, THz based inner layer reflectance analysis can be applied in the thin-film coating analysis of drug tablets as shown in Fig. 1.1(a) [1]. Thirdly, THz radiation is highly sensitivity to the water hydration state, which can be utilized in the concentration analysis of disaccharide water solutions as shown in Fig. 1.1(b) [2]. Finally, THz radiation is more sensitive to the vibration and interaction of molecules such as protein or polymer. Therefore, it can be utilized for the in-vitro cancer diagnosis (Fig. 1.1(c)) [3] as well as explosive detection [14].

TABLE 1.1: COMPARISON OF GENERAL SPECTROSCOPY AND IMAGING TECHNOLOGIES

Specifications	Ultrasound	MRI	THz	FTIR, NDIR, Ramen	CM, OCT	X-rays
Applications	Imaging	Imaging	Spectroscopy & Imaging	Spectroscopy	Imaging	Spectroscopy & Imaging
Radiation	Mechanical wave	Magnetic field	EM wave (Non- ionizing)	EM wave (Non- ionizing)	EM wave (Non- ionizing)	EM wave (Ionizing)
Frequency (Hz)	10^6 - 10^7	10^7 - 10^8	10^{11} - 10^{13}	10^{13} - 10^{15}	$\sim 10^{15}$	10^{16} - 10^{19}
Molecular interactions	No	No	Yes	No	No	No
Image reso- lution	~ 2 mm	~ 1 mm	$\sim 200\mu\text{m}$	-	0.1 - $10\mu\text{m}$	~ 15 nm
Penetration depth to human body	200-300mm	>1 m	1-3mm	0.5 - $5\mu\text{m}$	0.2 - $0.8\mu\text{m}$	>1 m

THz imaging also has several distinct advantages when compared to the other imaging techniques, such as ultrasound scan, magnetic resonance imaging (MRI), confocal microscopy (CM) and optical coherence tomography (OCT). Firstly, THz imaging has higher resolution than ultrasound scan or MRI due to its much shorter wavelength. It is also more sensitivity to the thin tissues due to a stronger reflection and attenuation in the water content. Secondly, compared to the existing optics based imaging methods such as CM and OCT, even though THz imaging system has lower resolution, it has much higher penetration depth due to its much longer wavelength. Recently, with remarkable contrast in skin and breast cancer demonstrated in THz images as shown in Fig. 1.1(d) [3, 4]. THz imaging system has been used as an intra-operative tool during breast cancer surgery in Guys hospital in London [15].

Figure 1.1: Application examples of THz spectroscopy and imaging: (a) non-destructive detection of crack initiation in a film-coated layer on a swelling tablet [1], (b) hydration state characterization in solution [2], (c) in-vitro breast cancer diagnosis [3], and (d) in-vivo skin cancer diagnosis [4].

1.1.1.2 Terahertz High-data-rate Wireless Communication

Even since the invention of radio in late 1800s, the pace of development of wireless communication has never stopped. In the past half century, the carrier frequency of communication system has rapidly increased from several megahertz (MHz) into multi-gigahertz (GHz) ranges to satisfy the growing bandwidth requirement. The recently developed 60-GHz systems with 5 ~ 9-GHz license-free band are able to provide a transfer speed up to 10 Gb/s for short distance data communication [5]. In order to further enhance the data transfer speed to multiple tens or hundreds of Gb/s for various applications such as ultra-high definition TV in the near future, we have to development the communication systems in THz regime with abundant bandwidth

resources. The application of THz communication systems can be mainly categorized into in-door data-links and system level data-transfer as shown in Fig. 1.2.

For the in-door data-links application, one THz wireless data transmission was initially demonstrated in 2009 at 300GHz with a photonics-based transmitter, this system is able to achieve a data rate of 12.5 Gb/s over 0.5-m distance [16]. The lab scale communication was also demonstrated at 625 GHz with a data rate of 2.5Gb/s in 2011 [17]. Most recent developments of semiconductor technologies have demonstrated a very clear potential of higher level integration with wireless I/Os for inter-chip or intra-chip communication [18], which is very likely to be achieved in THz. Recently, an integrated millimeter-wave integrated circuits (MMICs) THz transceiver has been demonstrated in 50nm mHEMT technology with a data rate of 25 Gb/s at 220 GHz [19]. Potentially, it is very promising for inter-chip or intra-chip communication in THz regime with high data rate and energy efficiency [20, 21].

Figure 1.2: Application examples of THz communication: (a) high-definition multimedia interface (HDMI) provided by WiGiga and Wireless-HD [5], and (b) THz intra-chip high speed data link between core and memory, and THz inter-chip high speed data link between core and core.

1.1.2 Optics based Terahertz System

The current optics based THz imaging system is developed by the well known electro-optic sampling technique [22] as shown in Fig. 1.3. When an ultra-short optical pulse (~ 50 femtoseconds) illuminates a non-linear semiconductor material such as zinc telluride (ZnTe), a very short electric pulse is generated at the input of the dipole antenna with THz power spectrum. The average power level of the THz signal generated in this way is in the order of nanowatts, The resulting THz pulse usually has a very wide bandwidth, of which the upper and lower frequency boundaries are

determined by the charge carriers acceleration in the semiconductor material and the antenna cut-off frequency, respectively. After penetrating through the sample under test, the resulting THz radiation is coherently detected in the time domain with both intensity and phase information, which can be further utilized in the non-destructive testing as well as 3D imaging. However, there are several drawbacks in the optics based time domain THz spectroscopy and imaging system. Firstly, the optics-based THz system is usually bulky, expensive and lack of portability. For example, Ti:Sapphire laser source is usually required for femtosecond pulse generation, and lots of mirrors and lens are needed for optical path adjustment. Secondly, the detection resolution and efficiency are limited. The absorption and reflection properties of tissues are usually spectral specific. As such, low spectrum resolution in frequency domain is resulted.

Figure 1.3: Schematic of a THz spectroscopy and imaging system by electro-optic sampling technique in transmission type

1.1.3 Electronics based Terahertz System

The THz imaging system can be potentially implemented by electronic approaches. With the rapid scaling of CMOS technology, it has become feasible to realize integrated circuits with standard CMOS process in THz regime towards low cost, portable and large-arrayed THz imaging system on a chip. Recently, several CMOS based transmitting and receiving components have been developed in THz [23–25]. As shown in 1.4(a) from [6], when the size of transistors is scaling down, the gap between CMOS transistors and three five group transistors is getting smaller and smaller. The ITRS (International Technology Roadmap for Semiconductors) projected roadmap of ft and

f_{max} for NMOS transistors is shown in Fig. 1.4(b). By the year 2020, the f_{max} of CMOS transistors will be higher than 1THz. Compared to the other semiconductor fabrication processes like SiGe, InP or GaAs, standard CMOS technology is always considered as the most cost effective solution to combine all digital, analog and RF system design. Moreover, CMOS based transmitter and receiver designs provide much higher flexibility in the power amplification and frequency control. As such, the THz imaging can be directly performed in frequency domain with improved detection resolution and efficiency. Therefore, it is of great interest to develop CMOS based THz imaging system with high resolution, low cost and high portability.

Figure 1.4: (a) Comparing cut-off frequencies for different FETs [6]; (b) ITRS projected f_{MAX} and f_t of NMOS transistors.

1.2 Design challenges in THz CMOS Imaging Systems

As shown in Fig. 1.5, THz waves suffers great loss when propagating inside free space. As a result, the major challenges in the CMOS imaging system design are to deal with the generation, transmission and detection of THz signals. Take 300-GHz signal for example, assuming the power generated by the signal source is -15 dBm and it suffers a free space path loss about 80-90 dB for 1 meter distance, if the antennas on both transmitter and receiver side can both provide 20-dB gain, the sensitivity of receiver must be better than -85 dBm to detect the signal. Usually in

Figure 1.5: Design challenges of THz imaging system to overcome the huge path.

a standard CMOS process, the transmitting power is limited by the maximum source drain voltage and the current, the receiver sensitivity is limited by the noise figure and bandwidth of receiver frontend, the gain and radiation efficiency of antennas are limited by the loss of metal and substrate and antenna size, respectively. In order to overcome the above difficulties and design a high performance THz imaging system in CMOS process, the design of high-Q passive structure is required in every part of the imaging system to replace the conventional transmission lines (T-line) or LC-tank resonators, which usually suffer from large size and low quality factor in THz and greatly limit the system performance.

1.3 Research Objective and Major Contributions

The main contribution of this work is the proposed metamaterial based CMOS transceivers for THz imaging and spectroscopy applications with significantly improved performance in THz signal generation, transmission and detection. As explored in this PhD thesis, on-chip metamaterial structures are demonstrated with several advantages over the conventional approaches based on T-lines or LC-Tanks. Firstly, non-resonant-type on-chip metamaterials, such as magnetic plasmon waveguide (MPW) or composite right/left handed T-line (CRLH-TL), can create a low loss medium with zero-phase propagation; secondly, resonant-type on-chip metamaterials,

such as T-line loaded with split ring resonator (TL-SRR) and complementary split ring resonator (TL-CSRR), can effectively increase the energy storage within a very compact area and resulting a much higher Q-factor. These unique features can be potentially utilized to resolve the fundamental limitations in the design of high performance THz imaging system. For examples, MPW can be applied in high power signal source designs for THz signal generation; CRLH T-line can be applied in the high gain antenna designs for THz signal transmission; DTL-SRR or DTL-CSRR can be applied in the high sensitivity receiver designs for THz signal detection.

For CMOS THz signal generation, the target is to improve the output power and power efficiency with wide frequency tuning range (FTR) as well as compact size. By coupling N oscillators in-phase, coupled oscillator network (CON) can effectively achieve an N times higher output power but also an N times less phase noise. The conventional on-chip coupling network by length of $\lambda/2$ or λ transmission-line is too bulky and lossy with difficulty for phase synchronization. Non-resonant-type meta-material such as magnetic plasmon waveguide (MPW) with zero-phase-shift property is applied to achieve in-phase low loss coupling with compact size, which enables the design of signal source with high power density and high efficiency. Particularly, this PhD thesis has summarized the following works

- One high output power and low phase noise 60GHz VCO is demonstrated by zero-phase-shift coupled oscillator network (CON). Each oscillator unit-cell is implemented by a MPW based zero-phase-shift coupler (ZPC), which has minimized loss for in-phase coupling condition when connecting with other oscillator unit-cells. The frequency tuning range is improved by inductive tuning at each unit-cell as well. A ZPC-based CON with 4 unit-cells is implemented at 60 GHz in 65-nm CMOS process with a compact core chip area of 0.11 mm². The measured results are: 2-mW peak output power, 2.2% power efficiency, -116.7dBc/Hz phase noise at 10 MHz offset and 15.8% frequency tuning range (FTR) centered at 63.1 GHz.
- One high output power and high efficiency injection-locked signal source is demonstrated by ZPC-based CON at 140 GHz. Each oscillator unit-cell is designed by an inter-digital ZPC with 0.4-dB loss at 70 GHz, and is further doubled to 140 GHz by a push-push frequency doubler. With four in-phase coupled unit-cells, high output power 140-GHz signal is generated at the center of the CON by an in-phase power combination of the second harmonics generated by push-push frequency doublers. Moreover, a 9.7% FTR centered at 133.5 GHz is achieved

by inductive tuning at each oscillator unit-cell. The proposed THz signal source is fabricated in 65-nm CMOS process with compact core chip area of 0.13 mm². The measured results are: 3.5-mW peak output power, 2.4% power efficiency and 26.9-mW/mm² power density.

For CMOS THz signal detection, the target is to improve the receiver sensitivity with compact size. The use of resonant-type metamaterial: transmission line (T-line) loaded with split ring resonator (TL-SRR) or complementary split ring resonator (TL-CSRR) can significantly improve both high-Q oscillation and oscillatory amplification within compact area. As such, one can achieve low phase noise oscillator for the design of high sensitivity super-regenerative receiver (SRX) with quench control. Particularly, this PhD thesis has summarized the following works

- Two low phase noise oscillators are designed with on-chip high-Q metamaterial based resonators: T-line loaded with split ring resonator (TL-SRR) and complementary split ring resonator (TL-CSRR) at 76 GHz and 96 GHz, respectively. By creating a sharp stop-band at the resonance frequency of TL-SRR or TL-CSRR, a standing wave is formed by the incident and reflected EM-wave with low loss and high power density, which results in a high-Q resonator. The 76-GHz TL-SRR based oscillator is measured with 2.7-mW core power consumption, -108.8-dBc/Hz phase noise at 10-MHz offset, and -182.1-dBc/Hz figure-of-merit (FOM), which is 4 dB better than that of a 76-GHz standing-wave oscillator implemented on the same chip. The 96-GHz TL-CSRR based oscillator is measured with 7.5-mW core power consumption, -111.5-dBc/Hz phase noise at 10-MHz offset and -182.4-dBc/Hz FOM, which is at least 6-dB better than the existing oscillators with LC-tank based resonators.
- Two super-regenerative receivers with quench-controlled metamaterial high-Q oscillators by TL-CSRR and TL-SRR are demonstrated with improved sensitivity over traditional LC-tank resonator based designs at 96 GHz and 135 GHz, respectively. With sharp stop-band introduced by the metamaterial resonators, high-Q oscillatory amplifications are achieved. The 96-GHz DTL-CSRR based SRX has a compact core chip area of 0.014 mm², and it is measured with power consumption of 2.8mW, sensitivity of -79dBm, noise figure (NF) of 8.5dB, and noise equivalent power (NEP) of 0.67 fW/ $\sqrt{Hz}$. The 135-GHz DTL-SRR based SRX has a compact core chip area of 0.0085 mm², and it is measured with power consumption of 6.2 mW, sensitivity of -76.8 dBm, NF of 9.7 dB, and NEP of 0.9 fW/ $\sqrt{Hz}$. The proposed SRXs have 2.8-4-dB sensitivity improvement and 60%

smaller core chip area when compared to the conventional SRX with LC-tank based resonator at similar frequencies. The 135-GHz SRX is also demonstrated for imaging applications after integrated with antenna.

For CMOS THz signal transmission, the target is to design wide band, high gain on-chip antennas with compact chip area as well as high efficiency. Substrate integrated waveguide (SIW) has been recently explored for the design of high quality factor (Q) passive devices from mm-wave to THz, which enables an on-chip antenna design that can leverage the advantages of both planar transmission line and non-planar waveguide with lower loss and wide band performance in a miniaturized cavity. Moreover, non-resonant-type metamaterial such as composite-right-left-handed (CRLH) T-line with nonlinear phase-to-length relationships enables more compact antenna design with even higher gain and efficiency. Particularly, this PhD thesis has summarized the following works

- One wide band 280-GHz on-chip circularly polarized SIW antenna is designed in CMOS process with compact area. By creating corner slots in SIW structure, the size of the proposed SIW antenna is reduced by 15% when compared to the conventional designs. As verified by the EM simulation, the proposed antenna has -0.5-dBi antenna gain and 32.1-GHz bandwidth centered at 268 GHz.
- One 280-GHz high gain on-chip 1D LWA with broadside radiation is demonstrated by 13-unit-cell periodic CRLH T-line in 65nm CMOS. The antenna radiation efficiency is improved by 12 times by stacking high-resistivity dielectric layer at 280 GHz. With the correlated measurement and EM validation from 220 to 325 GHz, the maximum radiation efficiency of 65%, and an antenna gain of 4.1 dBi is achieved at 280 GHz.

Finally, with the proposed transmitter and receiver designs, both narrow and wide band THz imager can be demonstrated at 135 GHz and 280 GHz, respectively. Particularly, this PhD thesis has summarized the following works

- One narrow band transmission-type THz imager is demonstrated at 135 GHz by integrating DTL-SRR based SRX with off-chip patch antenna array. Various pharmacy and security applications are demonstrated by the proposed imager as well as the sample transmission property characterization.
- One wide band transmission-type THz imager is demonstrated at 239 ~ 281 GHz with a high spectrum resolution and high sensitivity direct-conversion receiver

in 65nm CMOS process. The receiver consists of a circular-polarized substrate integrated waveguide (SIW) antenna, a single-gate down-conversion mixer and a power gain amplifier. The proposed receiver is measured with -2dB conversion gain, 42GHz bandwidth, -54.4-dBm sensitivity at 100-MHz detection resolution bandwidth, 6.6-mW power consumption and 0.99-mm² chip area. Moreover, frequency dependent biomedical imaging applications are demonstrated by the proposed imager.

- One wide band reflection-type THz imager is proposed at 280 GHz with integrated transceiver in 65-nm CMOS process, which consists of a high power CON based 280-GHz signal source, an 2D on-chip high-gain LWA array and a differential down-conversion receiver. The 280-GHz signal source is designed by connecting the 2nd harmonic outputs of two 140GHz CONs in parallel, which are both injection locked with the same phase and magnitude. It is simulated with an output power of +2.8 dBm, a power efficiency of 0.66%, and a FTR of 10.5% from 272 GHz to 302 GHz. The 2D on-chip high-gain LWA array is designed by connecting two 1D LWAs in parallel with 2×13 unit-cells. The proposed 2D LWA array is simulated with a broadside radiation pattern with 9.1dBi directivity and 41% radiation efficiency at 280 GHz. The differential down-conversion receiver is designed by integrating a differential single-gate mixer with one modified Cherry-Hooper amplifier based variable-gain amplifier with compact size. A bidirectional hybrid coupler with high impedance T-line is proposed to provide 90° phase shift for both RF and LO signals at the mixer input as well as to bypass the LO signal. The proposed receiver is simulated with a conversion gain of 46.6 dB, and a NF of 24.6 dB at 280 GHz. The whole transceiver has a compact size of 1 mm², and consumes 298.6 mW power operating under 1.2-V power supply. The transmitter is simulated with an equivalent isotropically radiated power (EIRP) power of 6.3 dBm, an EIRP density of 4.27 mw/mm²; the receiver is simulated with a maximum gain of 51dB and a sensitivity of -57.6 dBm.

1.4 Thesis Organization

The rest of the thesis is organized as follows. Chapter 2 briefly reviews the background of metamaterial and its applications in the THz imaging system. Chapter 3 explores the accurate modeling of THz devices in fractional-order with measurement

results. Chapter 4 discusses the design of THz CMOS signal sources with measurement results. Chapter 5 discusses the design of THz CMOS detectors with measurement results. Chapter 6 discusses the design of THz CMOS on-chip antennas with verification in EM-simulation and correlated measurement results. Chapter 7 demonstrates the CMOS transceiver design for both transmission and reflection types THz imaging. The thesis is concluded in Chapter 8 followed by the recommendation of future works.

Chapter 2

CMOS Metamaterial

2.1 Introduction

Metamaterial was first demonstrated by [26] with the use of split ring resonator (SRR) and metallic wire, among which SRR and metallic wire show the properties of negative permeability (μ) and negative permittivity (ε) at the resonance frequency, respectively. Metamaterial is not a traditionally defined material. It composes of many periodic or non-periodic unit cells. By giving different structure and property to these unit cells, the whole array of unit cells, which is the metamaterial, would show some properties that do not naturally exist. A more clear definition can be found in Fig. 2.1 [27], where both x and y axis are corresponding the the material relative permittivity (ε_r) and permeability (μ_r), respectively. Most natural materials lies on the horizontal line in the 1st quadrant ($\varepsilon_r > 0, \mu_r > 0$) with a relative permittivity larger than 1 and a nearly unity relative permeability. But with metamaterial, by giving different design for unit cells, theoretically we can construct a material located in any regions in Fig. 2.1 that enables many interesting applications. According to the transmission and reflection property, metamaterial can be categorized into two types: non-resonant-type and resonant-type.

Metamaterial in 1st and 3rd quadrants are transmission-types, where EM wave is able to propagate inside. The EM wave that propagate in the 1st quadrant ($\varepsilon_r > 0, \mu_r > 0$) has a positive phase velocity, which is a linear function of frequency. But when metamaterial appears in the 3rd quadrant ($\varepsilon_r < 0, \mu_r < 0$), it is called left-handed material. Left-handed material has a non-linear negative phase velocity, which means when the energy propagates forward in this material, the phase actually propagates backward. One application of left-handed material to composite with right-handed material to provide zero phase propagation. Note that transmission-type

Figure 2.1: Definition of metamaterial.

metamaterial with zero phase propagation is located on the X or Y Axes with $\epsilon_r = 0$ or $\mu_r = 0$.

Metamaterial in 2nd and 4th quadrants are resonant-types, where wave propagation is prohibited. When metamaterial appears in the 2nd quadrant ($\epsilon_r < 0, \mu_r > 0$), it is called electric plasma. Evanescent wave will also be formed to strongly reflect the EM wave entering it. Such medium can be achieved by coupled right-handed material with a negative ϵ structure such as complementary split ring resonator (CSRR). When metamaterial appears in the 4th quadrant ($\epsilon_r > 0, \mu_r < 0$), it is called magnetic plasma. Evanescent wave will be formed to strongly reflect the EM wave entering it. Such medium can be achieved by coupled right-handed material with a negative μ structure such as SRR.

2.2 Non-resonant-type Metamaterial

Non-resonant-type metamaterial such as magnetic plamson waveguide (MPW) or composite right/left handed (CRLH) T-line can be applied to achieve zero-phase-shifter with compact size. Magnetic coupled resonators have been extensively explored in the mm-wave filter designs since 1958 [28, 29]. But not until 2002, was MPW firstly

proposed as a transmission medium by E. Shamonina and his colleagues [30]. Recent years, MPW is also applied in the middle distance wireless energy transmission systems [31, 32]. On the other hand, the applications of CRLH T-line has also been widely explored in PCB scale for the designs of various passive devices below 10GHz such as antennas, filters, hybrid couplers, baluns and power combiners [33–38]. In addition, it can also be applied in the design of active devices such as distributed power amplifiers [39, 40]. Both MPW and CRLH T-line provide much more flexible phase control to the RF circuit with positive, negative or zero phase propagation modes, and their nonlinear phase-to-length relationships also enable much more compact designs compared to the ones with conventional T-line. When the operating frequency is further increased to THz region, potentially the application of the non-resonant-type metamaterial could largely reduce the size and increase the efficiency of on-chip devices. In this work, both MPW and CRLH T-line are explored for on-chip CMOS applications and used to overcome the fundamental limitations posed by conventional T-line for the CMOS designs from mm-wave to THz.

2.2.1 Magnetic Plasmon Waveguide

MPW with zero phase propagation can be introduced in the coupling network design with $2k/N = 0$ to largely improve the output power within compact area. It operates based on the inductive coupling between periodic resonators. The equivalent circuit of an ideal 1D MPW is shown in Fig. 2.2(a). The plasmon resonators are coupled by the magnetic flux between adjacent resonators, which are represented by the LC networks with mutual inductances (M). Assuming the magnetic coupling only exists between adjacent resonators, each unit-cell consists of two magnetic coupled resonators. As such, the dispersion relationship can be written as:

$$\omega_0^2/\omega^2 - 1 = \frac{2M}{L} \cos[(\alpha + j\beta)d] \quad (2.1)$$

where $j = \sqrt{-1}$, $\omega_0 = 1/\sqrt{LC}$ is the self-resonance frequency of LC resonator, d is the distance between adjacent unit-cells, α and β are the attenuation coefficient and phase constant, respectively. Fig. 2.2(b) shows the according dispersion diagram. One can observe that both α and β are zero at the lower stop bands boundary ω_L with

$$\omega_L = \omega_0/\sqrt{1 + 2M/L} \quad (2.2)$$

where the zero phase propagation exists. When multiple MPW unit-cells are serially connected, the in-phase EM-energy can be stored in each unit-cell in zero phase propagation mode. A zero phase propagation mode operation is important for not only power combination but also phase noise reduction. The noise coupling network becomes reciprocal in the zero phase propagation mode, and the total phase noise will be reduced by N times when coupling N free running oscillators [41].

Figure 2.2: (a) Equivalent circuit of magnetic plasmon waveguide (MPW); (b) dispersion diagram of MPW.

2.2.2 Composite Right/Left Handed T-line

CRLH T-line is a well-know metamaterial structure that can achieve positive, negative and zero phase propagation. As shown in the equivalent circuit of an ideal CRLH T-line unit cell in Fig. 2.3(a), it consists of both left handed circuit elements (L_L and C_L) and right handed circuit elements (L_R and C_R). The effective permittivity and permeability of one unit-cell of CRLH T-line are derived by

$$\begin{cases} \varepsilon_{eff} = \frac{C_R}{l} \left(1 - \frac{\omega_P^2}{\omega^2}\right) \\ \mu_{eff} = \frac{L_R}{l} \left(1 - \frac{\omega_S^2}{\omega^2}\right) \end{cases} \quad (2.3)$$

where l is the physical length of one unit-cell, ω is the angular frequency, ω_S and ω_P are the series and parallel resonance frequencies of unit-cell with $\omega_S = \frac{1}{\sqrt{L_R C_L}}$ and $\omega_P = \frac{1}{\sqrt{L_L C_R}}$.

One distinguished feature of metamaterial is observed in (2.3) that ε_{eff} and μ_{eff} are both negative when ω is smaller than ω_P and ω_S . We should note that when ω_S is different from ω_P , there is a band-gap region in between with zero propagation constant (β) and non-zero attenuation factor (α) as shown in Fig.2.3(b). Intuitively, one can apply this zero-propagation nature of CRLH T-line to realize the zero-phase-shifter. As β is zero inside the band-gap of CRLH T-line, there is a much flatter frequency dependence for phase-shifting ($\Delta\phi$) in the CRLH T-line.

It can be shown that the maximum α at the center of the band-gap is

$$\alpha_{MAX} = \frac{\omega_L}{l} \cdot \frac{|\omega_S - \omega_P|}{\sqrt{\omega_S \omega_P}} \quad (2.4)$$

where $\omega_L = \frac{1}{\sqrt{L_L C_L}}$ is the left-handed (LH) high-pass cut-off frequency. One can observe from (2.4) that α_{MAX} is proportional to the product of band-gap $|\omega_S - \omega_P|$ and LH cut-off frequency ω_L . Note that proper trade-off is required between α and zero-phase-shift bandwidth in the real design.

In addition, note that the propagation constant (β) outside the band-gap is

$$\beta = \frac{s(\omega)}{l} \cdot \frac{\omega_L}{\omega} \sqrt{\left(1 - \frac{\omega^2}{\omega_S^2}\right) \left(1 - \frac{\omega^2}{\omega_P^2}\right)} \quad (2.5)$$

where $s(\omega)$ is the sign-function. We can see from (2.5) that for a given frequency outside the band-gap, β is also be reduced by a smaller ω_L , which leads to a larger bandwidth of one zero-phase-shifter. Therefore, reducing ω_L is the key to achieve wide-band and low-loss zero-phase-shifter using CRLH T-line. A smaller ω_L requires

(a)

(b)

Figure 2.3: (a) Equivalent circuit model for the unit-cell of CRLH T-line; (b) dispersion diagram of CRLH T-line

to design larger L_L and C_L on chip.

As CRLH T-line based zero-phase-shifter dose not have any size limitation from

wavelength when compared to the traditional $\lambda/2$ T-line, it can be designed with a much more compact size and high quality factor.

2.3 Resonant-type Metamaterial

Figure 2.4: Layout topologies of SRR in (a) and CSRR in (b), and equivalent circuits of SRR in (c) and CSRR in (d).

Metamaterial based resonators have been explored recently for CMOS MMIC applications. The planar SRR structure can be considered as a magnetic dipole excited by the magnetic field (H-Field) along the ring axis as shown in Fig. 2.4(a). Fig. 2.4(c) shows the equivalent circuit of SRR unit cell, in which the equivalent inductance is coupled to the external applied magnetic flux. As the dual counterpart of SRR, CSRR shown in Fig. 2.4(b) was proposed by [42] based on the well-known complementary theory. CSRR shows the metamaterial property of negative permittivity (ϵ) at resonance frequency, and can be considered as an electric dipole excited by the electric field (E-Field) along the ring axis. Fig. 2.4(d) shows the equivalent circuit of CSRR unit cell, in which the equivalent LC resonator is driven by the external applied electric field.

Recent years, there are amount of works proposed for the oscillator design with high-Q Metamaterial resonators. SRR or CSRR based oscillator design is explored in PCB scale at $5.5 \sim 5.8$ GHz [43, 44]. TL-SRRs have also been studied on PCB substrate with operating frequencies below 10 GHz [45, 46]. A single-ended T-line

loaded with SRRs (STL-SRRs) was designed with silicon substrate for 60 GHz MMIC applications [47]. This structure with multiple SRRs occupies a large silicon area and has weak EM coupling between T-line and SRR load, both of which will contribute to more energy loss. A 24 GHz CMOS oscillator based on open-loop multiple-SRR is presented as another kind of metamaterial resonators in [48].

In the regime from millimeter-wave to THz, the challenge is how to design a low energy loss, strong EM coupling and high area efficiency metamaterial resonator for the MMIC applications. As explored in the mm-Wave region, both SRR and CSRR can be applied as the load for a T-line to build compact on-chip metamaterial-based resonators with high-Q factor. When coupling either SRR or CSRR as load to a host T-line, a plasmonic medium with single negative ε or μ could be formed, called electric or magnetic plasmonic medium. A sharp band-gap or stop-band is formed in such a medium at the resonant frequency such that the EM-wave can be perfectly reflected back into the host T-line to form a stable standing-wave. Ideally, the EM-energy is stored into the compact SRR or CSRR structure, where the energy density can be significantly increased with a high-Q factor.

Figure 2.5: (a) Standing-wave formed by perfect reflection at DTL-SRR; (b) equivalent circuit of DTL-SRR with condition to form perfect reflection.

2.3.1 T-line Loaded with Split Ring Resonator

As shown in Fig. 2.5(a), T-line loaded with SRR (TL-SRR) can be implemented on chip by the top most metal layer. By exciting SRR with the magnetic flux generated

from the differential current in the host T-line, a magnetic plasmonic medium is formed by TL-SRR in the vicinity of SRR resonance frequency. SRRs are excited by the magnetic flux generated by the differential current flowing in the host T-line. The equivalent circuit of TL-SRR unit-cell is depicted in Fig. 2.5(b). L and C are the intrinsic series inductance and shunt capacitance of T-line, while L_s and C_s are the equivalent inductance and capacitance of SRR. M is the mutual inductance between SRR and T-line.

The metamaterial property for TL-SRR can be analyzed by T-line model [49]. Recall that T-line is usually modeled by distributed series impedance (Z) and shunt admittance (Y) with determined ε and μ , respectively. It can be shown that $\varepsilon > 0$ and $\mu < 0$ condition can be satisfied by TL-SRR in the frequency range

$$\sqrt{\frac{1}{L'_s C'_s}} < \omega < \sqrt{\frac{L + L'_s}{L L'_s C'_s}} \quad (2.6)$$

where $L'_s = C_s M^2 \omega^2$ and $C'_s = L_s / (M^2 \omega^2)$ are the equivalent series inductance and capacitance of SRR. Note that M needs to be sufficiently high for a negative μ . As such, differential host T-line is deployed in the design with SRRs placed in between as close as possible.

Figure 2.6: (a) Standing-wave formed by perfect reflection at DTL-CSRR; (b) equivalent circuit of DTL-CSRR with condition to form perfect reflection.

2.3.2 T-line Loaded with Complementary Split Ring Resonator

What is more, as shown in Fig.2.6 (a). T-line loaded with CSRR (TL-CSRR) can be implemented on chip by engraving CSRRs on the T-line with the use of the top most metal layer. By exciting CSRR with the E-Field in the host T-line, An electric plasmonic medium is formed by TL-CSRR in the vicinity of CSRR resonance frequency. The equivalent circuit of TL-CSRR unit-cell is depicted in Fig.2.6 (b). L_C and C_C are the equivalent inductance and capacitance of the CSRR resonator. By comparing the equivalent circuit of TL-CSRR unit-cell with the one of T-line unit-cell, it can be observed that $\varepsilon < 0$ and $\mu > 0$ condition is satisfied in TL-CSRR in the frequency range

$$\sqrt{\frac{1}{C_c L_c}} < \omega < \sqrt{\frac{C + C_c}{C C_c L_c}} \quad (2.7)$$

where evanescent wave is formed.

Figure 2.7: Metamaterial based THz transceiver design including: (a) high power THz signal source by MPW based zero-phase coupled oscillator network (CON), (b) high gain THz antenna by CRLH T-line, and (c) high sensitivity THz super-regenerative receiver by TL-SRR/TL-CSRR based quench-controlled oscillator.

2.4 Metamaterial based THz Transceiver

Fig. 2.7 shows the block diagram of the proposed metamaterial based THz transceiver design. Non-resonant-type metamaterial can be used in the designs of high power signal sources and high gain on-chip antennas; Resonant-type metamaterial can be used in the designs of high sensitivity signal detection. In the design of high power signal sources by MPW based zero-phase coupled oscillator network (CON), N oscillators can be coupled in-phase to generate a N times higher combined output power as well as N times lower phase noise. In the design of high gain CRLH T-line based on-chip LWA, the zero-phase propagation in the CRLH T-line can generate in-phase radiation to largely increase the antenna gain in a very small area. In the design of high sensitivity super-regenerative receivers by TL-SRR/TL-CSRR based quench-controlled oscillators, with the sharp stop-band introduced by the metamaterial resonators, high-Q oscillatory amplifications are generated to largely improve the receiver sensitivity.

Chapter 3

CMOS THz Modeling

3.1 Introduction

Accurate device models that can take into account the loss from strong frequency-dependent dispersion and non-quasi-static effects must be considered in CMOS based THz design. As the most fundamental passive structure, the accurate modeling of transmission line (T-line) is very important in various designs [50, 51]. T-line is traditionally characterized by distributed integer-order RLGC model as shown in Fig. 3.1(a) [52]. Drude's classical relaxation-effect model is deployed for the skin effect with R_S [53, 54]. In addition, the loss due to dielectric polarization and dipole rotation can be modeled by a dielectric-loss of G_D . However, such an integer-order model is insufficient to describe the T-line performance at THz region because the loss term in T-line is difficult to model the dispersion loss and non-quasi-static effects [55], which can cause large deviation at THz frequency region. Such impact is further verified by the measurement results and circuit level simulations in this paper. Moreover, traditional T-line model has causality issue. Physically, the real and imaginary parts of both permittivity $\varepsilon(\omega)$ and permeability $\mu(\omega)$ in a propagation medium are not independent to each other, but follow the Kramers-Kronig relation [56]. As such, the extracted RLGC parameters in traditional T-line model may result in a non-causal response in the model that can induce both accuracy and convergence problems in the time-domain simulation.

The concept of fractional-order model has been examined to model capacitor (C) and inductor (L) at high frequency region. The I-V relation of a capacitor is found to follow the fractional-order [57], and the eddy current and hysteresis effect in inductors are also observed with fractional-order relation [58]. It motivates us to re-examine the RLGC T-line model at THz during the device characterization [59]. Note that

fractional-order model has been deployed to model the surface impedance [60] and describe the abnormal diffusion of voltage and current wave [61]. The fractional-order based impedance matching network [62,63] and resonator design [64] have also been studied. However, no studies investigating the model causality have been carried on with measurement verifications at THz.

In this chapter, two fractional-order T-line models have been developed for both CMOS on-chip conventional RLCG T-line modeling and metamaterial based CRLH T-line modeling at THz with the following advantages. Firstly, the fractional-order T-line models can describe dispersion and non-quasi-static effect in THz. Secondly, by properly deciding the range of fractional-order, the fractional-order RLCG T-line model does not have the causality issue. Lastly, the fractional-order models are still in compact forms that can be extracted from measurement results. The proposed fractional-order RLCG and CRLH T-line modes are verified by S-parameter measurement results in 10 MHz $\sim$ 110 GHz and 220 GHz $\sim$ 325 GHz, respectively. Compared to the conventional integer-order models, the proposed fractional-order T-line models demonstrate improved accuracy of characteristic impedance and propagation constant, which have significant impacts to THz circuit design.

Figure 3.1: RLGC unit-cell equivalent circuits of T-line: (a) integer-order model; and (b) fractional-order model.

3.2 Fractional-order T-line Model

3.2.1 Fractional Calculus

Generally, most of dynamic systems can be described with fractional dynamics, though the fractionality is rather low to be considered than the integer-order behavior. Recently, the fractional-order models are re-examined when considering loss terms in many fields [65] including electronics [57,66,67], electromagnetic [68], fluidic-dynamics [69], material technology [70], quantum mechanics [71], etc.

Fractional calculus was initiated by a question of half-order derivative by L' Hospital in 1695 and was generalized by Euler in [72]. In fractional calculus, the integration and differentiation can be generalized by the operator ${}_aD_t^\alpha$ [73],

$${}_aD_t^\alpha = \begin{cases} \frac{d^\alpha}{dt^\alpha}, & \alpha > 0 \\ \int_a^t (d\tau)^\alpha, & \alpha < 0 \end{cases} \quad (3.1)$$

where α is a real number, a and t are the lower and upper bounds. The differentiation and integration can be treated as the special cases when α equals 1 or -1, respectively. By Riemann-Liouville definition, the fractional operation can be expressed by the following equation ($n - 1 < \alpha < n$),

$${}_aD_t^\alpha f(t) = \frac{1}{\Gamma(n - \alpha)} \frac{d^n}{dt^n} \int_a^t \frac{f(\tau)}{(t - \tau)^{\alpha - n + 1}} d\tau, \quad (3.2)$$

where $\Gamma(\cdot)$ is the Euler's gamma function.

Assuming the lower bound $a = -\infty$, one can take the Fourier transform of 3.2 to obtain the generalized expression of fractional integral in frequency domain for $0 < \alpha < 1$,

$$\mathfrak{F}\{-\infty D_t^{-\alpha} f(t)\} = (j\omega)^{-\alpha} G(\omega) \quad (3.3)$$

Similarly, the generalized expression of fractional derivative in frequency domain is

$$\mathfrak{F}\{D^\alpha f(t)\} = (-j\omega)^\alpha G(\omega) \quad (3.4)$$

As shown in 3.3 and 3.4, the fractional-operator in frequency domain can be treated as the product of a magnitude scaling factor ω^α and a phase rotation factor $j^{-\alpha}$. Theoretically, the physical behavior of any electronic device can be described by these two fractional factors. More importantly, both scaling factor and rotation factor are

linked by a fractional-order α , which ensures the reality of one dynamic system with dissipation. Therefore, in order to examine the loss terms for electronic devices at THz, the aforementioned fractional calculus can be applied with many interesting observations as explored in the following sections.

3.2.2 Fractional-order Capacitance and Inductance

Fractional-order model for T-line can be built by introducing fractional-order terms in the conventional RLGC model as shown in Fig. 3.1 (b). A fractional-order capacitor model [57] with the I-V relation can be given by

$$I(t) = C' \frac{d^{\alpha_C} V(t)}{dt^{\alpha_C}} = C'_0 D_t^{\alpha_C} V(t) \quad (3.5)$$

where C' is the fractional capacitance with order α_C , and $\alpha_C \in (0, 1]$ is the fractional-order relating to the loss of capacitor.

Similarly, the I-V relation of fractional-order inductor [66] is

$$V(t) = L' \frac{d^{\alpha_L} I(t)}{dt^{\alpha_L}} = L'_0 D_t^{\alpha_L} I(t) \quad (3.6)$$

where L' is the fractional inductance with order α_L , and $\alpha_L \in (0, 1]$ is the fractional-order relating to the loss of inductor. The admittance and impedance of fractional-order capacitor and inductor can be obtained from (3.5) and (3.6) by

$$Y'(\omega) = \omega^{\alpha_C} C' e^{0.5\pi j \alpha_C} \quad (3.7)$$

$$Z'(\omega) = \omega^{\alpha_L} L' e^{0.5\pi j \alpha_L} \quad (3.8)$$

When α_L or $\alpha_C \neq 1$, we can expect the existence of real-parts at the right-hand sides of (3.7) and (3.8), which represent the frequency-dependent loss. Physically in a particular device, the fractional-order operator indicates the transfer of the energy storage to energy loss. As such, the distributed frequency-dependent terms are considered by L' and C' elements in fractional-order terms.

3.2.3 Fractional-order T-line Model

Note that the fractional-order T-line can be analyzed in a similar fashion as to the traditional T-line. The characteristic impedance (Z_0) of T-line can be found by $\sqrt{Z/Y}$, where Z and Y are the series impedance and shunt admittance, respectively.

Based on (3.7) and (3.8) with consideration of resistance R_0 and conductance G_0 , one can have

$$Z_0 = \sqrt{\frac{R_0 + \omega^{\alpha_L} L' e^{0.5\pi j \alpha_L}}{G_0 + \omega^{\alpha_C} C' e^{0.5\pi j \alpha_C}}} \quad (3.9)$$

In THz frequency region, ω is in the order of $10^{11} \sim 10^{13}$. At such a high frequency, we have $R_0 \ll \omega^{\alpha_L} L' e^{0.5\pi j \alpha_L}$ and $G_0 \ll \omega^{\alpha_C} C' e^{0.5\pi j \alpha_C}$, so (3.9) can be approximated as

$$Z_0 = \sqrt{\frac{L'}{C'}} \cdot \omega^{\frac{\alpha_L - \alpha_C}{2}} \cdot \left[\cos \frac{(\alpha_L - \alpha_C)\pi}{4} + j \sin \frac{(\alpha_L - \alpha_C)\pi}{4} \right] \quad (3.10)$$

Comparing characteristic impedance by the fractional-order and the conventional integer-order RLGC models, one can observe that the fractional-order Z_0 has nonlinear frequency dependency in THz. If $\alpha_L < \alpha_C$, magnitude of Z_0 has an inverse-square-root-like decreasing function of frequency. This reveals the existence of imaginary part in Z_0 , which accounts for the dispersion and non-quasi-static effects. Both effects are confirmed in the THz T-line measurements.

Moreover, the propagation constant (γ) is

$$\gamma = \alpha + j\beta \quad (3.11)$$

where α is the attenuation constant and β is the phase constant.

As $\gamma = \sqrt{ZY}$, with (3.7) and (3.8), one can have

$$\gamma = \sqrt{(R_0 + \omega^{\alpha_L} L' e^{0.5\pi j \alpha_L})(G_0 + \omega^{\alpha_C} C' e^{0.5\pi j \alpha_C})} \quad (3.12)$$

In THz frequency region, since $R_0 \ll |\omega^{\alpha_L} L' e^{0.5\pi j \alpha_L}|$ and $G_0 \ll |\omega^{\alpha_C} C' e^{0.5\pi j \alpha_C}|$, (3.12) can be approximated as

$$\gamma = \sqrt{L'C'} \cdot \omega^{\frac{\alpha_L + \alpha_C}{2}} \cdot \left[\cos \frac{(\alpha_L + \alpha_C)\pi}{4} + j \sin \frac{(\alpha_L + \alpha_C)\pi}{4} \right] \quad (3.13)$$

Comparing propagation constant by the fractional-order and the conventional integer-order RLGC models, one can observe that the fractional-order γ also has nonlinear frequency dependency in THz. The attenuation constant α will become non-zero when $\alpha_L + \alpha_C < 2$ in (3.13), and the energy loss is introduced accordingly. What is more, note that the S-parameters of T-line are determined by Z_0 and γ [74].

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} \frac{A+B/Z_0-CZ_0-D}{A+B/Z_0+CZ_0+D} & \frac{2(AD-BC)}{A+B/Z_0+CZ_0+D} \\ \frac{2}{A+B/Z_0+CZ_0+D} & \frac{-A+B/Z_0-CZ_0+D}{A+B/Z_0+CZ_0+D} \end{bmatrix}. \quad (3.14)$$

where A, B, C and D are the transfer matrix of uniform T-line:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cosh \gamma l & Z_0 \sinh \gamma l \\ \frac{\sinh \gamma l}{Z_0} & \cosh \gamma l \end{bmatrix}. \quad (3.15)$$

By substituting (3.9) and (3.12) into (3.14), one can obtain the S-parameters of the fractional-order T-line model. As such, the models can be verified by the S-parameters measurement results. In addition, it is well known that the propagation velocity of the EM wave front (V_p) equals to ω/β . In the fractional-order T-line model, one can have

$$V_p = \frac{\omega^{2-\frac{\alpha_L+\alpha_C}{2}}}{\sqrt{L'C'} \sin \frac{(\alpha_L+\alpha_C)\pi}{4}} \quad (3.16)$$

In the fractional-order T-line model, when $\alpha_L + \alpha_C < 2$ in (3.16), V_p becomes non-linearly frequency dependent in high frequency region like THz. This reveals dispersion effect when propagating through lossy media. On the other hand, in the integer-order T-line model, when $\alpha_L + \alpha_C = 2$ in (3.16), V_p equals $1/\sqrt{L'C'}$, which is not frequency dependent, to model the dispersion effect in high frequency region like THz.

3.2.4 Fractional-order CRLH T-line Model

Note that the fractional-order CRLH T-line can be analyzed in a similar fashion as fractional-order T-line. Base on the integer-order CRLH T-line model with general series and shunt loss terms (R and G) in Fig. 3.2(a), fractional-order terms are introduced in the fractional-order CRLH T-line model as shown in Fig. 3.2(b). Note that R_0 and G_0 only represent the series resistance and the shunt conductance at DC condition, respectively. Based on 3.7 and 3.8 with consideration of R_0 and G_0 , one can have the characteristic impedance of CRLH T-Line as

$$Z_0 = \sqrt{\frac{R_0 + \omega^{\alpha_{LS}} L'_S e^{0.5\pi j \alpha_{LS}} + \frac{1}{\omega^{\alpha_{CS}} C'_S e^{0.5\pi j \alpha_{CS}}}}{G_0 + \omega^{\alpha_{CP}} C'_P e^{0.5\pi j \alpha_{CP}} + \frac{1}{\omega^{\alpha_{LP}} L'_P e^{0.5\pi j \alpha_{LP}}}}} \quad (3.17)$$

From the equivalent circuit of CRLH T-line, we find that each cell consist parallel

Figure 3.2: Equivalent circuits of CRLH unit-cell: (a) integer-order model; and (b) fractional-order model.

and series LC resonators, and there is a gap between parallel resonate frequency (ω_P) and series resonate frequency (ω_S). When $\omega = \omega_P$, Z_0 reach maximum value, when $\omega = \omega_S$, Z_0 reach its minimum value, thus Z_0 shows a peek-valley or valley-peek curve as frequency grows within the whole range. This will be verified by simulation and measurement results shown in Fig. 3.14. According to (3.17), real terms of series resonator and parallel conductance are expressed as

$$\begin{cases} \Re(Z) = R_0 + \omega^{\alpha_{L_S}} L'_S \cos(0.5\pi j\alpha_{L_S}) + \frac{\cos(0.5\pi j\alpha_{C_S})}{\omega^{\alpha_{C_S}} C'_S} \\ \Re(Y) = G_0 + \omega^{\alpha_{C_P}} C'_P \cos(0.5\pi j\alpha_{C_P}) + \frac{\cos(0.5\pi j\alpha_{L_P})}{\omega^{\alpha_{L_P}} L'_P} \end{cases} \quad (3.18)$$

One can observe from (3.18) that, different from conventional integer order model, fractional model introduces frequency-depended terms to loss and conductance equations that greatly affect the peak and valley magnitudes of Z_0 .

In THz frequency region, ω is in the order of 10^{11} 10^{13} , such that $R_0 \ll \omega^{\alpha_{L_S}} L'_S e^{\frac{j\alpha_{L_S}\pi}{2}}$, and $G_0 \ll \omega^{\alpha_{C_P}} C'_P e^{\frac{j\alpha_{C_P}\pi}{2}}$. At a frequency that much lower than zero-phase-shift frequency, the left-handed terms ($\frac{1}{\omega^{\alpha_{C_S}} C'_S e^{\frac{j\alpha_{C_S}\pi}{2}}}$ and $\frac{1}{\omega^{\alpha_{L_P}} L'_P e^{\frac{j\alpha_{L_P}\pi}{2}}}$) becomes dominant in (3.17). Thus Z_0 can be approximated as

$$Z_0 = \sqrt{\frac{L'_P}{C'_S}} \cdot \omega^{\frac{\alpha_{LP} - \alpha_{CS}}{2}} \cdot \left[\cos \frac{(\alpha_{LP} - \alpha_{CS})\pi}{4} + j \sin \frac{(\alpha_{LP} - \alpha_{CS})\pi}{4} \right] \quad (3.19)$$

On the other hand, at a frequency that much lower than zero-phase-shift frequency, the right-handed terms ($\omega^{\alpha_{LS}} L'_S e^{0.5\pi j \alpha_{LS}}$ and $\omega^{\alpha_{CP}} C'_P e^{0.5\pi j \alpha_{CP}}$) becomes dominant in (3.17). As such, Z_0 can be approximated as

$$Z_0 = \sqrt{\frac{L'_S}{C'_P}} \cdot \omega^{\frac{\alpha_{LS} - \alpha_{CP}}{2}} \cdot \left[\cos \frac{(\alpha_{LS} - \alpha_{CP})\pi}{4} + j \sin \frac{(\alpha_{LS} - \alpha_{CP})\pi}{4} \right] \quad (3.20)$$

From (3.19) and (3.20), one can observe that Z_0 is a frequency-depended parameter, which is determined by $\omega^{\frac{\alpha_{LP} - \alpha_{CS}}{2}}$ and $\omega^{\frac{\alpha_{LS} - \alpha_{CP}}{2}}$ at low and high frequencies outside the band-gap, respectively. For an on-chip CRLH T-line design, the fractional orders for inductive elements (α_{LS} and α_{LP}) is usually smaller than that of capacitive elements (α_{CS} and α_{CP}) due to a lower Q factor of inductors. Therefore, both $\omega^{\frac{\alpha_{LP} - \alpha_{CS}}{2}}$ and $\omega^{\frac{\alpha_{LS} - \alpha_{CP}}{2}}$ are negative value. As a result, Eq. (3.17) becomes a negative function of frequency outside the band-gap.

Moreover, since R_0 and G_0 are negligible at THz, the propagation constant of CRLH T-Line can be written as

$$\gamma = \sqrt{(\omega^{\alpha_{LS}} L'_S e^{0.5\pi j \alpha_{LS}} + \frac{1}{\omega^{\alpha_{CS}} C'_S e^{0.5\pi j \alpha_{CS}}}) \cdot (\omega^{\alpha_{CP}} C'_P e^{0.5\pi j \alpha_{CP}} + \frac{1}{\omega^{\alpha_{LP}} L'_P e^{0.5\pi j \alpha_{LP}}})} \quad (3.21)$$

Assuming the fractionality for both inductance and capacitance are all constants that $\alpha_{LS} = \alpha_{LP}$ and $\alpha_{CS} = \alpha_{CP}$, (3.11) can be simplified as

$$\gamma = \sqrt{\omega^{\alpha_{LS} + \alpha_{CP}} L'_S C'_P e^{0.5\pi j (\alpha_{LS} + \alpha_{CP})} + \frac{1}{\omega^{\alpha_{CS} + \alpha_{LP}} C'_S L'_P} e^{-0.5\pi j (\alpha_{CS} + \alpha_{LP})}} \quad (3.22)$$

The zero-phase-shift frequency (ω_0) can be obtained from (3.22) with $\beta = 0$:

$$\omega_0 = (L'_S C'_P C'_S L'_P)^{\frac{1}{\alpha_{LS} + \alpha_{CP} + \alpha_{CS} + \alpha_{LP}}} \quad (3.23)$$

Eq. (3.23) reveals an exponential relationship between the prefactors and fractional order terms, which can be used as a guideline in the fractional order modeling of CRLH

T-line network.

3.3 Model Extraction and Causality Analysis

T-line is a passive, linear and time-invariant (LTI) network. The extracted T-line model is thereby needed to be causal. The extraction flow of fractional-order T-line model is introduced with the additional causality checking and enforcement followed by comparison with the traditional integer-order counterpart.

3.3.1 Fractional-order Model Extraction

A fractional-order model parameters extraction flow for T-line at THz is illustrated in Fig. 3.3. The extraction begins with the measurement data obtained from a Vector Network Analyzer. Firstly, the measurement data is converted into transfer matrix (T matrix) for an easy operation, and the error terms contributed by the testing pads are removed by de-embedding process. Secondly, characteristic impedance Z_0 and propagation constant γ are calculated from de-embedded T-matrix according to [75]. Afterwards, one can define the modeling frequency interval $[\omega_1, \omega_2]$ in THz region based on his interests.

From (3.10), one can have

$$\alpha_L - \alpha_C = 2 \log_{\frac{\omega_1}{\omega_2}} \left| \frac{Z_0(\omega_1)}{Z_0(\omega_2)} \right| \quad (3.24)$$

where $Z_0(\omega_1)$ and $Z_0(\omega_2)$ are the characteristic impedances at frequencies ω_1 and ω_2 in THz region, respectively.

From (3.13), one can have

$$\alpha_L + \alpha_C = 2 \log_{\frac{\omega_1}{\omega_2}} \left| \frac{\gamma(\omega_1)}{\gamma(\omega_2)} \right| \quad (3.25)$$

where $\gamma(\omega_1)$ and $\gamma(\omega_2)$ are the propagation constants at frequencies ω_1 and ω_2 in THz region, respectively. By combining (3.24) and (3.25), α_L and α_C can be obtained in the fractional-order model; and by substituting α_L and α_C into (3.24) and (3.25), fractional-order L' and C' can be obtained as well. Note that L' and C' are the p.u.l. (per-unit-length) prefactors with corresponding units of $V s^{-\alpha_L} A^{-1}/m$ and $A s^{-\alpha_C} V^{-1}/m$, respectively, but not p.u.l. inductance and capacitance anymore.

Moreover, in order to apply the fractional-order T-line model in the time-domain

simulator likes Cadence Spectre. The model needs to be converted from frequency domain into time-domain by rational functional approximation

$$f(s) \approx \sum_{j=1}^N \frac{c_j}{s - a_j} + d + sh \quad (3.26)$$

Here, N is the rational order, a_j and c_j are the poles and residues in complex conjugate pairs, d and h are real. The coefficients in (3.26) can be obtained by vector fitting algorithm as introduced in [76]. Note that the error introduced by the frequency to time conversion is well controlled by increasing the rational-fitting order.

Figure 3.3: Fractional-order T-line modeling parameters extraction flow.

3.3.2 Causal LTI System and Causality Enforcement

To understand the causality for the extracted fractional-order T-line model, the fundamentals of causal LTI system are first reviewed here. In a LTI system, the impulse response $h(t)$ to an input $x(t)$ can be expressed as [77]:

$$y(t) = x(t) * h(t) = \int_{-\infty}^{+\infty} h(t - \tau) x(\tau) d\tau \quad (3.27)$$

where $x(t)$ and $y(t)$ represent the input and output voltages, currents or powers of T-line network, and $h(t)$ is the corresponding admittance or impedance state matrix.

The principle of causality states that there is no effect happened before its cause. As such, a causal LTI system $h(t)$ can be mathematical defined as:

$$h(t) = 0, \forall t < 0. \quad (3.28)$$

The causality of T-line model can be verified by this definition in time-domain. (3.28) can be equivalently represented as

$$h(t) = \text{sign}(t)h(t). \quad (3.29)$$

where sign function $\text{sign}(t)$ equals -1 when $t < 0$ and equals 1 when $t > 0$.

By taking Fourier transform of (3.29), we can obtain the impulse response of $h(t)$ in frequency domain with a complex function

$$H(\omega) = \mathfrak{F}\{h(t)\} = \Re[H(\omega)] + j\Im[H(\omega)]. \quad (3.30)$$

with

$$\Re[H(\omega)] = \frac{2}{\pi} \int_0^\infty \frac{\omega' \Im[H(\omega')]}{\omega'^2 - \omega^2} d\omega'. \quad (3.31)$$

and

$$\Im[H(\omega)] = -\frac{2\omega}{\pi} \int_0^\infty \frac{\Re[H(\omega')]}{\omega'^2 - \omega^2} d\omega'. \quad (3.32)$$

Here $\Re[H(\omega)]$ and $\Im[H(\omega)]$ are the coefficients of real and imaginary parts of $H(\omega)$, respectively, which are both real numbers. Equation (3.28) is also addressed as Kramers-Kronig relation or Hilbert transform [56]. Note that (3.31) and (3.32) are bidirectional equations, which reveal the dependency of real and imaginary part of impulse response, and also provide a necessary and sufficient condition for a causal LTI system. Note that the causality of a LTI system can be enforced by correcting the real or imaginary part in (3.31) and (3.32) with truncation. However, truncation error could be also introduced with largely reduced accuracy for simulation.

One criteria to verify causality by tabulated S-parameters is to measure the error difference (e_{ij}) between imaginary part and its Hilbert transform of real part

$$e_{ij}(\omega_n) = |\text{Hilbert}\{\Re[S_{ij}(\omega_n)]\} - \Im[S_{ij}(\omega_n)]|. \quad (3.33)$$

Whether a system is causal or non-causal is determined by the error threshold that can be tolerated in the numerical analysis [78]. A smaller e_{ij} is usually desired to

ensure the accuracy and the convergence in simulation. A flow of causality verification and enforcement is illustrated in Fig. 3.4. Firstly, a very wide band initial tabulated S-parameter needs to be generated by numerical calculation from the model under investigation. Secondly, both the real and imaginary parts are extracted, and Hilbert transform is applied to the real parts according to (3.32). Finally, the causality is verified by calculating e_{ij} from (3.33); and enforced by replacing the imaginary part of the original tabulated data with the Hilbert transform of the real part in the frequency band of interest. Note that the bandwidth of initial data must be much larger than the final frequency band after causality enforcement to minimize reconstruction and discretization errors [79].

Figure 3.4: Causality verification and enforcement flow.

3.3.3 Causality of T-line Model

The relationship between real and imaginary parts in (3.31) and (3.32) also relates the amplitude and phase. Provided the magnitude, one can calculate the phase and vice versa. According to the theory of linear system, any causal and stable impulse response $H(\omega)$ can be decomposed into the production of the minimum phase function and all-pass function [80] by

$$H(\omega) = H_{min}(\omega) \cdot H_{all}(\omega). \quad (3.34)$$

where $H_{min}(\omega)$ and $H_{all}(\omega)$ are the minimum phase function and all-pass function, respectively. Since $H_{all}(\omega)$ does not contain any magnitude information, the system causality can also be ensured if it satisfies the minimum phase function defined by the following condition [81, 82]:

$$\lim_{\omega \rightarrow \infty} \left[\frac{\gamma(\omega)}{j\omega} \right] \Rightarrow 0. \quad (3.35)$$

For T-line, the minimum phase function can be calculated by substituting (3.13) into (3.35) as

$$\lim_{\omega \rightarrow \infty} \left(\frac{\gamma(\omega)}{j\omega} \right) = \lim_{\omega \rightarrow \infty} \left[\omega^{\frac{\alpha_L + \alpha_C}{2}} \cdot \sqrt{L'C'} \cdot \left(\sin \frac{(\alpha_L + \alpha_C)\pi}{4} - j \cos \frac{(\alpha_L + \alpha_C)\pi}{4} \right) \right]. \quad (3.36)$$

We can observe that (3.36) shows very different responses for fractional-order and integer-order T-line models. For the fractional-order model, (3.36) equals zero when $\alpha_L + \alpha_C < 2$. So the causality is always ensured as the minimum phase function condition when (3.35) is satisfied. On the other hand, α_L and α_C are both equal to one for the integer-order T-line model, and (3.36) results in a constant value of $\sqrt{L'C'}$, where L' and C' become the normal inductance and capacitance, respectively. Thus the minimum phase function condition in (3.35) is violated and model becomes non-causal.

Note that the major reason of non-causal issue in the traditional integer T-line model is due to the linear frequency dependence of the propagation constant $\gamma(\omega)$ when $\alpha_L + \alpha_C = 2$ in (3.13). This cannot model the dispersion loss and non-quasi-static effects in the high frequency application like THz. The reality of integer-order T-line model is lost in THz region, and so is the causality. In contrast, the non-ideal effects are considered in the proposed fractional-order T-line model by fractional-order dispersion terms, which can largely improve the model reality. As such, both the model accuracy and causality are improved. The causality of the fractional-order model can also be verified in numerical calculation by computing the error terms in (3.33), which will be discussed in the following section.

3.4 Measurements

3.4.1 T-line Fractional-order Model Verification

Figure 3.5: T-line testing structure, (a) die photo, and (b) detailed dimensions.

As shown in Fig. 3.5(a), a coplanar waveguide transmission line (CPW-TL) testing structure with RF-PADs is fabricated with Global Foundry 1P8M 65nm CMOS process, of which the dimensions are given in Fig. 3.5(b). The CPW-TL is implemented on the top metal layer with thickness of $3.3 \mu\text{m}$. It is measured on a CASCADE Microtech Elite-300 probe station by Agilent PNA-X (N5247A) with frequency sweep up to 110 GHz. The measurement setup of S-parameters up to 110 GHz is illustrated in Fig. 3.6. The reference plane of PNA is calibrated to the ends of GSG probes by SOLT method. Note that both the probes and the impedance standard substrate are provided by Cascade Microtech. RF-PADs on both sides are de-embedded from the measurement results with "open-short" method. TABLE 3.1 summarizes extracted model parameters of both integer-order and fractional-order models based on measurement results. The parameters of traditional integer-order model are extracted according to the procedure by [75]. The parameters of fractional-order model are extracted according to Section 3.3.1.

The resulting S-parameters and characteristic impedance (Z_0) of integer-order and fractional-order RLGC models are compared in Fig. 3.8. We can observe that both the traditional integer-order model and the proposed fractional-order model can fit the measurement results in magnitude in Fig. 3.7. Here a relatively large deviation is observed in magnitude of S11 between the simulation and measurement results. This deviation comes from the equipment noise and calibration error, which is unavoidable as the absolute magnitude of S11 is small ($-15 \sim -50 \text{ dB}$). Moreover, the phase delay of both traditional integer-order model and the proposed fractional-order model

Figure 3.6: Measurement setup of on-wafer S-parameter testing up to 110GHz.

agree well with the measurement results as shown in Fig. 3.8. However, it is observed that characteristic impedance Z_0 in the traditional integer-order RLGC model has deviated from measurement results above 10 GHz, and almost approaches a constant above 40GHz. On the other hand, the fractional-order RLGC model closely fits the measured Z_0 up to 110 GHz, because it can accurately consider frequency-dependence loss yet in a compact RLGC form. At 100 GHz, the Z_0 from fractional-order and measurement results are 34.3 Ω , which is 3.1 Ω lower than the one from integer-order model. Note that such difference will keep increasing with frequency and largely affects the model accuracy in traditional integer-order T-line model at THz. Physically, the values of fractional-order terms (α_L and α_C) model the frequency-dependent dispersion loss of device at THz. For the T-line fabricated by on-chip CMOS process, larger loss is observed in metal layer than in dielectric layer. As a result, α_L has a relatively large deviation from 1, while α_C is close to 1. But note that a slight change in the order-terms (α_L and α_C) could bring huge changes in the prefactors (L' and C') in THz region as observed in (3.7) and (3.8).

3.4.2 CRLH T-line Fractional-order Model Verification

In order to minimize the characterization error of each CRLH T-line unit-cell, one 13-cell CRLH T-line is fabricated with Global Foundry 1P8M 65nm CMOS process. As shown in Fig. 3.9, it has a chip size of $145\mu\text{m} \times 660\mu\text{m}$ excluding the RF Pads. The layout and dimension of each unit cell is shown in Fig. 3.10. The top most aluminum layer (LB) is exclusively employed as signal layer for the maximum distance to the

Figure 3.7: Verification of fractional-order T-line model with measurement results for magnitude of S11 and S21 in dB.

Figure 3.8: Verification of fractional-order T-line model with measurement results for phase delay of S21 and characteristic impedance.

bottom ground layer (M1) to improve the radiation efficiency. Various components in the CRLH T-line cell in Fig. 3.2(a) are synthesized by on-chip structures. The L_P of the CRLH T-line is synthesized by a microstrip line connected to the ground and C_S is implemented with inter-digital capacitor. Both right-handed elements L_S and C_P are contributed by the intrinsic parasitic. Note that a mesh structure is applied

TABLE 3.1: MODELING PARAMETERS OF INTEGER-ORDER AND FRACTIONAL-ORDER RLGC MODEL FOR T-LINE

Integer-Order Model			Fractional-Order Model		
Parameter	Value	Unit	Parameter	Value	Unit
L	247.5	nH/m	α_L	0.862	-
C	0.188	nF/m	L'	10022	$Vs^{-\alpha_L} A^{-1}/m$
R	1200	Ω/m	α_C	0.988	-
R_S	12.56	$m\Omega/m \cdot rad$	C'	0.278	$As^{-\alpha_C} V^{-1}/m$
G	0.079	S/m	R	1200	Ω/m
G_D	19.33	$pS/m \cdot rad$	G	0.079	S/m

in ground layer to satisfy the metal density rule.

Figure 3.9: Chip micrograph of fabricated CRLH T-line in 65nm CMOS process.

Figure 3.10: Dimension of each unit cell and layer configurations of LWA.

The 13-cell CRLH T-line design is verified by circuit simulation in ADS from 220 to 325 GHz. From this we get the integer-order and fractional-order simulation results.

The fabricated 13-cell CRLH T-line structure is measured on probe station (CASCADE Microtech Elite-300) with VNA extender (VDI WR3.4-VNAX). Two waveguide GSG probes with $50\ \mu\text{m}$ pitch are used for the S-parameter measurement from 220 to 325 GHz, as shown in Fig. 3.11. Note that the testing pads and traces are de-embedded (open, short) from both sides with recursive modeling technique [83]. We also compare the measurement results of fabricated CRLH T-line with integer-order and fractional-order circuit simulation as well as EM-simulation by HFSS. The circuit simulations are conducted with the equivalent circuits of unit-cell shown in Fig. 3.2, and the values of circuit elements are summarized in TABLE 3.2, obtained by curve fitting technique.

Figure 3.11: Measurement setup of on-wafer S-parameter testing from 220GHz to 325GHz.

TABLE 3.2: MODELING PARAMETERS OF INTEGER-ORDER AND FRACTIONAL-ORDER MODELS OF CRLH T-LINE UNIT-CELL

Integer-Order Model			Fractional-Order Model		
Parameter	Value	Unit	Parameter	Value	Unit
L_S	15.6	pH	$\alpha_{L_S}/\alpha_{L_P}$	0.9847/0.9766	-
C_S	14.7	fF	$\alpha_{C_S}/\alpha_{C_P}$	0.9939/0.9973	-
G	1.3	mS	L'_S	14	$V_S^{-\alpha_{L_S}} A^{-1}$
R	2.8	Ω	C'_P	1732.1	$A_S^{-\alpha_{C_P}} V^{-1}$
C_P	13.8	S/m	L'_P	39.41	$V_S^{-\alpha_{L_P}} A^{-1}$
L_P	28.3	pH	C'_S	1408	$A_S^{-\alpha_{C_S}} V^{-1}$
-	-	-	R_0/G_0	0.3396/902	Ω/mS

As shown in Fig. 3.12 and 3.13, the phase and magnitude of S21 are almost identical for both fractional-order and integer-order models in the measured frequency

range of 220-325GHz. But the extracted phase constant (β) of fractional-order model is closer to measurement than that of integer-order one while considering the dispersion effects. More importantly, fractional-order model accurately fits the measurement results at the frequency with $\beta = 0$, which is the boundary between left-handed and right-handed regions, while that from integer-order model is 13 GHz less. Moreover, Fig. 3.14 shows a remarkable difference between integer-order and fractional-order results in terms of characteristic impedance Z_0 . The measurement Z_0 fit very well to fractional-order model at zero-phase-shift region from 260 GHz, also a smaller error of Z_0 at low frequency region compared to integer-order fitting result. The average accuracy improvement of 78.8% is obtained by fractional-order model compared to the integer-order counterpart with correlated measurement and simulation results of Z_0 . Moreover, the measurement results of CRLH T-line agree well with the EM simulation results for the frequency range of 220 ~ 325 GHz.

Figure 3.12: Measurement, EM, integer-order and fractional-order circuits simulation results: magnitude of S21 and S11 in dB.

Figure 3.13: Measurement, EM, integer-order and fractional-order circuits simulation results: phase of S21 in degree and the absolute value of extracted phase constant β .

3.4.3 Causality Verification and Comparison

The causality of the proposed fractional-order T-line model can be verified by comparing imaginary parts of S-parameters with the Hilbert transform of real parts. Then, the error term $e_{ij}(\omega_n)$ is calculated by (3.33) as discussed in Section 3.3. For the purpose of comparison, the causality of traditional integer-order T-line model is also verified in the same way. The tabulated results for both models are obtained by two-port S-parameter simulation in Agilent Advanced Design System (ADS) based on the extracted model parameters shown in TABLE 3.1. For a two-port network, four sets of complex S-parameter results can be obtained including S11, S22, S12 and S21. However, according to the reciprocal property of T-line structure ($S_{11} = S_{22}$ and $S_{12} = S_{21}$), only S11 and S21 are considered in the causality analysis. In order to minimize reconstruction and discretization errors [79] introduced by finite spectrum, the S-parameter simulation is conducted from 0Hz to 20THz with a step size of 1GHz.

Figure 3.14: Measurement, EM, integer-order and fractional-order circuits simulation results: characteristic impedance of CRLH T-line (Z_0).

Firstly, the causality of return loss (S_{11}) is verified for both integer and fractional order T-line models. Figs. 3.15 and 3.16 show the comparison between $\Im(S_{11})$ and the value obtained by Hilbert transformation from real part $\text{HilbertReal}(S_{11})$ for both integer-order and the proposed fractional-order T-line models in the frequency range of $0.001 \sim 1$ THz, respectively. For the traditional integer-order RLCG T-line model, the $\Im(S_{11})$ starts to deviate from the causal response at 10 GHz and shows large deviation in $0.1 \sim 1$ THz as depicted in Fig. 3.15. But for the proposed fractional-order T-line model as shown in Fig. 3.16, we can observe that the $\Im(S_{11})$ closely fits the causal response obtained from Hilbert transformation $\text{Hilbert}\{\Re(S_{11})\}$. The error magnitude of $\Im(S_{11})$ from both models are compared in Fig. 3.17, where a dramatic error reduction is observed by the application of fraction order T-line model. Note that the error magnitude is calculated by

$$e_{11} = |\text{Hilbert}\{\Re(S_{11})\} - \Im(S_{11})| \quad (3.37)$$

Secondly, the causality of return loss (S_{21}) is verified for both integer and fractional order T-line model. The comparison between $\text{Imag}(S_{21})$ and causal response for both models are illustrated in Figs. 3.18 and 3.19. For the traditional integer-order RLCG T-line model, the $\Im(S_{21})$ obtained from the integer-order RLCG T-line model deviates from the causal response ($1 \sim 10$ GHz) as depicted in Fig. 3.18. But for the proposed fractional-order T-line model as shown in Fig. 3.19, we can observe that the $\Im(S_{21})$ of

Figure 3.15: Causality verification by Hilbert transformation: integer-order T-line model S11.

Figure 3.16: Causality verification by Hilbert transformation: fractional-order T-line model S11.

the fractional-order T-line model closely fits the causal response. A clear comparison by error magnitude of $\Im(S_{21})$ from both models is illustrated in Fig. 3.20, where a dramatic error reduction is also observed by the proposed fractional-order T-line model. Note that the error magnitude is calculated by

Figure 3.17: Causality verification by Hilbert transformation: error magnitude comparison of S11

$$e_{21} = |\text{Hilbert}\{\Re(S_{21})\} - \Im(S_{21})| \quad (3.38)$$

Note that since both e_{11} and e_{21} for fractional-order T-line model are rather small, the causality enforcement by (3.31) and (3.32) is not required. The resulting frequency model can be directly used to estimate the time-domain model by the rational fitting. As such, the best accuracy could be ensured in the time-domain simulation such as Transient Analysis or Periodic Steady State (PSS) Analysis. But for the traditional integer-order RLCG T-line model, the non-causal effect could have convergence issues. One way to alleviate the causality issue of the integer-order RLCG T-line model is by truncating the model data with the causality enforcement, but the accuracy is lost in this way as discussed in Section 3.3.

3.5 Conclusion

Accurate device model is critical for CMOS based THz circuit design. Since transmission line (T-line) is one of the most fundamental passive devices commonly used in the THz circuit design, an in-depth study of T-line model at THz is thereby important. Note that dispersion and non-quasi-static effects are difficult to be modeled by traditional methods such as the integer-order RLCG model with causality. By the

Figure 3.18: Causality verification by Hilbert transformation: integer-order T-line model S21.

Figure 3.19: Causality verification by Hilbert transformation: fractional-order T-line model S21.

proposed compact and causal fractional-order T-line model, the causality concern is resolved for T-line via considering the frequency-dependent dispersion loss and non-quasi-static effect at THz. The measured results have confirmed that the proposed fractional-order RLGC T-line model and CRLH T-line model have improved accuracy

Figure 3.20: Causality verification by Hilbert transformation: error magnitude comparison of S21

over the traditional integer-order models from mm-wave to THz region. Accordingly, the proposed fractional-order T-line models will be applied in the design for CMOS based THz circuits in the following sections.

Chapter 4

CMOS THz Signal Generation

4.1 Introduction

CMOS-based THz signal sources have been recently demonstrated for compact system-on-chip implementation [84–92] with applications in high-data-rate communication and non-invasive imaging. As THz signal suffers from large propagation loss, the generated signal strength must be strong enough to have sufficient signal to noise ratio (SNR) for detection, which imposes grand challenges for the high power signal source designs in CMOS.

It is challenging to design a high output power THz signal source by single CMOS oscillator source with high efficiency as well as wide frequency tuning range (FTR). In CMOS process, one single oscillator usually has small output power that limited by single CMOS transistor with commonly observed low output power level [41, 85, 93], which is incapable of delivering a strong THz signal for transmission and processing. Similar to the approach utilized in power amplifier [94, 95], output power combining of multiple CMOS oscillators can be considered to achieve a large output power at THz frequency. Several coupled-oscillator-network (CON) structures have been proposed for the high output power [92, 96]. A phase/delay tuning design is utilized in [92] but cannot ensure the in-phase coupling condition. In-phase synchronization by half λ transmission line (T-line) is reported in [96], which is bulky and lossy when deployed for a large-scale phase-arrayed design. There is great interest to explore a high output power signal source by combining the outputs of an array of CMOS oscillators such as CON [97].

The combined output power of CON is maximized when all the oscillator unit-cells are in-phase coupled with a phase difference of $2n\pi$ ($n = 0, 1, 2, \dots$). In the conventional CON design [97], adjacent oscillators are coupled by a T-line or an equivalent

delay network with at least an electrical length of half λ ($n = l$), which is bulky and lossy and hence largely reduces the output power and efficiency. The recent work in [94] has shown that the CMOS metamaterial devices can be utilized for phase-arrayed design such as zero-phase-shifters ($n = 0$) for power combing in CMOS mm-wave PAs with very high output power. Compared to the coupling by half λ T-line, the zero-phase-shifter can result in zero-phase coupling with low loss and compact area.

Recently years, plasmon polaritons based waveguide structures have been explored for low loss energy transfer [31,32,98–101]. As one type of the plasmon polaritons based waveguide structures, Magnetic plasmon waveguide (MPW) has been already applied to the middle distance wireless energy transmission systems [31, 32] as well as one-dimensional sub-wavelength power transfer in the mm-wave range [98–100]. Similar to CRLH T-line, MPW has a zero-phase propagation mode. But it also has several advantages over CRLH T-line in the CON design. Firstly, MPW has lower propagation loss in the practical CMOS layout. As introduced in Sec. 2.2.2, there will be a band-gap in CRLH T-line if the series and parallel resonance frequencies of unit-cell are not same, which will contribute the propagation loss. On the other hand, MPW only has one zero-phase mode at the boundary of stop-band, where its attenuation factor is zero. Secondly, the plasmon resonators inside MPW can be easily transferred into oscillator unit-cells by replacing their equivalent capacitance with transistors based negative resistance. As such, a distributed zero-phase CON is proposed with magnetic-plasmon-waveguide (MPW) based oscillator unit-cells. When a number of zero-phase oscillator unit-cells are serially connected in a ring with a centralized placement of active devices, in-phase coupling can be achieved with low loss. The resulting CON output can be significantly improved with a reduced phase noise because outputs of all zero-phase coupled (ZPC) oscillator unit-cells are synchronized for an in-phase combination. With the further use of inductive tuning [102], a wide FTR can be achieved for each ZPC oscillator unit-cell.

In this chapter, firstly, a zero-phase CON with 4 unit-cells is implemented at 60 GHz in 65nm CMOS process. The measured results show a peak output power of 2 mW with 2.2% power efficiency, phase noise of -116.7 dBc/Hz at 10-MHz offset and FTR of 16% from 58 to 69.1 GHz. Secondly, an injection-locked THz signal source with zero-phase coupled oscillator network is proposed to provide high output power and low phase noise within compact chip area. A zero-phase oscillator unit-cell is developed by inter-digital coupler with 0.4-dB loss at 70 GHz. With four in-phase coupled unit-cells and push-push frequency doublers, high output power signal is generated at the center of the proposed CON at 140 GHz. The measured results show a peak output

power of 3.5 mW with 2.4% power efficiency, power density (or output power/area) of 26.9 mW/mm² and 9.7% FTR centered at 133.5GHz

4.2 MPW with capacitance loaded coupled-line

In this work, magnetic plasmon waveguide (MPW) with zero phase propagation is introduced in the coupling network design with $2k/N = 0$ to largely improve the output power within compact area. A zero phase propagation is not only important for power combination but also the phase noise reduction. The noise coupling network becomes reciprocal in zero-phase mode, and the phase noise at CON output becomes $1/N$ of a single free running oscillator [41].

Figure 4.1: (a) Equivalent circuit of differential ZPC loaded with parasitic capacitance (cross-coupled pairs); (b) on-chip realization of inter-digital coupled T-lines.

The MPW unit-cell can be implemented on-chip by coupled T-line based resonator with C contributed by the parasitic capacitances of transistors as shown in Fig. 4.1(a). The two-port Y-parameters for a conventional coupled T-lines structure can be expressed as [103]:

$$\begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} = \begin{bmatrix} \frac{-j(Y_{0o}+Y_{0e}) \cot \theta}{2} & \frac{-j(Y_{0o}-Y_{0e}) \cot \theta}{2} \\ \frac{-j(Y_{0o}-Y_{0e}) \cot \theta}{2} & \frac{-j(Y_{0o}+Y_{0e}) \cot \theta}{2} \end{bmatrix} \quad (4.1)$$

where Y_{0o} , Y_{0e} denotes the odd-mode and even-mode admittance, respectively; $\theta = \beta l$ is the electrical length of the coupler.

When two identical capacitors (C) are introduced on both sides of the coupler, the two-port Y-parameters becomes:

$$\begin{bmatrix} Y_{11}' & Y_{12}' \\ Y_{21}' & Y_{22}' \end{bmatrix} = \begin{bmatrix} \frac{-j(Y_{0o}+Y_{0e})\cot\theta}{2} + j\omega C & \frac{-j(Y_{0o}-Y_{0e})\cot\theta}{2} \\ \frac{-j(Y_{0o}-Y_{0e})\cot\theta}{2} & \frac{-j(Y_{0o}+Y_{0e})\cot\theta}{2} + j\omega C \end{bmatrix} \quad (4.2)$$

Eq. (4.2) can be converted into S-paramaters according to the method introduced in [104];

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} \frac{1+Z_0^2(Y_{0o}\cot\theta-\omega C)(Y_{0e}\cot\theta-\omega C)}{\Delta Y} & \frac{jZ_0(Y_{0o}-Y_{0e})\cot\theta}{\Delta Y} \\ \frac{jZ_0(Y_{0o}-Y_{0e})\cot\theta}{\Delta Y} & \frac{1+Z_0^2(Y_{0o}\cot\theta-\omega C)(Y_{0e}\cot\theta-\omega C)}{\Delta Y} \end{bmatrix} \quad (4.3)$$

with

$$\Delta Y = Z_0^2[\omega C(Y_{0o} + Y_{0e})\cot\theta - \omega^2 C^2 - Y_{0o}Y_{0e}\cot^2\theta] + jZ_0[2\omega C - (Y_{0o} + Y_{0e})\cot\theta] + 1 \quad (4.4)$$

As such, the coupling phase (ϕ) can be expressed as:

$$\phi = \frac{\pi}{2} - \tan^{-1} \left\{ \frac{2\omega Z_0 C - Z_0(Y_{0o} + Y_{0e})\cot\theta}{1 + Z_0^2[\omega C(Y_{0o} + Y_{0e})\cot\theta - \omega^2 C^2 - Y_{0o}Y_{0e}\cot^2\theta]} \right\}. \quad (4.5)$$

when the impedance of both ends are perfectly matched ($Z_0^2 Y_{0o} Y_{0e} = 1$), a zero coupling phase condition ($\phi = 0$) is satisfied in (4.5) with

$$\left[\cot(\beta l) - \frac{\omega C}{Y_{0e}} \right] \left[\cot(\beta l) - \frac{\omega C}{Y_{0o}} \right] = 1. \quad (4.6)$$

The required physical length l can be derived as

$$l = \frac{1}{\omega\sqrt{\mu\varepsilon}} \cot^{-1} \left\{ \frac{\omega C(Y_{0o} + Y_{0e})}{2Y_{0o}Y_{0e}} + \sqrt{1 + \left[\frac{\omega C(Y_{0o} - Y_{0e})}{2Y_{0o}Y_{0e}} \right]^2} \right\} \quad (4.7)$$

Note that (4.7) is obtained as the minimum positive solution of (4.6), which provides the smallest feature size of zero-phase-coupler in the practical IC layout.

With inter-digital configuration in layout, as shown in Fig. 4.1(b), lower coupling loss can be achieved [105]. The coupling coefficient in zero-phase mode $|S_{21_{ZP}}|$ can

be derived from (4.3) and (4.6):

$$|S21_{ZP}| = \left| \frac{(Y_{0e} \cot \theta - \omega C) - (Y_{0o} \cot \theta - \omega C)}{(Y_{0e} \cot \theta - \omega C) + (Y_{0o} \cot \theta - \omega C)} \right|. \quad (4.8)$$

which can be optimized for start-up condition by a higher odd-mode admittance (Y_{0o}) or a lower even-mode admittance (Y_{0e}).

Clearly, a low loss can be obtained by a much smaller physical length l and a lower Y_e for multiple ZPC based oscillator unit-cells under the zero-phase condition. Compared to the conventional coupler design with T-line by single strip on each side, the proposed ZPC structure can simultaneously increase Y_{0o} and reduce Y_{0e} , which can be further optimized based on the relation of coupler length l vs. loaded capacitances and even-mode admittance as shown in Fig. 4.2. The coupling loss needs to be compensated to start oscillation, which means $|S21_{G_m}| \cdot |S21_{ZP}| > 1$, where $S21_{G_m}$ is the equivalent gain of shunt negative conductance from active devices.

Figure 4.2: The calculated l for zero coupling phase vs. loaded capacitance and even-mode characteristic admittance at 60GHz.

4.3 Zero-phase CON in Close-looped

Coupled oscillator network (CON) [96] is a well-known structure to synchronize output power and reduce phase noise. For a closed-loop CON with N oscillators, the phase shift ($\Delta\phi$) between adjacent oscillators need satisfy the condition of $\Delta\phi = 2k\pi/N$, ($k = 0, \pm 1, \pm 2, \dots$) as illustrated in Fig. 4.3.

$$\text{Loop Phase} = 2k\pi, (k=0, \pm 1, \pm 2, \dots)$$

Figure 4.3: Closed loop coupled oscillator network with center combined output.

The combined output admittance ($Y_{OUT}(\omega_0)$) and current ($I(t)$) of all oscillators can be calculated as:

$$\begin{cases} Y_{OUT}(\omega_0) = \sum_{i=1}^n Y_i(\omega_0) \\ I_{OUT}(t) = I_0 \cdot \sum_{i=1}^n \cos(\omega t + \phi_i) \end{cases} \quad (4.9)$$

where Y_i , I_0 and ϕ_i are the output impedance, the amplitude and phase of the output current from each oscillator unit-cell, respectively. Clearly, $I(t)$ is maximized as $N \cdot I_0$ when all oscillator outputs are in-phase ($2k/N = 0, \pm 1, \pm 2, \dots$). Because Y_{OUT} is also N times larger by parallel connecting N oscillator outputs, the total available output power is N times increased by the CON due to $P_{OUT} = 0.5 \cdot |I_{OUT}|^2 / Y_{OUT}$ when compared to that of a single free running oscillator. However, if the coupling

network is implemented by the conventional T-line, at least an equivalent length of $\lambda/2$ is required with $2k/N = 1$, which is not compact with large loss.

The resulting phase noise ($L(\Delta\omega)$) at frequency offset $\Delta\omega$ can also be improved under the zero-phase condition satisfied for the CON (in $1/f^2$ region) is [106]:

$$L(\Delta\omega) = 10\log \left(\frac{8\pi Z_0 \omega_0^2 \overline{i_T^2}}{NP_{diss} Q_L^2 \Delta\omega^3} \right) \quad (4.10)$$

where $\overline{i_T^2}$ is the squared noise current density; P_{diss} is the power dissipated; and ω_0 is the oscillation frequency. Z_0 and Q_L are the impedance and the quality factor of the coupling T-line, respectively. Ideally, with N oscillator unit-cells coupled, the phase noise is N times smaller compared to the single free-running oscillator. Note that the similar improvement cannot be achieved by spending the same amount power at one single oscillator. Firstly, increasing the supply voltage close to the breakdown voltage has serious reliability issue; secondly, phase noise cannot be reduced when increasing the supply voltage.

4.4 60GHz VCO Design by ZPC-Based CON

4.4.1 Differential ZPC Unit-cell

Fig. 4.4 shows the schematic and layout to implement a differential ZPC by the top most copper layers (M5, M6) and Aluminium layer (AL) with inter-digital coupling topology in 65-nm CMOS process. The average length of coupler is $182\mu\text{m}$. The sizes of transistors are pre-determined by the required output power and the frequency range. Their parasitic capacitances are extracted from post-layout simulation as the load of one ZPC. These capacitances are then incorporated into the EM simulation to satisfy Eq. 4.6. The design satisfies the zero-phase condition at 60GHz as illustrated in Fig. 4.5. In addition, the S11 is smaller than -20 dB and the differential S21 is greater than -0.5 dB at the vicinity of 60 GHz, which confirm a low coupling loss. As a comparison, a conventional coupler using two coupled T-lines with the minimum allowed gap ($1.5\mu\text{m}$) is also simulated. With the same capacitance load (40 fF), the proposed ZPC obtains 2-dB better S21, leading to a lower coupling loss. The propagation constant of the proposed ZPC is illustrated in Fig. 4.6. β is negative before 60 GHz, which presents the left-handed property. A zero β is achieved at 60 GHz leading to the zero-phase coupling condition. At the same time, the attenuation constant α is also minimized at around 60 GHz, resulting in the minimum coupling

loss. As shown in Fig. 4.6, the negative relative permittivity (ϵ_r) realized by the CMOS on-chip ZPC confirms the metamaterial characteristic.

Figure 4.4: Layout of inter-digital differential ZPC with effective electrical length controlled by MOS switches.

4.4.2 ZPC-based Oscillator with Tuning

Since a broadband zero coupling phase is desired, based on (4.6) and (4.7) the frequency tuning can also be achieved by changing the effective length of the coupler. The inductive loading method [102] is applied to achieve the wide tuning range. Two metal loops are formed above the coupler in the aluminum PAD layer (LB). By configuring the on-off status of MOS switches $M_{a1,2}$ and $M_{b1,2}$ shown in Fig. 4.4, the effective electrical length of the coupler is changed, which in turn changes ZPC oscillator frequency as shown below

$$\omega_{0,mn} = \frac{1}{\sqrt{C_{eq}(L_{eq} + mM_A + nM_B)}} \quad (4.11)$$

where $C_{eq}L_{eq}$ defines the maximum operation frequency; $m = 0, 1$, $n = 0, 1$ denotes the modes of configurations; M_A and M_B are the loaded mutual inductances. As such, the minimum FTR without considering varactor tuning can be calculated by

$$FTR = 2 \left[\frac{\omega_{0,00} - \omega_{0,11}}{\omega_{0,00} + \omega_{0,11}} \right] \quad (4.12)$$

Figure 4.5: Simulated magnitude of S11 and S21 in dB and phase of S21 of proposed the proposed coupler structure with comparison of S21 magnitude to the conventional one by coupled T-line.

In order to achieve higher FTR, larger M_A and M_B are required.

The loading capacitance is contributed by the parasitic of the cross-coupled NMOS pair (M_1 and M_2), output buffers (M_3 and M_4) and varactors (D_1 and D_2). M_1 and M_2 provide a negative resistance to compensate the energy loss. The sizes of M_1 and M_2 are optimized with the maximum tuning range. The drain-source current of cross-coupled pair is controlled by a tail current connected NMOS M_5 , of which the biasing can be adjusted. To facilitate output impedance matching and isolate the VCO core from peripherals, common source NMOS M_3 , M_4 are employed as output buffers. The width of M_3 , M_4 are optimized for high output power.

4.4.3 60-GHz Zero-phase-coupled Oscillator Network

The schematic of the distributed zero-phase-coupled VCO network is shown in Fig. 4.7. Four differential ZPC unit-cells are connected in serial with a closed-loop. Their power outputs are combined at the geometry center of the layout. A CPW T-line with 120 μm length and 50 Ω characteristic impedance is used to connect the center output to the RF PADs. The cross-section of CPW T-line is also shown in in Fig.

Figure 4.6: The extracted dispersion diagram and relative permittivity (ϵ_r).

4.7. According to the EM-Simulation at 60 GHz, the CPW T-line has an insertion loss and Q-factor (Q) of 0.1 dB and 15, respectively. Note that all the tail currents of all oscillator unit-cells are controlled by the same current mirror with diode-connected NMOS.

4.4.4 Measurements

As shown in Fig. 4.8, the distributed 4-way ZPC-based CON was fabricated in UMC 65-nm CMOS process with f_T/f_{MAX} of 170/190 GHz. The core chip area is $330\mu\text{m} \times 320\mu\text{m}$ excluding Pads. It was measured on CASCADE Microtech Elite-300 probe station and Agilent PNA-X (N5247A), E5052 source signal analyzer with spectrum swept up to 70 GHz. Bias-T, probe and cable loss are calibrated before the measurements. The proposed VCO consumes 91-mW DC power under 1.2-V power supply. Note that the differential outputs are on different side of the chip. Since

Figure 4.7: Schematic of proposed VCO with differential outputs combined at the center.

the measurement is performed at single-ended RF output with GSG Pads, +3 dB is added to the output power level. In addition, the simulation results are obtained from Cadence Spectre post-layout simulation.

Fig. 4.9 shows the simulated and measured output power of VCO in each mode. Similar to the simulation results, the measured output power of VCO is highest when $S_a = 1.2$ V and $S_b = 1.2$ V. This is mainly due to a higher transconductance of NMOS transistors at a lower frequency. The proposed VCO achieves the highest output power of 2 mW (3 dBm) at $V_{Tune} = 0$ condition, where the oscillation frequency is 58.1 GHz. The according maximum DC-RF efficiency is 2.2%, which is defined by P_{OUT}/P_{DC} , where P_{OUT} and P_{DC} are the output power and DC power consumption of VCO, respectively. Fig. 4.10 shows four modes of VCO with different frequency bands obtained from both simulation and measurement, which are generated by switching the inductive loadings from S_a and S_b . The entire 60-GHz band under IEEE 802.15.3c standard is completely covered in both simulation and measurement. Note that the measured FTR of proposed VCO is 15.8% from 58.1 to 68.1 GHz, which is slightly lower than the simulation result of 16.8% from 58.4 GHz to 69.1 GHz. Fig. 4.11 shows the measured and simulated phase noise at 60 GHz output frequency. The measured phase noise correlates very well with the simulation results. A -116.7-dBc/Hz phase noise is observed at 10-MHz offset, which about 2.5-dB lower than the simulation result at the same frequency offset.

Figure 4.8: Die micrograph of the fabricated 60GHz VCO chip in 65nm CMOS.

The performance of the proposed VCO is summarized in TABLE 4.1 with comparison of other similar designs at 60 GHz. It can be observed that the proposed design has the highest output power of +3 dBm, the highest power efficiency of 2.2% and the widest FTR of 15.8%. Note that the phase noise of the proposed VCO is very close the best reported result of -118.8 dBc/Hz at 10-MHz offset in [107]. However, the output power of the proposed VCO is almost 10 times (9.6-dB) higher. This leads to the state-of-art figure of merit (FOM) and figure of merit with tuning range (FOMt)

Figure 4.9: Measured and simulated output power and power efficiency of proposed VCO over entire 60GHz band.

Figure 4.10: Measured and simulated VCO FTR under various switches configurations to cover 58.3-64.8GHz continuously.

as well, which are defined by the following equations:

$$\begin{cases} FOM = PN(\Delta f) - 20\log\left(\frac{f_{osc}}{\Delta f}\right) + 10\log\left(\frac{P_{diss}}{1mW}\right) \\ FOMt = PN(\Delta f) - 20\log\left(\frac{f_{osc}}{\Delta f} \times \frac{FTR}{10}\right) + 10\log\left(\frac{P_{diss}}{1mW}\right) \end{cases} \quad (4.13)$$

where $PN(\Delta f)$ is the phase noise at the offset frequency Δf , f_{osc} is the oscillation frequency and P_{diss} is the DC power consumption in mW. The highest output power and lowest phase noise both confirm the feasibility of applying the proposed ZPC in mm-wave IC designs. The maximum power density of the proposed VCO is 18.4 mW/mm², which is more than 8 times higher than the previous VCO design in [107]. Note that power density is defined as the output power generated in unit chip area (P_{OUT}/A_{CORE}).

Figure 4.11: Measured and simulated VCO phase noise at 60GHz.

4.5 140GHz Injection-locked Signal Source by ZPC-based CON

Fig. 4.12 shows the block diagram of the proposed 140-GHz signal source, of which the core is a 70-GHz CON with four zero-phase-coupled oscillator unit-cells. Since the output signals after frequency doublers are still in-phase, they are directly combined at the center of CON to generate a four times higher output power. The oscillation frequency of CON is controlled by the injection locking method. Compared

TABLE 4.1: PERFORMANCE COMPARISON OF STATE-OF-THE-ART VCO DESIGNS AROUND 60 GHz

Parameters	[93]	[85]	[41]	[107]	This Work
Technology	0.13- μ m CMOS	32-nm CMOS SOI	65-nm CMOS	90-nm CMOS	65-nm CMOS
f_{osc} (GHz)	56.5	102.2	40	57	63.1
FTR (%)	10.3	4.1	10.5	14.3	15.8
Phase Noise @10MHz (dBc/Hz)	-108 @10M	-100.8 @10M	-85 @10M	-118.8 @10M	-116.7 @10M
Output Power (dBm)	-18	-30.7	-13	-6.6	+3
Power Efficiency (%)	<0.16	0.013	<0.1	1.5	2.2
FOM (dBc/Hz)	-173.1	-172.45	-162	-184.3	-172.9
FOM _t (dBc/Hz)	-173.4	-164.75	-162.4	-187.4	-177.3
A_{CORE} (mm ²)	0.06	0.0014	-	0.1	0.11
P_{OUT}/A_{CORE} (mw/mm ²)	0.26	0.6	-	2.2	18.4

to the direct frequency control by a 70-GHz phase lock loop (PLL) with the bulky and power hungry frequency dividers, injection locking method has higher power and area efficiency. In this work, the 70-GHz injection signal is obtained by doubling the frequency of a 35-GHz reference input, which can be easily generated by an on-chip or off-chip signal generator. The design of each circuit block is shown in following section.

4.5.1 Zero-phase Oscillator Unit-cell at 70GHz

Fig. 4.13 shows the schematic and layout of on-chip MPW based oscillator unit-cell with coupled T-line implemented in the top most copper layer (M8) and parasitic capacitances from transistors in 65-nm CMOS process. Here an inter-digital coupling topology is deployed to largely increase the magnetic coupling inside each unit-cell. Both input and output of the unit-cell are on the same side due to the dumbbell shape routing with an effective length of 40 μ m. Switch-controlled inductive loadings by Sa and Sb are applied to increase the number of the available zero-phase modes of unit-cell as well as the tuning range of CON. The unit-cell EM-simulation results and the dispersion diagram extracted by method introduced in [108] without any inductive loadings are shown in Fig. 4.14 and 4.15, respectively, where a very small insertion loss of 0.4dB is observed in zero-phase mode at 70 GHz. Note that a parasitic capacitance of 40fF from active devices is also considered in the simulation. Moreover,

Figure 4.12: Block diagram of proposed 140GHz signal source with center combined output from four 70GHz zero-phase-coupled oscillator unit-cells.

the metamaterial properties of proposed on-chip unit-cell design are verified by a similar dispersion diagram to Fig. 2.2(b) except the loss induced non-zero α at zero-phase mode.

4.5.2 70GHz Zero-phase Coupled Oscillator Network

The schematic of the 70-GHz CON is shown in Fig. 4.16. Four MPW based oscillator unit-cells are serially connected in a closed-loop form. Due to the strong in-phase inductive coupling inside each zero-phase oscillator unit-cell, the differential output signals at locations A, B, C and D have the same phase, magnitude and frequency, which are locked to the injected 70-GHz reference signal with largely amplified strength. The oscillation signal is generated by compensating the energy loss in each unit-cell with a negative resistance formed by cross-coupled NMOS pair (M1 and M2). Usually larger M1 and M2 are preferred to ensure the oscillation condition, but the available output power will be correspondingly reduced. Additionally, in order to reduce the impacts of the process variation, a central symmetrical layout is deployed and all active devices are placed as closed as possible to the geometrical center of CON.

Figure 4.13: Schematic of on-chip ZPC-based oscillator unit-cell at 70-GHz band with inter-digital coupled T-line and switch-controlled inductive loadings.

4.5.3 70GHz to 140GHz Output Frequency Doublers

Fig. 4.17 shows the schematic of four 70-GHz to 140-GHz push-push frequency doublers with center combined output. The 70-GHz differential output signals at A, B, C and D are coupled to the push-push frequency doubler by 28fF DC-block capacitors. Therefore, all the frequency doublers can be externally biased to the threshold level (VG1) to maximize the frequency conversion efficiency. The resulting four in-phase 140-GHz output signals are directly tied together to generate a high power output signal at the center with a combined output impedance of 50 Ω . Moreover, a LC resonator based AC-GND is applied to reduce the output leakage of the 70-GHz fundamental signal. A CPW T-line with 210- μ m length and 50- Ω characteristic impedance is used to connect the center output to the RF PADs. The cross-section of CPW T-line is also shown in in Fig. 4.17. According to the EM-Simulation at 140 GHz, the CPW T-line has an insertion loss and Q-factor (Q) of 0.45 dB and 13.5, respectively.

Figure 4.14: EM simulation results of 70-GHz ZPC with inter-digital coupled T-line.

4.5.4 35-GHz to 70-GHz Input Reference Frequency Doubler

Fig. 4.18 shows the schematic of the 35-GHz to 70-GHz Reference Frequency Doubler. One transformer based balun is deployed to generate a differential 35-GHz reference signal to drive M3 and M4, which also have the threshold level biasing. However, transformer based balun suffers from the outputs mismatch above 60 GHz. As such, another Marchant balun with inter-digital coupling is deployed at 70 GHz to have balanced differential outputs as well as low insertion loss. As verified by EM simulation in Fig. 4.19, the proposed Marchant balun has an average intrinsic loss of 1.1 dB at 70 GHz. The magnitude and phase mismatches at 70 GHz are only 0.4 dB and 4 degrees, respectively. The 70-GHz differential signal is then injected into the CON by a common source buffer stage (M5 and M6). Fig. 4.20 shows the post-layout simulation results of the entire frequency doubler with 5-dBm reference power. The conversion gain is above -15 dB in 33-39 GHz. Moreover, a good input matching is observed with S11 smaller than -6 dB in 33-37.5 GHz.

Figure 4.15: Extracted dispersion diagram of 70 GHz ZPC with inter-digital coupled T-line.

Figure 4.16: Schematic of injection-locked 70-GHz CON with 4 oscillator unit-cells.

4.5.5 Measurements

As shown in Fig. 4.21, the proposed injection-locked 140GHz signal source is implemented in GlobalFoundries 65-nm CMOS RF process with f_T/f_{MAX} of 180/530 GHz. The die area including DC and RF pads is $750 \times 550 \mu^2\text{m}$, and the CON core area is 0.13 mm^2 . It was measured on a probe station with a 6-pin DC probe for biasing, a normal GSG probe for reference signal input and a D-band GSG to waveguide probe for output, which is connected to the R&S FSUP signal source analyzer with a

Figure 4.17: Schematic of four 70-GHz to 140-GHz push-push frequency doublers with center combined output.

Figure 4.18: Schematic of 35GHz to 70GHz frequency doubler with inter-digital Marchand Balun at 70 GHz.

Figure 4.19: EM simulation results of the proposed Marchant balun at 70GHz.

D-band waveguide harmonic mixer. Operating from a 1.2-V power supply, the CON core of signal source consumes 145 mW, while the input frequency doubler consumes another 3.8 mW. Note that the simulation results are obtained from Cadence Spectre post-layout simulation.

TABLE 4.2: PERFORMANCE COMPARISON WITH RECENTLY PUBLISHED THZ SIGNAL SOURCE

Parameters	[84]	[92]	[87]	[91]	This Work
Technology	65-nm CMOS	65-nm CMOS	SiGe	65-nm CMOS	65-nm CMOS
f_{osc} range (GHz)	113.6-118.8	100-110	125-138	159-169	127-140
P_{OUT} (dBm)	-3.5	4.5	3	1	5.4
Power Efficiency (%)	2.2	5.2	4	1.38	2.4
FTR (%)	4.4	9.5	9.8	6.1	9.7
A_{CORE} (mm ²)	0.21	0.18	0.21	0.1	0.13
P_{OUT}/A_{CORE} (mw/mm ²)	2.1	15.6	9.5	4.16	26.9

Fig. 4.22 shows the output power obtained from both measurement and simulation. By controlling the inductive loading modes of MPW unit-cells, the entire tuning range of signal source from 127 to 140 GHz is covered by three available bands. Compared

Figure 4.20: Post-layout simulation results of the 35-GHz to 70-GHz frequency doubler.

Figure 4.21: Die micrograph of the fabricated THz source in CMOS.

Figure 4.22: Measured and simulated output power of signal source with three different inductive loading modes.

to the simulation results, the measured output power is 0.5 ~ 2.5-dB higher in 128 ~ 132 GHz and 1.5 ~ 3-dB lower in 134 ~ 140 GHz. This is probably due to the process variation. The maximum output power of 5.4 dBm is observed at 132 GHz in mode 2 with a DC-RF efficiency of 2.4%. Fig. 4.23 shows the measured and simulated spectrum of 132-GHz output signal when locked to a 5-dBm reference signal at 33 GHz. Note that the simulated spectrum is offset to the same output signal power level at 132 GHz. Compared to the simulation results, a raised noise floor is observed in the measurement within ± 25 MHz of 132 GHz. This is probably contributed by the noise coupled from DC power supplies, because there are not any de-coupling capacitors in the DC probe. In such case, a phase noise of -104.9 dBc/Hz is measured at 25-MHz offset. Moreover, the maximum power density of the proposed signal source is 26.9 mW/mm², which is defined as the output power generated in unit chip area (P_{OUT}/A_{CORE}). The performance of the proposed signal source is summarized in TABLE 4.2 with comparison to the recent state-of-the-art THz source designs in both CMOS and SiGe processed. It can be observed that the proposed design has the highest output power and power density.

Figure 4.23: Measured and simulated output spectrum of the proposed signal source at 132GHz.

4.6 Conclusion

The CMOS high output power signal sources are demonstrated from mm-wave to THz by coupled oscillator network with ZPC based oscillator unit-cells, which helps largely increase the output power and efficiency by the in-phase power combination. The fabricated 60-GHz VCO in 65-nm CMOS has a compact core chip area of 0.11 mm², and it is measured with 2-mW output power, -116.7-dBc/Hz phase noise at 10MHz offset, and 15.8% frequency tuning range (FTR) centered at 63.1 GHz. The fabricated 140GHz signal source with injection locking in 65-nm CMOS has a compact core chip area of 0.13 mm², and it is measured with 3.5-mW peak output power, 2.4% power efficiency, 26.9-mW/mm² power density and 9.7% FTR centered at 133.5 GHz. In the following section, CMOS based THz signal detection will be discussed.

Chapter 5

CMOS THz Signal Detection

5.1 Introduction

Recently, CMOS-based THz detectors have been developed [23–25, 109–111]. One can thereby develop a low-cost, portable and large-arrayed THz imaging system in CMOS. One transmission-typed design is shown in Fig. 5.1, where each THz image pixel consists of a receiver and an antenna. However, the THz radiation signal strength is usually weak when generated by CMOS and it will be further attenuated by absorption and diffraction during the propagation. The main challenge is to design a high-sensitivity receiver that can compensate the weak signal source and the path propagation loss with both narrow band or wide band approaches. Moreover, a large-arrayed receiver is desired with improved spatial resolution and also image capturing speed, requiring a compact design of each THz image pixel [112].

The recent super-regenerative receiver (SRX) [110, 113] topology can achieve high-sensitivity within narrow band, which is desired for THz imager that has relative low data rate. As depicted in Fig. 5.2(a), the core of a SRX is a quench-controlled oscillator, which consists of a resonator with positive feedback network to realize an oscillatory amplification. When a periodic quench-control signal is applied, the average of detected signal envelope is amplified for the injected RF signal from LNA. One compact and high-sensitivity SRX for a THz imager requires a compact and high quality factor (Q) resonator. Note that the Q of the traditional LC-tank based resonator [114] has significant performance degradation with large area at THz frequency region.

Recently, a metamaterial-based resonator has been explored in [115] to improve the Q with compact area at mm-wave frequency region. A split-ring-resonator (SRR) can be designed in CMOS process top-metal layer to provide negative permeability (μ) for

Figure 5.1: CMOS THz imager array.

mm-wave propagation. When loading SRR to a host transmission-line (TL-SRR), the integrated structure becomes a non-transmission medium with single-negative property ($\mu \cdot \varepsilon < 0$) in the vicinity of resonance frequency. A sharp stop-band is thereby formed such that the incident mm-wave can be perfectly reflected at SRR load with a stable standing-wave established in the host T-line [115]. Compared to the traditional LC-tank based resonator, TL-SRR has stable EM-energy storage within a compact area, which results in much higher Q factor. As such, it becomes relevant to study the CMOS on-chip SRR for the compact and high-sensitivity SRX design of the THz imager.

In this chapter, firstly, the design of on-chip metamaterial resonator are explored based on differential T-line loaded with SRR and CSRR beyond 70 GHz. As a demonstration of idea, two oscillators based on SRR and CSRR resonators are implemented with 65nm CMOS process at 76 GHz and 96 GHz, respectively. The state-of-the-art performance shows that the phase noise and FOM of SRR achieves -108.8 dBc/Hz and

Figure 5.2: (a) Block diagram of super-regenerative receiver; (b) impact of resonator Q-factor to receiver sensitivity.

-182.1 dBc/Hz (@10 MHz offset), respectively. The power is reduced dramatically to 2.7 mW compared to the existing designs on SWOs [116–118]. And the CSRR oscillator shows the-state-of-art phase noise of -111.5 dBc/Hz (@10-MHz offset) and FOM of -182.4 dBc/Hz. Secondly, two super-regenerative receivers with quench-controlled metamaterial high-Q oscillators by TL-CSRR and TL-SRR are demonstrated with improved sensitivity over traditional LC-tank resonator based designs at 96 GHz and 135 GHz, respectively. With sharp stop-band introduced by the metamaterial resonators, high-Q oscillatory amplifications are achieved. The 96-GHz DTL-CSRR based SRX has a compact core chip area of 0.014 mm², and it is measured with power consumption of 2.8 mW, sensitivity of -79 dBm, noise figure (NF) of 8.5 dB, and noise equivalent power (NEP) of 0.67 fW/ $\sqrt{Hz}$. The 135-GHz DTL-SRR based SRX has a compact core chip area of 0.0085 mm², and it is measured with power consumption of 6.2 mW, sensitivity of -76.8 dBm, NF of 9.7 dB, and NEP of 0.9 fW/ $\sqrt{Hz}$. The proposed SRXs have 2.8-4 dB sensitivity improvement and 60% smaller core chip area when compared to the conventional SRX with LC-tank based resonator at similar frequencies.

5.2 Oscillators Design with SRR and CSRR

To demonstrate the high-Q performance of the proposed metamaterial resonator and its improvement in phase noise, Two MMIC oscillators based on metamaterial

resonators have been designed in 65-nm CMOS (STM 7-metal-layer). The first one is operated at 76 GHz using the differential T-line loaded with SRR (DTL-SRR), and the second one is operated at 96 GHz using the differential T-line loaded with SRR (DTL-SRR). Note that the design of slow-wave shielding is implemented for both MMIC oscillators with loss reduction. The slow-wave shielding strips are designed by the bottom metal layer M1 with both width and pitch of 1 μm . The two MMIC oscillators are designed and verified with Agilent ADS Momentum for EM simulation and Cadence Spectre for oscillator circuit simulation.

5.2.1 Differential TL-SRR Resonator Design

5.2.1.1 Stacked SRR Layout

The on-chip SRR can be implemented in a stacked fashion with on-chip multi-layer interconnect [47]. As shown in Fig. 5.3(a), one SRR unit-cell is realized by the top two metal layers stacked alternatively, considering a trade-off among resonant frequency, area and loss. When its size is fixed, S_{21} of TL-SRR with different stacked layers is shown in Fig. 5.3(b). It is found that more stacked layers result in lower resonant frequency, but suffer from lower Q at the same time. With the increased resonant frequency, TL-SRR reveals a steeper and higher rejection property, which means a higher Q . Thus TL-SRR shows the potential application for on-chip MMIC designs.

Figure 5.3: (a) Stacked SRR unit-cell designed by metal layers of M7 and M6; (b) S_{21} simulation results with different stacking methods.

Fig. 5.4(a) shows a differential T-line with stacked on-chip SRR (DTL-SRR) in CMOS process, of which the cross-section is illustrated in Fig. 5.4(c). The two loaded SRR unit-cells are excited by the axial magnetic field generated by the host

Figure 5.4: Geometries of resonators with slow-wave shielding: (a) differential T-line loaded with stacked SRR, (b) T-line based standing-wave resonator, and (c) cross-section of BEOL.

T-line. It has the following advantages in Q improvement. Firstly, as the SRR-load is metamaterial with stop-band property, it results in large impedance with the open circuit condition formed. Thus EM energy can be perfectly reflected in the host T-line. Secondly, the differential design provides local ground to reduce EM loss and enhance the EM-energy coupling. For example, the magnetic field generated by the differential T-line is equidirectional and superimposed when applied to the two SRR unit-cells. Thus a stronger coupling between T-line and SRR is achieved with larger mutual capacitance and mutual inductance, which can store more EM-energy with less EM-energy leakage into the substrate. Due to the stronger EM coupling, the DTL-SRR needs less number of SRR unit-cells than STL-SRR when the same rejection property is achieved. This makes the DTL-SRR achieve higher area efficiency as well. To strengthen the coupling between T-line and SRRs, a shortest distance (or gap) between SRRs and T-line is selected with the consideration of the process limitation (1.5 μm in STM 65nm CMOS). Lastly, floating metal shielding is also employed in this design to further reduce the substrate loss.

5.2.1.2 Comparison with Single-ended TL-SRR Resonator

In the following, detailed analysis for the enhancement of Q factor is shown with comparison between the DTL-SRR and STL-SRR. Assuming both terminals of SRR unit-cell observe the same characteristic impedance (Z_0). The reflection coefficient

can be estimated at the position TL-SRR unit cell by

$$\Gamma = \frac{R'_s}{R'_s + 2Z_0} = \frac{k^2}{k^2 + \frac{2Z_0L_s}{R_sL}}. \quad (5.1)$$

One can have two observations from (5.1). Firstly, if the Q factor of SRR is sufficiently high that $k^2 \gg 2Z_0L_s/R_sL$, Γ is approaching unity, which means a perfectly reflection of EM-wave at the SRR-load. Secondly, Γ increases with k for a given SRR with a finite Q. Thus improving k is the means to enhance the EM-energy reflection efficiency. Note that the coupling coefficient k is often limited by the geometry mismatch between T-line and SRR.

As a result, in order to have a high-Q DTL-SRR design, one needs to have the reflection coefficient Γ as high as possible. One can observe from (5.1) that Γ increases with the coupling coefficient k between SRRs and T-line, In the single-ended T-line as shown in Fig. 5.5(a), the magnetic flux can not be fully covered between the SRR and T-line. This is illustrated in Fig. 5.5(c) as there is partial of the magnetic flux leaked to the open space regardless the distance between SRR and T-line. In contrast, the differential T-line shown in Fig. 5.5(b) does not have this limitation. As one can see from Fig.5.5(d), it is possible to have SRR fully cover the magnetic flux generated by the differential T-line. Thus, high EM coupling coefficient can be achieved with a high Γ for the DTL-SRR structure than the STL-SRR structure.

To further validate the high-Q of the DTL-SRR, EM simulation (Agilent ADS momentum) is performed for STL-SRR and DTL-SRR structures shown in Fig. 5.5(a) and (b). The conductivity of top most metal layers M6 and M7 are 4.6×10^7 S/m, the metal layers M1~M5 are 4.1×10^7 S/m and the silicon substrate is 10 S/m according to the 65-nm CMOS process files. The simulation result of reflection coefficient (Γ) is plotted against different gap sizes in the Smith Chart as shown in Fig. 5.6. Note that the resonance happens when the imaginary part of Γ equals to zero. One can observe the reflection coefficient of DTL-SRR at resonance frequency is much higher than that of STL-SRR. Moreover, the reflection coefficient is increased for a smaller gap size. For example, at the minimum gap of 2 μm that is allowed by the design rule, the reflection coefficient of the differential T-line is 10.6% higher than that of single-ended T-line. Since the minimum gap size is limited by the design rule, the maximum reflection coefficient one can obtain is around 0.9. The Q factor for both resonators are also compared by the reflection coefficient as shown in Fig. 5.7. As discussed, a high reflection coefficient of DTL-SRR can be directly transferred into a high Q. One can observe that the Q of DTL-SRR is around 20 ~ 40% higher than

Figure 5.5: T-line based SRR excitation: (a) single-ended approach, (b) differential approach, (c) magnetic field distribution of single-ended approach, and (d) magnetic field distribution of differential approach.

that of STL-SRR with the same gap size.

5.2.1.3 Comparison with Standing-Wave Resonator

The proposed DTL-SRR resonator is further compared with the standing-wave resonator using coplanar stripline (CPS). As shown in Fig. 5.4, they are both designed under the same resonance frequency and are also provided with floating metal shielding to reduce substrate loss.

The optimization of the two structures is conducted with the full-wave EM simulator (Agilent Momentum). As for DTL-SRR based metamaterial resonator, the stacked SRR unit-cell is designed with the top two metal layers (M7, M6). M7 and M5 are used for the design of the host T-line and the floating metal strips for shielding of the two resonators, respectively. The sizes of T-line, SRR and floating metal strips are carefully selected to obtain the desired frequency. Moreover, for CPS based standing-wave resonator, its Q factor also depends on the width and the separation of the T-line, the width of the floating metal strip and the spacing between two adjacent floating metal strips. Due to the parasitic capacitance of the cross-coupled NMOS transistors and the layout-dependent parasitic effect, the physical length of CPS is shorter than the ideal length of $\lambda/4$. The detailed physical sizes are shown in Fig. 5.4 and one can observe that the use of SRR has 40% area reduction than the use of CPS.

Figure 5.6: Simulated Γ plot in smith chart of SRR/T-line unit cell at resonance.

Note that the Q of one resonator can be described by

$$Q = \omega \frac{\text{Average_energy_stored}}{\text{Energy_loss/second}}. \quad (5.2)$$

As such, one can compare the Q factors of the DTL-SRR with the standing-wave resonator as follows. Firstly, the smaller size of the DTL-SRR leads to a lower substrate loss, and hence the denominator above decreases. Secondly, for the DTL-SRR, the strong EM-energy coupling between the SRR and T-line with perfectly reflection can enhance the energy storage capability with the nominator in (5.2) increased. Thereby, one can expect that a higher quality factor can be achieved by DTL-SRR than the standing-wave resonator, which is further validated by the measured experiment results.

Figure 5.7: Simulated Γ and quality factor (Q) of SRR/T-line unit cell at resonance.

5.2.2 Differential TL-CSRR Resonator Design

It is not feasible to directly deploy the etched CSRR from ground [42] on chip due to the lossy substrate in CMOS process. To realize a low-loss and high-Q implementation for on-chip CSRR, the CSRR can be etched directly on metal layer using signal lines as shown in Fig.2.6. Compared with the previous method [42], CSRRs on metal layer can form much stronger coupling between T-line and CSRR because they are on the same metal layer. More EM-energy can thereby be stored in the resonator and n-turn results in a higher Q-factor.

In this work, a differential CSRR structure is proposed in Fig. 5.8 to provide a compact area. Both inputs are designed on the same side to provide an ac-ground for easy dc-supply, which further reduces the potential coupling to the lossy substrate. Note that the metamaterial property of the proposed differential T-line loaded CSRR and its high-Q feature can be illustrated through simulation as follows.

For example, the metamaterial property can be calculated from S-parameters by

Figure 5.8: On-chip differential T-line loaded with CSRR.

[119, 120]

$$\begin{aligned}
 \cos(nkd) &= \frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}}, \\
 z &= \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}}, \\
 \varepsilon &= \frac{n}{z}, \mu = nz.
 \end{aligned} \tag{5.3}$$

where n is the refractive index, z is the wave-impedance, k is wave-factor, and d is the physical length. Because the metamaterial is considered as a passive medium, the signs of n and z in (5.3) can be determined by two requirements: $\Im(n) \geq 0$ and $\Re(z) \geq 0$, where $\Im(n)$ and $\Re(z)$ denote the imaginary part and real part of n and z , respectively.

Based on (5.3), one can characterize the metamaterial resonator as follows. The proposed DTL-CSRR structure in Fig. 5.8 is implemented on chip with resonance frequency biased around 100 GHz. ADS Momentum is used for the EM simulation to obtain the S-parameters. As shown in Fig. 5.9, a negative μ can be observed within a narrow band near the resonance frequency. As stated earlier, the negative μ and positive ε create the electric plasma, where the propagating EM-wave become evanescent waves and is largely reflected backward. The deep rejection frequency band with a sharp cut-off can be viewed from S12 plot in Fig. 5.9, which corresponds to a high-Q performance.

The Q-factor can be estimated from the simulation by $Q = f_0/BW_{-3dB}$, where

f_0 is the resonance frequency and BW_{3dB} is the bandwidth. As such, the obtained Q-factor is 65, which is much higher than the normal Q value by a resonator composed of LC-tank at similar frequency, around 30 as indicated in [121].

Figure 5.9: EM characterization of the proposed differential CSRR resonator.

Figure 5.10: Voltage distribution of the DTL-SRR based resonator.

5.2.3 76GHz Oscillator Design with Differential TL-SRR Resonator

Firstly, as shown in Fig. 5.10, DTL-SRR based metamaterial oscillator has the property of position-dependent voltage-current amplitudes. If a cross-coupled transistor pair is connected to the opened ends of the two striplines in the host T-lines, an oscillation can be sustained at the specified frequency according to the length of the striplines. The incident-wave energy is injected by the cross-coupled inverters propagates in forward waves along the T-line toward the short point; the energy is reflected at SRR load; and the reverse-wave has a superposition of the incident-wave and leads to a resonance if in phase. Stronger wave reflection means less loss and higher Q of the resonator for oscillation.

As a result, with the use of the proposed metamaterial resonator by DTL-SRR, a 76GHz oscillator is designed. As shown in Fig. 5.11, a pair of cross-coupled pair of NMOS transistors is deployed as the negative resistor to compensate the energy loss in the resonator. Source-follower output buffers are implemented with off-chip bias T to save area. The supply-voltage for the resonator is provided from the middle point of the T-line short-circuit termination, which is a differential zero ac-voltage point. External current source is fed into the chip through a dc-PAD and mirrored to the core circuit of oscillator to control the power consumption. In addition for the comparison purpose, the standing-wave resonator by coplanar stripline (CPS) under the same resonant frequency is also implemented in the same chip with geometries shown in Fig. 5.4(b). For a fair comparison, the proposed two oscillators have the same size and layout designs of cross-coupled NMOS transistors and output buffers, except for resonators. The channel length of transistor is chosen to the minimum 60 nm allowed by the technology to reduce the parasitic effects, while the width is 6 μm (each finger width is 1 μm) for both the cross-coupled pair and the source-follower output buffers. The detailed physical sizes in Fig. 5.4 show that the use of SRR has 40% area reduction than the use of CPS.

5.2.4 96GHz Oscillator Design with Differential TL-CSRR Resonator

Next, a 96GHz oscillator is also implemented in the same CMOS 65nm process with the use of metamaterial resonator by DTL-CSRR. Similarly, the loss from the resonator is compensated by a cross-coupled pair of NMOS transistors, as shown in Fig. 5.12. To obtain the maximum f_{MAX} of NMOS transistors, the individual finger

Figure 5.11: Schematic of the DTL-SRR based oscillator.

width is designed to be $1\ \mu\text{m}$ [122], and the total finger number is designed to be 8 to sustain the oscillation on slow-corner while minimizing parasitic capacitance. In order to isolate the oscillator from the peripheral circuits and also to provide enough output power, the output is designed together with on-chip buffer and RF choke. It is composed of quarter-wavelength slow-wave T-line and de-coupling capacitor. All four transistors in the circuit are self-biased. The dc-supply voltages for the core oscillator and buffer-stage are provided separately to identify the individual current consumption. Note that the 1-metal-layer design of resonator used for demonstration in Fig. 2.6 is modified to be a 2-metal-layer design for oscillator implementation. Other

Figure 5.12: Circuit diagram of the 96GHz CMOS oscillator with the use of the proposed metamaterial resonator by CSRR structure.

than the benefit of size reduction, the stacked structure is also expected to improve the Q -factor. Because the stacked two CSRRs are in the opposite direction, they are excited in the odd mode at resonance by the E-field in between, which avoids the E-field to penetrate through the substrate. As such, the impact of the lossy substrate to the resonator Q -factor can be further reduced.

Figure 5.13: Equipment setup for W-band 75~110GHz phase noise measurement.

5.2.5 Measurements

The proposed 76-GHz SRR and 96-GHz CSRR oscillators were both implemented in STM 65-nm CMOS RF process with f_T/f_{MAX} of 170/230 GHz. As shown in Fig. 5.13, the RF and DC signals are connected through a CASCADE Microtech Elite-300 probe station. The single-ended output of the chip is connected to a phase-noise analyzer FSU-P50 from Rohde & Schwarz (R&S) for the phase noise measurement at millimeter-wave frequency region. To measure the signal frequency in 75 ~ 110 GHz, W-band harmonic mixer FS-WR10 is used for down-conversion. The external Bias-T is required in the measurement for DTL-SRR and SWO based oscillators at 76 GHz.

5.2.5.1 Results of 76GHz Oscillator with DTL-SRR

As shown by the die photo in Fig. 5.14, both DTL-SRR based oscillator and the standing-wave-oscillator (SWO) by coplanar stripline are implemented side by side on the same chip with the same resonant frequency. The sizes excluding RF-PADs is $310 \times 210 \mu\text{m}^2$ (0.06 mm^2) for DTL-SRR, and $310 \times 270 \mu\text{m}^2$ (0.08 mm^2) for SWO. The proposed DTL-SRR oscillator consumes only 2.7 mW from a 1-V power supply, which is slightly higher than the power consumption of SWO of 2.63 mW from the same 1-V power supply. The measured oscillation frequency of the DTL-SRR and SWO oscillators are both observed at 76.1 GHz as shown in the spectrum diagram in Fig. 5.15 and 5.16, where one spur contributed by the harmonic mixer can be observed at 200 MHz away from the oscillation frequency. Here a signal identification function of

spectrum analyzer is used to verify the source of spurs. However, signal power level could be distorted by such identification function, which should be turned off when capturing the measurement results. The output power is -16 dBm by subtracting the loss of cable and Bias-T.

Fig. 5.17 shows the phase noise measurement and simulation results of the DTL-SRR and SWO based oscillators. It can be observed that the variation of phase noise becomes worse as close to the oscillation frequency, induced by the modulation of $1/f$ noise. Thus more stable phase noise results at 10 MHz offset are used for comparison. One can see that the phase noise of DTL-SRR based oscillator is 4.2 dB better than that of SWO at 10 MHz offset at the oscillation frequency. This phase noise improvement is very close to 2.6 dB observed from simulation results. Note that the phase noise of both oscillators are approaching the same level (around -116 dBc/Hz) from 10-MHz to 100-MHz frequency offset. This is because the phase noise level of the oscillator is lower than the noise level of spectrum analyzer in the measurement.

Figure 5.14: Chip photo of the DTL-SRR based oscillator and SWO.

TABLE 5.1 summarizes the performance of the proposed DTL-SRR based oscillator, the SWO realized in the same chip, and the previous oscillators by LC-tank or SWOs at the similar frequency and process. and achieves a phase noise of -108.8 dBc/Hz at 10 MHz offset and a FOM of -182.1 dBc/Hz. Note that the energy loss is reduced due to the higher power efficiency, thus the power consumption of DTL-SRR

Figure 5.15: Spectrum of the 76GHz DTL-SRR oscillator.

Figure 5.16: Spectrum of the 76GHz SWO.

Figure 5.17: Phase noise measurement and simulation results of the 76GHz DTL-SRR oscillator and SWO.

based oscillator is much smaller than the SWO based oscillator while the phase noise is minimized. As a summary, when compared to the existing LC-tank or standing-wave resonator based oscillator [116–118], the phase noise and FOM are improved by 2.7 dB and 7.6 dB on average, respectively. when compared to the standing-wave resonator based oscillator implemented on the same chip, the phase noise and FOM are improved by 4.2 dB and 4.1 dB, respectively.

TABLE 5.1: PERFORMANCE COMPARISON OF STATE-OF-THE-ART OSCILLATOR DESIGNS AROUND 70 GHz

Parameters	[116]	[117]	[118]	This work	This work
Oscillator Design	LC-tank	SWO	SWO	SWO	DTL-SRR
f_{osc} (GHz)	70.2	71.1	76.5	76.1	76.1
Power Supply (V)	1.2	1.2	1.5	1.0	1.0
Power _{core} (mW)	5.4	6	14.3	2.63	2.7
Phase Noise @ 10-MHz offset (dBc/Hz)	-106.1	-104	-108.4	-104.6	-108.8
FOM (dBc/Hz)	-175.8	-173.2	-174.5	-178	-182.1
Technology	65-nm CMOS	65-nm CMOS	65-nm CMOS	65-nm CMOS	65-nm CMOS

5.2.5.2 Results of 96-GHz Oscillator with DTL-CSRR

The die photo of the 96-GHz Oscillator with DTL-CSRR is shown in Fig. 5.18 with area of $430 \times 320 \mu\text{m}^2$ (0.14mm^2) excluding RF-PADs. Both the oscillator core and output buffer are supplied with 1.2-V power supply, of which the current consumption is 6.24 mA and 6.53 mA, respectively. Thus the power consumption for oscillator core circuit is 7.5 mW. Fig. 5.19 shows the measured spectrum when the output frequency is 96.36 GHz. Same as the oscillator with DTL-SRR, there is a 200 MHz offset spur generated by the harmonic mixer. Fig. 5.20 shows the measured phase noise at this frequency of -111.5 dBc/Hz at 10-MHz offset, which is 4.5dB higher than the simulation results. This is probably due to the noise coupled from DC power supply.

Figure 5.18: Chip photo of the DTL-CSRR based oscillator.

As described in TABLE 5.2, the measurement results of the proposed 96-GHz oscillator by DTL-CSRR resonator show the state-of-the-art performance when compared to the recent oscillators designed at 100 GHz using the traditional on-chip LC-tank resonators. Clearly, the proposed high-Q metamaterial resonator structure shows the best phase noise result of -111.5 dBc/Hz at 10-MHz offset and FOM of -182.4 dBc/Hz

Figure 5.19: Spectrum of the 96-GHz oscillator with DTL-CSRR.

at 96 GHz. As a summary, when compared to the existing designs by LC-tank or standing-wave resonator based oscillator [123–125], the phase noise and FOM are improved by 13 dB and 16 dB on average, respectively.

TABLE 5.2: PERFORMANCE COMPARISON OF STATE-OF-THE-ART OSCILLATOR DESIGNS AROUND 100 GHz

Parameters	[123]	[124]	[125]	This Work
Oscillator Design	SWO	LC-tank	LC-tank	DTL-CSRR
f_{osc} (GHz)	104	128	100	96
Power Supply (V)	1.48	1.2	1	1.2
Power _{core} (mW)	25.3	9	30	7.5
Phase Noise @10-MHz offset (dBc/Hz)	-105	-105	-85	-111.5
FOM (dBc/Hz)	-171.3	-177.6	-150.2	-182.4
Technology	130-nm CMOS	90-nm CMOS	90-nm CMOS	65-nm CMOS

Figure 5.20: Phase noise measurement and simulation results of 96-GHz oscillator with DTL-CSRR.

5.3 High Sensitivity THz Receivers with Resonant-type Metamaterial

5.3.1 Fundamentals of Super-regenerative Amplification

Generally, a SRX consists of a quench-controlled oscillator injected by an external signal and an envelope detector. The process of injecting external signal into a quench-controlled oscillator is firstly reviewed to understand the operation of SRX, called super-regenerative amplification (SRA).

5.3.1.1 Equivalent Circuit of SRA

A simplified circuit model of SRA is shown in Fig. 5.21. The resonator is modeled by RLC block, and its oscillation is quench-controlled by a time-dependent negative resistance $-1/G_m(t)$, where G_m is the equivalent conductance determined by the associated active devices. The external signal injected is modeled as a time-dependent current source $I_i(t)$. $V_o(t)$ is the output voltage. The resonance frequency is $\omega_0 = 1/\sqrt{LC}$; the quality factor is $Q_0 = R/Z_0 = 0.5\zeta_0^{-1}$; Z_0 and ζ_0 are the characteristic impedance

Figure 5.21: Simplified equivalent circuit model of super-regenerative amplifier.

and quiescent damping factor, respectively.

Assuming $G_m(t)$ varies much slower than ω_0 such that a quasi-static condition holds in the system to have a time-varying transfer function in s-domain by

$$\frac{V_o(s, t)}{I_i(s)} = \frac{Z_0 \omega_0 s}{s^2 + 2\zeta(t) \omega_0 s + \omega_0^2}, \quad (5.4)$$

where $\zeta(t) = \zeta_0[1 - G_m(t)R]$ is the instantaneous damping factor.

A second-order linear time variant system can be observed from (5.4). By varying $\zeta(t)$, the pole can be shifted between left and right sides of s-plane periodically. In other words, the oscillation starts in SRA when $\zeta(t)$ is negative, and stops when $\zeta(t)$ is positive. Note that (5.4) is only valid when SRA works in linear mode, such that $V_o(s, t)$ is small enough to prevent significant distortion in each quench cycle. Generally, SRA working in linear mode is preferred in the application of millimeter-wave imaging since it has a better sensitivity than that in the logarithmic mode [?, 126].

After Laplace transform, (5.4) can be written as a second-order differential equation in time domain:

$$v_o''(t) + 2\zeta(t)\omega_0 v_o'(t) + \omega_0^2 v_o(t) = Z_0 \omega_0 I_i'(t). \quad (5.5)$$

Assuming the oscillation is fully quenched in each cycle, such that $v_o(t)$ is independent of the previous ones. For a particular quench cycle $t \in (t_a, t_b]$ with $t_a < 0 < t_b$, if $\zeta(t)$ is positive for $t \in (t_a, 0]$ and negative for $t \in (0, t_b]$, (5.5) can be written as [?]:

$$v_o(t) = \frac{Z_0}{s(t)} \int_{t_a}^t I_i'(\tau) s(\tau) \sin[\omega_0(t - \tau)] d\tau \quad (5.6)$$

where $s(t) = e^{\omega_0 \int_0^t \zeta(\lambda) d\lambda}$ is called the sensitivity function, and it reaches maximum when $t=0$; and decays rapidly with t . As a result, the SRA is only sensitive to the

input $I'_i(t)$ in the time window centered at $t=0$ when $\zeta(t)$ turns from positive to negative.

5.3.1.2 Frequency Response of SRA

The frequency response of SRA can be analyzed with convolution model [127]. For a ac input with $I_i(t) = I_0 \sin(\omega_i t + \phi_i)$, the output waveform can be approximated by

$$v_o(t) \approx \frac{Z_0 \omega_i I_0}{2s(t)} |S(\Delta\omega)| \sin(\omega_0 t + \phi_i) \quad (5.7)$$

where $\Delta\omega = \omega_0 - \omega_i$ and $S(\omega)$ is the Fourier transform of $s(t)$. In the application of millimeter-wave imaging, we are more interested in the envelope of v_o , which is

$$Env[v_o(t)] = \frac{Z_0 \omega_i I_0}{2s(t)} |S(\Delta\omega)|. \quad (5.8)$$

Assuming ω_i is very close to ω_0 ($\Delta\omega \ll \omega_i$), a quasi-static condition holds in (5.8) that the frequency response of $Env(v_o(t))$ is determined by $|S(\Delta\omega)|$.

For a typical ramping quench signal with time variant conductance $G_m = \frac{1}{R}(kt+1)$, where k is the normalized ramping slope of G_m with the unit of $1/s$, the instantaneous damping factor is $\zeta(t) = -\frac{k}{2Q_0}t$. Thus the envelope of $v_o(t)$ can be solved by

$$Env[v_o(t)]_{ramp} = \frac{\sqrt{\pi} Z_0 \omega_i I_0}{\Omega_0} e^{\frac{\Omega_0^2}{4} t^2} e^{-\frac{\Delta\omega^2}{\Omega_0^2}}, \quad (5.9)$$

where $\Omega_0 = \sqrt{k\omega_0/Q_0}$ is a constant that determines the frequency response of SRA, e.g., the 3-dB bandwidth of SRA equals $1.177\Omega_0$. As such, one can observe that for the given k and ω_0 , the bandwidth is inversely proportional to Q_0 .

5.3.1.3 Sensitivity of SRA

The sensitivity of SRA is defined as the minimum detected power that means the induced output signal power is the same as its variance:

$$S_{SRA} = P_{min}|_{I_x^2=\sigma_x^2} = \frac{I_0^2 R}{2}|_{I_x^2=\sigma_x^2} \quad (5.10)$$

where I_x is the equivalent induced ac in SRA in response to the ac input I_i , and σ_x^2 is the variance of I_x .

As discussed in [127], for a typical ramp-damping function with normalized ramping slope of k , we have

$$I_x = \frac{I_0 \omega_0 \sigma_s}{2}, \sigma_x^2 = \frac{N \omega_0^2 E_g}{2} \quad (5.11)$$

where $\sigma_s = \sqrt{\frac{2Q_0}{\omega_0 k}}$ is the SRA time constant with a unit of $s/\sqrt{rad}$, $E_g = \sigma_s \sqrt{\pi}$ is the energy of density function, and N is the noise power density with $N = 4K \cdot T \cdot F/R$. Note that K and F denote the Boltzmann constant and noise factor of SRA contributed by active devices, respectively.

As such, the sensitivity of SRX can be found by substituting (5.11) into (5.10):

$$S = 2KTF \sqrt{\frac{k\omega_0}{\pi Q_0}}. \quad (5.12)$$

Note that the receiver noise figure (NF) can be approximated as [110]:

$$NF = \frac{S}{K \cdot T \cdot B}. \quad (5.13)$$

Note that the NF of a SRX is independent of quench signal. For a typical 3-dB bandwidth of the SRX ($B = 1.177\Omega_0$), the NF becomes 0.958 F. In addition, the noise equivalent power (NEP) can be calculated by $S/\sqrt{B}$:

$$NEP = 1.38KTF \sqrt[4]{\frac{k\omega_0}{\pi Q_0}}. \quad (5.14)$$

Note that k is usually determined by the frequency of the quench signal and the sampling rate of one SRA. Therefore, it can be observed from (5.12) and (5.14) that, for a given ω_0 and k , the sensitivity and NEP are inversely proportional to the square-root and the fourth-root of Q_0 , respectively. So the resonator with higher Q will significantly improve the sensitivity within the interested bandwidth for imaging application.

5.3.2 Super-regenerative Receivers Design by Quench-controlled Oscillators

Two SRXs working at 96 GHz and 135 GHz are implemented in CMOS process to demonstrate the advantages of applying quench-controlled Oscillators with metamaterial resonators in super-regenerative receivers (SRX). The fundamentals of quench-controlled oscillator design is introduced first.

5.3.2.1 Quench-controlled Oscillation

High-Q Resonance with Standing Wave In the practical on-chip resonator design with finite Q of SRR or CSRR, the reflection coefficient ($|\Gamma|$) depends on the number of cascading TL-SRR or TL-CSRR unit-cells. Fig. 5.22 shows the circuit level simulation of TL-CSRR at 96-GHz resonance frequency with following observations. First, the reflection coefficient $|\Gamma|$ is more sensitive to the cells number when Q is below 200. Second, $|\Gamma|$ can be improved by cascading more unit-cells.

Figure 5.22: Reflection coefficient of T-line loaded with CSRR unit-cells.

Voltage Controlled Negative Resistance The oscillation can be sustained by compensating the reflection loss ($|\Gamma| < 1$) with a negative resistance. Similarly, a quench-controlled oscillating can be achieved by controlling voltage controlled negative resistance (VCNR), which determines the instantaneous damping factor ($\zeta(t)$) of (5.4) as discussed in Section II. The sensitivity of SRX is also a function of $\zeta(t)$ that determined by VCNR. Usually a cross-coupled NMOS pair is applied for the differential negative resistance design as depicted in Fig. 5.23, where the tail current of the cross-coupled NMOS pair (I_D) can be quench-controlled by another NMOS biased in the saturation region. The equivalent differential negative conductance between nodes

'a' and 'b' can be expressed as below by neglecting the channel-length modulation

$$G_m = \frac{gm_2}{2} = \frac{I_D}{2V_{od2}} \quad (5.15)$$

where gm_2 and V_{od2} are the transconductance and overdrive voltage of cross-coupled NMOS FETs, respectively. Note that I_D can be obtained by

$$I_D = \frac{W_1}{2L_1} \mu_n C_{ox} V_{od1}^2 = \frac{W_2}{L_2} \mu_n C_{ox} V_{od2}^2 \quad (5.16)$$

where W_1 , L_1 and $V_{od1} = V_Q - V_T$ are the channel width, length and overdrive voltage of tail NMOS; W_2 and L_2 and V_{od2} are the channel width, length and overdrive voltage of the cross-coupled NMOS pair; $\mu_n C_{ox}$ and V_T are the process related parameters. As such, (5.15) can be written as a function of V_Q by

$$G_m = \frac{\mu_n C_{ox} (V_Q - V_T)}{4} \sqrt{\frac{2W_1 W_2}{L_1 L_2}} \quad (5.17)$$

One can observe from (5.17) that G_m is linearly controlled by V_Q , of which the slope is determined by the product of W_1/L_1 and W_2/L_2 . Note that the oscillation starts when $1/G_m < R$ and stops when $1/G_m > R$. As such, $(W_1/L_1)(W_2/L_2)$ must be large enough to satisfy the oscillation start conduction ($1/G_m < R$). However, large W_2/L_2 will introduce additional parasitic capacitance, which will be counted into the resonator rank and reduce the oscillation frequency. Moreover, in order to provide sufficient head room for the cross-coupled NMOS pair, W_1/L_1 is selected several times larger than W_2/L_2 .

Figure 5.23: Reflection loss compensation by cross-coupled NMOS pair with controlled tail current.

5.3.2.2 SRX Design by TL-CSRR

Folded Differential T-line Loaded with CSRR TL-CSRR structure cannot be directly employed for the SRX design. Firstly, the single-ended approach will bring large common-mode noise in the oscillator; secondly, cascading more unit-cells will increase area overhead. A folded differential T-line loaded with CSRR (DTL-CSRR) structure is proposed to reduce area by half while doubling the number of the unit-cell [128]. As shown in Fig. 5.24, two cascaded TL-CSRR unit-cells (with CSRR size of $60 \times 60 \mu\text{m}^2$) are folded in the two top most metal layers (M6 and M7).

The S-parameters of the proposed DTL-CSRR structure is verified by EM simulation tool EMX with a parasitic capacitance of 40 fF from transistors. Both ϵ and μ of DTL-CSRR are extracted from the simulation results according to (5.3), which become both complex numbers due to the existence of loss factor induced imaginary parts. The metamaterial property is illustrated by the real parts of ϵ and μ in Fig. 5.25. At the vicinity of 105-GHz resonance frequency, an electric plasmonic medium is formed with $\epsilon < 0$ and $\mu > 0$. A stop-band is thereby formed within a narrow bandwidth of 1.8 GHz, where the Q factor is found to be 58 by $Q = \omega_0/\omega_{3dB}$ from the differential impedance (Z_{diff}) between P1 and P2.

Figure 5.24: Layout for CMOS on-chip implementation of DTL-CSRR for 96GHz SRX.

96-GHz DTL-CSRR-based SRX Fig. 5.26 depicts the schematic of DTL-CSRR-based SRX. DTL-CSRR is firstly connected to a differential negative resistance formed

Figure 5.25: EM-simulation based comparison of DTL-CSRR and LC-tank resonator for CMOS 96GHz SRX design.

by cross-coupled NMOS (M2 and M3). To further improve the detection efficiency, a virtual ground at 96 GHz is formed by two $\lambda/4$ stubs. The size of M4 is designed as 4 times of M2 and M3. Note that W_{Total} and W_{Single} are the total and individual finger width of transistors in Fig. 5.26, respectively. And the channel length of every active device is 60nm. The remaining circuit consists of a common source input buffer (M1) for current injection and an envelope detector formed by M5 and M6. The common source stage (M1) is designed for input signal injection and also reverse isolation from the oscillator to the input. The size of M1 is optimized with consideration of minimized parasitic capacitance as well as the input matching. Similarly, M5 and M6 also need to be minimized but doing so will reduce the detection efficiency. To solve this problem, a capacitance coupling by C1 and C2 is introduced between the outputs of oscillator tank and the envelope detection. Firstly, the capacitance loading from M5 and M6 are reduced by series connection of the coupling capacitors; secondly, M5 and M6 are biased externally by large resistors (R1, R2) to optimize the detection; thirdly, 1/f noise from M5 and M6 is also isolated.

5.3.2.3 SRX Design by TL-SRR

Differential T-line Loaded with SRR The TL-SRR structure with horizontal placement of SRRs (Fig. 2.5) is also not suitable for the practical implementation for SRX, mainly due to the large area overhead. Compared to TL-CSRR, TL-SRR

Figure 5.26: Schematic of CMOS 96GHz SRX with DTL-CSRR.

inherently has better layout flexibility because SRRs can be vertically stacked within a compact area. One differential T-line loaded with stacked SRRs (DTL-SRR) is proposed in this work for the application of 135 GHz SRX design in 65nm CMOS RF process.

As shown in Fig. 5.27(a), the DTL-SRR is designed by stacked SRRs with the same dimensions of $24 \times 24 \mu\text{m}^2$ in 4 metal layers (M5 to M8). All SRRs are closely coupled to the same host T-line implemented in the top most metal layer (M8). The overall size of the proposed DTL-SRR is $35 \times 34 \mu\text{m}^2$. For the purpose of comparison, a traditional LC-tank resonator is designed in M8 metal layer as shown in Fig. 5.27(b), which has the same resonance frequency of 135GHz. The S-parameters of both structures are also verified by EMX with the same parasitic capacitance of 16fF. As shown in Fig. 5.28, at the vicinity of 140-GHz resonance, $\epsilon > 0$ and $\mu < 0$, and a magnetic plasmonic medium is formed. As a result, a stop-band is formed at 140 GHz within a narrow bandwidth of 3.5 GHz. The Q factor of the DTL-SRR resonator is 40, which is more than 2 times of the Q of the LC-tank resonator. Moreover, the DTL-SRR resonator layout area ($1190 \mu\text{m}^2$) is less than half of the LC-tank resonator ($2500 \mu\text{m}^2$).

Such a Q factor enhancement effect can also be explained by the strong phase non-linearity in the frequency range closed to SRRs resonance. Note that Q factor

Figure 5.27: Layout for CMOS on-chip implementation of DTL-SRR for 135GHz SRX.

Figure 5.28: EM-simulation based comparison of DTL-SRR and LC-tank resonator for CMOS 135GHz SRX design.

can also be obtained by phase based method:

$$Q = \frac{\omega_0}{2} \cdot \left| \frac{d\angle Z(j\omega)}{d\omega} \right|. \quad (5.18)$$

where $\angle Z(j\omega)$ is the phase of resonator impedance. Fig. 5.29 (a) shows the impedance diagram of both DTL-SRR and LC-Tank without any capacitor loading. A resonance generated by the SRR loadings is observed at 167 GHz for DTL-SRR. Such resonance causes non-linear phase shift at 140 GHz. Fig. 5.29 (b) shows that DTL-SRR has

much stronger phase non-linearity than that of LC-Tank around 140 GHz. As shown in Fig. 5.29 (c), both structure has the same resonance frequency of 140 GHz after including the ideal capacitance ($C = 16$ fF). The phase non-linearity in DTL-SRRs increases the phase gradient of Z_{Diff} at 140 GHz, resulting into a higher Q according to (5.18).

135-GHz DTL-SRR-based SRX Fig. 5.30 depicts the schematic of 135-GHz DTL-SRR-based SRX. Firstly, transformer based matching network is applied to the input matching for M1 for the electrostatic discharge (ESD) protection when integrating with the antenna; secondly, the virtual ground formed by two $\lambda/4$ T-line is replaced by the high-Q MOM capacitor to further reduce the chip area; thirdly, the detected envelope signal V_{ENV} is directly averaged by an on-chip low-pass filter formed by R3 and C3 at the output.

5.3.3 Measurements

5.3.3.1 DTL-CSRR-based SRX at 96 GHz

As shown in Fig. 5.31(a), the proposed DTL-CSRR-based SRX is implemented in UMC 65-nm CMOS process with f_T/f_{MAX} of 170/190 GHz. The total die area is $500 \times 440 \mu\text{m}^2$ including a core area of 0.014 mm^2 with resonator and active devices. The SRX is measured on probe station (CASCADE Microtech Elite-300) with RF input signal provided by Agilent PNA-X (N5247A) with T/R modules (N5260), of which the output power is calibrated in the range of $-85 \sim -10$ dBm by Agilent Spectrum Analyzer E4407B with a W-band waveguide harmonic mixer (11970W). Note that the output power of T/R module can be controlled by N5247A, but its minimum output power level that can be calibrated is limited by the sensitivity of spectrum analyzer. A 12.5MHz sinusoid quench-control signal is applied by function generator (AFG3022) with voltage swept in $0 \sim 250$ mV. The receiver operates under 1-V power supply with a power consumption of 2.8 mW.

The receiver gain is defined as $Gain(dB) = 20 \log V_{out} - P_{in}$, where V_{out} is the normalized output voltage of receiver and P_{in} is the input power level of proposed SRX in dBW. The actual gain of SRX is not totally independent of P_{in} as what has been discussed in (5.9) in Sec. 5.3.1. A relative higher P_{in} of -20 dBm is applied to have sufficient V_{out} at the frequency outside the bandwidth. Fig. 5.32(a) shows the measured and simulated gain and input S11, where the maximum gain of proposed SRX at -20-dBm P_{in} is 21 dB at 95.5 GHz and a 3-dB bandwidth of 560 MHz is

(a)

(b)

(c)

Figure 5.29: Impedance Diagram of DTL-SRR and LC-Tank in Globalfoundries 65-nm CMOS process, (a) real and imaginary parts of Z_{Diff} , (b) phase of Z_{Diff} , (c) phase of Z_{Diff} when the ideal 16-fF capacitor is included.

Figure 5.30: Schematic of CMOS 135GHz SRX with DTL-SRR.

Figure 5.31: Die micrographs: (a) CMOS 96GHz SRX with DTL-CSRR, and (b) CMOS 135GHz SRX with DTL-SRR.

observed. The measured gain is quite close to the post layout simulation result. A good input match is also observed with S11 smaller than -14 dB from 94.6 GHz to 96.6 GHz.

Fig. 5.32(b) shows the normalized V_{out} against P_{in} as well as the responsivity, where an almost linear response is observed with a sensitivity of -78 dBm and a maximum responsivity of 6.02 MV/W. Note that responsivity is calculated by V_{out}/P_{in} . The measured responsivity is very close to the simulation results especially when the input power level is below -70dBm. The noise equivalent power (NEP) of proposed SRX is $0.67 \text{ fW}/\sqrt{\text{Hz}}$, which is calculated by $S/\sqrt{B}$, where S is the sensitivity and B is the 3-dB bandwidth [110]. Finally, the noise figure (NF) is found to be 8.5 dB by $S/(K \cdot T \cdot B)$ at room temperature of 290K.

5.3.3.2 DTL-SRR-based SRX at 135GHz

As shown in Fig. 5.31(b), the proposed DTL-SRR-based SRX is implemented in GlobalFoundries 65-nm CMOS RF process with f_T/f_{MAX} of 180/530 GHz. It has a total die area is $570 \times 460 \text{ } \mu\text{m}^2$, and a core area of 0.0085 mm^2 . For the purpose of imaging system integration, the receiver chip is firstly attached to a test board (FR4, 1.6-mm thickness). As shown in Fig. 5.33, the whole test board is placed on probe station for SRX measurement. As such, the SRX can be easily integrated with imaging system by replacing GSG probe with bonding wires connected to an 135GHz antenna. The RF input signal is provided by a VDI D-Band signal generator, of which the output power is calibrated in the range of -85 ~ -10 dBm by the R&S FSUP signal source analyzer with a D-band waveguide harmonic mixer. A 12.5-MHz sinusoid quench-control signal is applied from function generator (Agilent 33250a) with voltage sweep-range of 0~400mV. Operating from a 1-V power supply, the receiver consumes 6.2 mW.

Fig. 5.34(a) shows the measured gain when P_{in} is -18 dBm, where the maximum gain of proposed SRX is 15.5dB at 134.8GHz and the 3-dB bandwidth is 530 MHz. The measured gain is also quite close to the post layout simulation result, but the center frequency of measurement results is 5-GHz lower than that from simulation, which is probably due to the inaccurate transistor model above 100 GHz. Fig. 5.34(b) shows normalized V_{out} against P_{in} as well as the responsivity, where the receiver sensitivity (S) and the maximum responsivity are observed as -76.8 dBm and 4.82 MV/W, respectively. Note that the measured responsivity is very close to the simulation results especially when the input power level is below -75dBm. And the NEP is calculated to be $0.9 \text{ fW}/\sqrt{\text{Hz}}$. A near linear relationship between V_{out} and P_{in} is observed when the input power is below -40 dBm, which can be utilized in post-data processing to generate THz images.

(a)

(b)

Figure 5.32: Measurement and simulation results of CMOS 96GHz SRX, (a) gain and input S11, and (b) output voltage (V_{out}) and responsivity vs. input power (P_{in}).

Figure 5.33: Measurement setup of CMOS 135GHz SRX with DTL-SRR.

5.3.3.3 Comparison and Discussion

The performance of the measurement results of proposed SRXs is summarized in TABLE 5.3 as well as the previous state-of-the-art receiver designs. Compared to the direct conversion receiver [129], SRXs has a 16 ~ 22 dB better sensitivity due to a narrower receiver bandwidth. Compared to the traditional SRX designs with LC-tank resonator [110, 113], the proposed SRXs are showing 30% ~ 50% reduced NEP, 2.8 ~ 4dB better sensitivity and 60% area reduction, which makes them well suitable for the portable THz imaging with large sensor array.

5.4 Conclusion

High-Q oscillations can be achieved by metamaterial resonators such as DTL-SRR and DTL-CSRR. As demonstrated in this section, Both DTL-SRR and DTL-CSRR can be applied in the oscillators design with low phase noise, low power, and compact chip. Both oscillators have 4~6dB better phase noise than that of the standing-wave oscillator by a CPS with similar operating frequencies. Both DTL-SRR and DTL-CSRR can also be applied in the super-regenerative receivers design with quench-controlled oscillators to achieve better sensitivities. When compared to the conventional SRX design with LC-tank resonator at the similar frequency, both proposed

(a)

(b)

Figure 5.34: Measurement and simulation results of CMOS 135GHz SRX, (a) gain, and (b) output voltage (V_{out}) and responsivity vs. input power (P_{in}).

TABLE 5.3: PERFORMANCE COMPARISON OF STATE-OF-THE-ART RECEIVERS FOR IMAGING APPLICATION

Parameters	[110]	[113]	[129]	DTL-CSRR	DTL-SRR
Topology	SRX	SRX	DC	SRX	SRX
Resonator Type	LC-tank	LC-tank	-	TL-CSRR	TL-SRR
Technology Node	65-nm CMOS	65-nm CMOS	180-nm BiC-MOS	65-nm CMOS	65-nm CMOS
f_{osc} (GHz)	144	183	103	95.5	135
Power (mW)	2.5	13.5	225	2.8	6.2
Sensitivity (dBm)	-74	-72.5	-56	-78	-76.8
Bandwidth (GHz)	0.94	1.4	20	0.56	0.53
NF (dB)	10.2	9.9	15	8.5	9.7
NEP (fW/ $\sqrt{Hz}$)	1.3	1.5	17.8	0.67	0.9
Core Area (mm ²)	0.021	0.013	0.75	0.014	0.0085

SRXs at 96 GHz and 135 GHz shows 2.8 ~ 4 dB improved sensitivity. Especially, the proposed SRXs at 135 GHz has 60% reduced core chip area In the following section, CMOS based THz antenna design will be discussed.

Chapter 6

THz On-chip SIW and LWA Antennas

6.1 Introduction

In order to compensate huge propagation loss of THz signal, in addition to the high power signal sources and high sensitivity receivers, an effective THz imaging system also requires antenna and antenna array with high-directivity radiation pattern [130] and high efficiency with sufficient bandwidth, which imposes grand challenges for an on-chip antenna design in CMOS. Firstly, the realization of highly directive radiation is not trivial. The conventional right-handed antennas (patch, dipole and etc.) [24,131] have positive phase-and-length relation that usually results in non-in-phase radiation and large sized design. Secondly, due to close distance between top metal layer and ground, the radiation efficiency of on-chip antenna is not high. Moreover, each antenna elements must be as compact as possible to form antenna array in limited chip area. The previous on-chip antenna works [24, 113, 131, 132] have either low gain or narrow bandwidth as well as ignored polarization issue. For example, given the THz source with a linearly polarized radiation, the longitudinal polarization may be turned into transverse direction after penetrating through the tissue [8]. If the antenna at receiver is also linearly polarized, the detection efficiency may be largely reduced due to the mismatch in the polarization directions.

On-chip antennas have more stringent requirements on the antenna size than the off-chip ones. They are usually implemented at a frequency higher than 200GHz, of which a structure with equivalent electrical length of $\lambda/4$ or $\lambda/2$ can be fit into the chip scale with good efficiency. There are two types of commonly used right-handed

on-chip antenna designs: patch antennas [24, 113, 132] and dipole antennas [129, 131]. Typically a patch antenna has broadside radiation pattern and needs single-ended feeding. It has relative higher gain (~ 1 dBi) and radiation efficiency ($\sim 25\%$) due to the shielding of metal grounding layer between the patch structure and lossy silicon substrate. However, the fractional bandwidth of a patch antenna is small ($\sim 2.5\%$), which is limited by the distance between on-chip patch and ground metal layers. Moreover, the actual size of patch antenna is usually large if considering a ground plane with at least twice the size of the patch structure. Dipole antenna has an omnidirectional radiation that can be excited differentially. Although there are many advantages in dipole antenna such as compact size and wide bandwidth ($>15\%$), it suffers from a low gain (~ -8 dBi) and a poor radiation efficiency ($<10\%$) as a large fraction of the radiated power is lost in the silicon substrate,

Substrate integrated waveguide (SIW) structures have been recently explored for the design of high quality factor (Q) passive devices in both mm-wave and THz regions [133–135]. SIW can be regarded as a dielectric-filled rectangular waveguide with surrounding walls and metal layers on the top and bottom surfaces, which leverages the advantages of both planar transmission line and non-planar waveguide with lower loss and wide band performance in a miniaturized cavity for on-chip antenna design. SIW antennas designs are proposed in both PCB scale [133] and chip scale [135] with a wide bandwidth and a high gain. In [135], a 400-GHz linear polarized on-chip SIW antenna is demonstrated in SiGe process with -0.55 -dBi gain and 7.8% relative bandwidth. However, its dimension has to satisfy an equivalent electrical length of $\lambda/2$, which should be further miniaturized when designed on-chip. In the first part of this work, one wide band 280-GHz on-chip circularly polarized SIW antenna is design in CMOS process with compact area. By creating corner slots in SIW structure, the size of the proposed SIW antenna is reduced by 15% when compared to the conventional designs. As verified by the EM simulation, the proposed antenna has -0.5 -dBi antenna gain and 32.1 -GHz bandwidth centered at 268 GHz.

On the other hand, the recently explored left-handed metamaterial [49] can provide negative phase with a nonlinear phase-and-length dependence, which can realize a zero-phase EM-wave propagation or radiation with compact area on chip [136], ideally for a broadside radiation antenna design. In the second part of this work, the on-chip THz leaky wave antenna (LWA) design is explored with periodic composite right/left handed (CRLH) transmission-line (T-line) structure with broadside radiation under a zero phase propagation condition. Moreover, stacking high resistivity dielectric layer is deployed to improve efficiency. A 13-cell on-chip CRLH T-line is fabricated in 65nm

Figure 6.1: Geometrical configuration of the proposed SIW antenna with four corner slots

CMOS process and is characterized from 220 GHz to 325 GHz. The corresponding LWA design shows broadside radiation pattern, 65% efficiency and 4.1-dBi gain at 280 GHz.

6.2 On-chip Antenna Design with SIW

6.2.1 Circularly Polarized SIW Antenna with Corner Slots

A compact circular-polarized SIW antenna design is designed in CMOS process with corner slots, of which the geometrical configuration is illustrated in Fig. 6.1. The operating frequency of a SIW antenna is determined by the cavity dimension. It can be approximated by the following equation by considering the cavity resonance model [137].

$$f_{mnp} = \frac{c}{2\sqrt{\mu_r \varepsilon_r}} \sqrt{\left(\frac{m}{L_{eff}}\right)^2 + \left(\frac{n}{W_{eff}}\right)^2 + \left(\frac{k}{h}\right)^2} \quad (6.1)$$

where $L_{eff} = L$ and $W_{eff} = W$ are the effective length and width of the substrate integrated cavity, c is the speed of light in free space, μ_r and ε_r are the relative

permeability and permittivity of the dielectric material inside the cavity and h is the cavity height. The resonance modes with lowest order are used for a minimum antenna size, considering h is much smaller than the wavelength when designed on-chip, the available resonance frequencies left are f_{210} and f_{120} with

$$\begin{cases} f_{210} = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \sqrt{\frac{4}{L_{eff}^2} + \frac{1}{W_{eff}^2}} \\ f_{120} = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \sqrt{\frac{1}{L_{eff}^2} + \frac{4}{W_{eff}^2}} \end{cases} \quad (6.2)$$

After introducing corner slots with 45° to the edge, both L_{eff} and W_{eff} can be approximated by the total effective length of the center rectangular cavity and two keystone cavities as shown in Fig. 6.1:

$$\begin{cases} L_{eff} = L_{rectangular} + 2L_{keystone} \\ W_{eff} = W_{rectangular} + 2W_{keystone} \end{cases} \quad (6.3)$$

where $L_{rectangular} = L - \sqrt{2}L_s$, $W_{rectangular} = W - \sqrt{2}L_s$, $W_{keystone}$ and $L_{keystone}$ are the effective lengths of each keystone cavity in the X and Y directions, respectively. The effective length of keystone cavity can be approximated by the following equations [138]:

$$\begin{cases} L_{keystone} = \frac{L_s R_{tl}}{1.152} \\ W_{keystone} = \frac{L_s R_{tw}}{1.152} \end{cases} \quad (6.4)$$

with

$$\begin{cases} R_{tl} = \frac{L_s(2W - \sqrt{2}L_s)}{\sqrt{2}WL_s} \\ R_{tw} = \frac{L_s(2L - \sqrt{2}L_s)}{\sqrt{2}LL_s} \end{cases} \quad (6.5)$$

With (6.3), (6.4) and (6.5), L_{eff} and W_{eff} can be simplified as:

$$\begin{cases} L_{eff} = L + 1.042L_s - \frac{1.737L_s^2}{W} \\ W_{eff} = W + 1.042L_s - \frac{1.737L_s^2}{L} \end{cases} \quad (6.6)$$

As observed from (6.6), both the effective width and length are extended by a factor of L_{eff}/L or W_{eff}/W after introducing corner slots, which means the antenna size can be reduced by the same ratio at a particular frequency. However, the reduction ratio is also limited by higher order effects. As shown in (6.6), L_{eff} and W_{eff} reach their maximums of $1.15L$ and $1.15W$ when $L_s = 0.3W$ and $0.3L$, respectively, which is equivalent to a 15% size reduction in each dimension of SIW antenna.

Note that the lengths of center slots $L1$ and $W1$ can be calculated by:

$$\begin{cases} L1 = \frac{1}{2f_{L1}\sqrt{\mu_{eff}\varepsilon_{eff}}} \\ W1 = \frac{1}{2f_{W1}\sqrt{\mu_{eff}\varepsilon_{eff}}} \end{cases} \quad (6.7)$$

where f_{L1} and f_{W1} are the resonant frequencies of center slots, μ_{eff} and ε_{eff} are the equivalent permeability and permittivity in the center slots, respectively. In a conventional SIW antenna design [133], to maximize radiation efficiency, both center slots need to have the same resonant frequencies as the respective resonance mode: $f_{L1} = f_{120}$ and $f_{W1} = f_{210}$. By properly adjusting the cavity dimensions, f_{120} and f_{210} can be close to each other that a circularly polarized radiation is generated at a frequency in between. However, the antenna designed in such method has a narrow bandwidth, because only two resonance modes exist. In this work, the radiation bandwidth is extended by introducing additional resonance modes in the antenna design, where four resonance modes are generated by designing f_{L1} and f_{W1} slightly lower and higher than f_{120} and f_{210} , respectively. Moreover, the antenna performance is further improved from the following two aspects. Firstly, the radiation efficiency is increased with the cavity height, which could be achieved by selecting a CMOS process option with a large number of stacking layers. Secondly, the metal loss of SIW walls is largely reduced by replacing metal vias with metal bars. Note that vertical connection by metal bars is an option provided in the standard CMOS process to connect many vias horizontally (Fig. 6.2) if the metal density is not critical in the particular area.

6.2.2 On-chip Implementation in 65-nm CMOS

The proposed on-chip SIW antenna is designed in 65nm CMOS process with 9 metal layers as shown in Fig. 6.2. A composite dielectric material with silicon dioxide (SiO_2) and silicon nitride (Si_3N_4) is enclosed in the cuboid cavity ($410 \mu\text{m} \times 410 \mu\text{m} \times 9 \mu\text{m}$) formed by the top most aluminum layer (AL), bottom most copper layer (M1) and metal walls constructed by metal layers and via bars (M1-AL). The chip area required for SIW antenna is 0.17 mm^2 . Two $17\text{-}\mu\text{m}$ wide rectangular slots with different lengths ($325 \mu\text{m}$ and $360 \mu\text{m}$) are crossed at the center of AL layer to create four resonance modes with perpendicular polarization directions at 270 GHz. A rectangular slot ($120 \mu\text{m} \times 30 \mu\text{m}$) are created at each corner to reduce the antenna size at the desired operating frequency as discussed in Sec. 6.2.1. Different from the SIW design in Printed Circuit Board (PCB), a uniform metallic plane is not available in CMOS process according to the metal density rules. As such, bottom side (M1) is

Figure 6.2: Design of on-chip integrated circular-polarized SIW antenna in CMOS 65nm process.

implemented in mesh type that tiny square slots ($3.5 \times 3.5 \mu\text{m}^2$) are placed with $12\text{-}\mu\text{m}$ pitch. The antenna input is fed by a micro-strip line with characteristic impedance of 40Ω , which is implemented by M8 and M1 layers for signal and ground, respectively.

6.2.3 EM Simulation

The proposed antenna structure is verified by a full wave simulation in Ansoft HFSS. Fig. 6.3 shows the simulated radiation pattern at 270 GHz. It has a broadside radiation pattern with 6.2-dB directivity and a 21.4% radiation efficiency at 270 GHz. As a result, a -0.5-dB antenna gain is obtained as illustrated in the simulated 3D radiation pattern in Fig. 6.4. Fig. 6.5 shows the simulated E-field distribution of SIW cavity at 270 GHz, where a circularly polarized field distribution is observed with two modes (f_{120} and f_{210}) excited with a phase difference of 90° . The simulated 3dB-axial-ratio frequency range is 265.6-274.5GHz as observed from Fig. 6.6. Fig. 6.6 also shows the S11 of the proposed antenna. As a result of the difference between W1 and L1, f_{120} is slightly different from f_{210} . Thereby four resonances are generated including f_{120} , f_{210} , f_{L1} and f_{W1} , resulting a very wide -6dB S11 bandwidth of 32.1GHz centered at 268GHz.

Figure 6.3: HFSS Simulation results of proposed SIW antenna directivity at 270 GHz.

6.3 On-chip Antenna Design with CRLH T-line

6.3.1 CRLH T-line Based Leaky Wave Antenna Design

The recently explored left-handed metamaterial can provide negative phase with a nonlinear phase-and-length dependence, which can realize a zero-phase EM-wave propagation or radiation with compact area on chip [139], ideally for a broadside radiation antenna design. In this section, the CRLH T-line is studied for leaky wave antenna (LWA) design at THz with broadside radiation under a zero phase propagation condition.

CRLH T-line is a well-know metamaterial structure that can achieve positive, negative and zero phase propagation. A CRLH T-line consists of number of periodic unit-cells. The traveling wave can be generated in the CRLH T-line based LWA with

$$\phi(y, z) = \phi_0 e^{-jk_y y} e^{-\gamma z} \quad (6.8)$$

where $\gamma = \alpha - j\beta$ is the propagation constant of the travelling wave in CRLH T-line; $k_y = \sqrt{k_0^2 - |\beta|^2}$ is wave factor of radiated power along y axial; and k_0 is wave number in free space.

Figure 6.4: HFSS simulation results of proposed SIW antenna radiation pattern at 270 GHz.

When signal travels along the antenna surface (z axis) as shown in Fig. 6.7, the energy leak-out is confined in parallel with k_0 . If the phase velocity is slower than that of light, or $|\beta| > |k_0|$, the signal is attenuated exponentially along y axial. In this case the antenna works in slow-wave region. On the other hand if $|\beta| < |k_0|$, the antenna is in the fast-wave mode with a real k_y , which is a desired condition for LWA to operate. Note that the main beam radiation angle can be described by

$$\theta_{MB} = \arcsin(|\beta|/|k_0|) \quad (6.9)$$

Therefore, the radiation pattern becomes broadside when $|\beta|$ equals to zero, and beam steering can be observed along broadside radiation with high directivity for THz communication.

6.3.2 280-GHz LWA Design by CRLH T-line

Based on the fabricated 13-cell CRLH T-line in Sec. 3.4.2, one 280-GHz LWA is also implemented in Global Foundry 65-nm CMOS process. P1 in Fig. 3.10 is selected

Figure 6.5: HFSS simulation results of polarized vector electric field in the slots of SIW cavity at 270 GHz.

as the antenna input, and P2 is left open circuit. As shown Fig. 6.8, a standard high resistivity silicon layer ($750 \Omega \cdot \text{cm}$) with a thickness of $100 \mu\text{m}$ is placed on top of the antenna surface to enhance the radiation efficiency.

6.3.3 EM Simulation

The proposed CRLH T-line based LWA design is verified by EM simulation in HFSS. As shown in Fig. 6.9, the maximum efficiency was 6.4% in $230 \sim 290$ GHz, and it is enhanced to $40.5 \sim 65.2\%$ after stacking the dielectric layer of with high resistivity of Si. The maximum enhancement of 26 times is achieved. After the enhancement of efficiency, the maximum antenna gain of 4.1 dBi is achieved at 280GHz. As illustrated in the radiation pattern shown in Fig. 6.10, a broadside radiation is observed at 280 GHz when $\beta = 0$. Note that the zero phase propagation at 280 GHz also provides higher gain and efficiency than the negative phase ($\beta < 0$ at 290 GHz) and positive phase ($\beta > 0$ at 250 GHz) propagation with tilted radiation direction. Note that the zero- β frequency of 280GHz shown in Fig. 6.10 is lower than the 303 GHz shown in Fig. 3.13. This is mainly because the increase of equivalent permittivity due to the

Figure 6.6: HFSS simulation results of input S_{11} and antenna axial ratio on broadside radiation direction.

Figure 6.7: Operation diagram of leaky wave antenna: (a) $\beta > 0$, (b) $\beta = 0$, and (c) $\beta < 0$.

stacking of a high resistance silicon layer.

Figure 6.8: Stacking of high resistivity Si layer on-top of the LWA.

Figure 6.9: Radiation efficiency enhancement by stacking high resistivity Si layer.

6.4 Conclusion

The designs of THz CMOS wide band and high gain on-chip antennas are demonstrated in compact area by high-Q passive structures such as substrate integrated waveguide (SIW) and metamaterial based composite right/left handed transmission

Figure 6.10: Gain radiation patterns for the proposed antenna at three frequencies: $f=250\text{GHz}$ ($\beta < 0$, backward radiation), $f=280\text{GHz}$ ($\beta = 0$, broadside radiation), and $f = 290\text{GHz}$ ($\beta > 0$, forward radiation).

line (CRLH T-line). In the wide band 280-GHz on-chip circularly polarized SIW antenna design in 65nm CMOS process, by creating corner slots in SIW structure, the size of the proposed SIW antenna is reduced by 15% when compared to the conventional designs. As verified by the EM simulation, the proposed antenna has -0.5-dBi antenna gain and 32.1-GHz bandwidth centered at 268 GHz. In the CRLH T-line based leaky wave antenna (LWA) design in 65-nm CMOS process, stacking of dielectric layer with high resistivity of Si is utilized to improve the LWA efficiency. With correlated measurement and EM validation from 220 GHz to 325 GHz, a broadside radiation pattern is achieved at 280 GHz with 65% radiation efficiency and 4.1dBi antenna gain at 280GHz. Both antenna structures can be potentially deployed for the design of on-chip antenna array in the THz biomedical imaging systems. In the following section, CMOS based Transceiver design for THz Imaging will be discussed.

Chapter 7

CMOS THz Transceiver for Imaging

7.1 Introduction

High performance THz imaging systems can be constructed by the proposed on-chip metamaterial based signal sources, receivers and antennas in the previous sections. As illustrated in Fig. 7.1, A high power THz signal firstly generated by MPW based zero-phase CON and then radiated by the CRLH T-line based on-chip LWA. After penetrating through the sample under test, the resulting THz signal is received by a high sensitivity super-regenerative receivers by TL-SRR/TL-CSRR based quench-controlled oscillators. With the proposed transmitter and receiver designs, both a narrow band and a wide band THz imaging systems can be demonstrated at 135 GHz and 280 GHz, respectively.

In this chapter, firstly, a narrow band transmission type THz imager is demonstrated at 135 GHz with various pharmacy and security applications. To further enhance a wide band CMOS THz imaging system at 280 GHz with high sensitivity, and high spectrum resolution, a heterodyne receiver architecture is required [140]. As shown in Fig. 7.2, a high sensitivity CMOS wide band transmission-type THz imager is also demonstrate by integrating the circular polarized substrate integrated waveguide (SIW) antenna introduced in Sec. 6.2 with a heterodyne receiver, which consists of a down-conversion mixer and a power gain amplifier (PGA). The down-conversion mixer with single-gate topology can achieve 80-GHz bandwidth with a conversion gain of -19 dB. The three-stage PGA achieves 150-MHz bandwidth for the detection resolution. The entire imager is measured with -2-dBi conversion gain

Figure 7.1: Metamaterial based THz imaging system

over 42-GHz bandwidth, -54.4-dBm sensitivity at 100-MHz detection resolution bandwidth, 6.6-mW power consumption and 0.99-mm² chip area with high contrast images measured.

Figure 7.2: THz image system with heterodyne receiver for high spectrum resolution detection and circular-polarized antenna for the tolerance of depolarization effect.

In addition, reflection based THz imager is also required for in-vivo skin cancer diagnosis. Compared to the transmissive imaging system, reflective type has higher requirement of transmitter power, receiver sensitivity and the control of path of incident

and reflected signal. As such, a reflective CMOS THz imaging system is also proposed based on simulation results in this work with on-chip integrated THz transceivers, which has been constructed by a differential heterodyne receiver as well the on-chip CRLH-TL leaky wave antenna (LWA) and zero-phase CON based signal source demonstrated in previous sections.

7.2 135-GHz Narrow Band Imager by DTL-SRR-based SRX

7.2.1 THz Imaging by SRA Detection

THz radiation is usually attenuated due to absorption and scattering during the propagation [141], which can be modeled by

$$I_{\gamma_1} = I_{\gamma_0} e^{-\int_{\gamma_0}^{\gamma_1} \alpha_e(z) dz} \quad (7.1)$$

where I_{γ_0} and I_{γ_1} are the incoming and outgoing radiance intensity along path (γ_0, γ_1) ; and $\alpha_e(z)$ is the extinction coefficient, which is the summation of absorption (α_a) and scattering (α_s) coefficients. For a homogeneous material placed between (γ_0, γ_1) , α_a is a constant. The scattering only happens at the interface with scattering coefficients of $\alpha_s(\gamma_0)$ and $\alpha_s(\gamma_1)$. The received power (P_R) in a transmissive-type THz imaging system is [142]

$$P_R(dBm) = P_T(dBm) + G_R(dBi) - L(dB) - 8.686[\Delta\gamma\alpha_a + \alpha_s(\gamma_0) + \alpha_s(\gamma_1)](dB) \quad (7.2)$$

where P_T is the effective isotropic radiated power (EIRP) of transmitter, G_R is the receiver antenna gain, and L is the path loss without any objects placed in the propagation path, including both the free space path loss (FSPL) and atmosphere absorption.

As shown in (5.8), the envelope of receiver output is proportional to the injected current or the square root of input power. A DC output can be obtained by averaging $Env[v_o(t)]$ in each periodic quenching cycle.

$$V_{DC} = \frac{\omega_i Z_0 |S(\Delta\omega)| \sqrt{P_R}}{\sqrt{2R}(t_b - t_a)} \int_0^{t_b} \frac{1}{s(t)} dt \quad (7.3)$$

As such, the received power could be detected by measuring V_{DC} from the SRA output.

As a result, the THz image of an object can be further obtained by the 2D scanning of V_{DC} with fixed P_T , G_R and L . Moreover, by analyzing V_{DC} as well as P_R with various object thickness ($\Delta\gamma$), one can further find the absorption coefficient of the object under test.

7.2.2 Narrow band imaging at 135 GHz

The SRX can be integrated with the THz imaging system by replacing GSG probe with bonding wires connected to an 135-GHz antenna. It is demonstrated by wires bonding from the input of the proposed 135-GHz SRX to a 2×4 antenna array with hybrid series/parallel feeding network as show in Fig. 7.3(a). The receiver and antenna must be well aligned to minimize the connection loss, which is estimated to be 3~5dB according to the EM simulation in Ansoft HFSS. A 2x4 antenna array using hybrid series/parallel feed is designed and fabricated in Roger RT5880 with size of 8×8 mm². The antenna has 15.4dBi simulated gain at 135 GHz. Its input is matched to 50ohm with measured S11 below -10dB from 124 to 139GHz. Detailed information of the antenna array design is shown in [143]. The entire THz imaging setup is also shown in Fig. 7.3(b). The 135-GHz radiation from a VDI source (0-dBm output power) is received by proposed SRX after propagating through the objects under test, which is hold by an X-Y moving stage (STANDA) placed in the middle. Although a substantial portion of the object is illuminated due to the divergent beam from the source antenna, only the power propagating to the direction of receiver is detected. As such, a high resolution image can be obtained without focus lens. The resulting V_{out} at each X-Y stage position is recorded into a 2D matrix, which can be plotted in colored image with by Matlab with JET colormap.

Fig. 7.18 shows the imaging results by the proposed CMOS THz image system. Fig. 7.4 demonstrates the detection of knife, perfume, and coin inside a hand-bag. These items can be clearly identified in the image, because different material types like metal, plastic and liquid have different absorbtion and reflection properties to the THz radiation. Fig. 7.5 shows that one can differentiate between a moisturized Panadol pill and a dry one. Due to the strong water absorption at THz frequencies, a moisturized Panadol has higher absorbtion than the dry one. Fig. 7.6 shows the imaging of various types of eaten oil including sunflower, olive, fresh soybean and soybean that has been used once. Note that four petri dishes are used to hold the oil samples.

The imaging system can also be applied in transmission analysis to characterize

Figure 7.3: (a) PCB integration of CMOS 135-GHz SRX with antenna; (b) THz imaging measurement setup with the proposed receiver chip integrated on PCB and object under test fixed on an X-Y moving stage.

the material in the propagation path. The absorption ratio of each oil type can be identified by comparing the received power under different sample volume with the help of (7.2) in Section II, and it is depicted in the box chart in Fig. 7.7. It is interesting to

Figure 7.4: Images captured by imaging system with the proposed 135-GHz SRX receiver: knife, perfume, and coin in hand-bag.

Figure 7.5: Images captured by imaging system with the proposed 135-GHz SRX receiver: moisturized and normal Panadol pills.

observe that the soybean oil that has been used once has higher absorption to the 135-GHz energy than the fresh one. With the significantly improved receiver sensitivity, the proposed CMOS THz imager results in high contrast images and it can be further utilized in the analysis of moisture level as well as the identification of particular liquid content.

Figure 7.6: Images captured by imaging system with the proposed 135-GHz SRX receiver: various types of oil.

Figure 7.7: Absorption ratio of various types of oil detected at 135 GHz.

7.3 240 ~ 280-GHz Wide Band Imager with Heterodyne Receiver

Since THz radiation is highly sensitive to the crystal lattice vibration, hydrogen-bond as well as intermolecular interactions, it results in unique spectroscopy fingerprints for many materials. There are two design targets for receiver to enable the spectroscopy analysis in a sub-THz imaging: high spectrum selectivity and wide frequency range of operation. Diode detection based receivers [24, 129, 131] can achieve the latter target, but not the former one; while super-regenerative based receivers [113, 132] can achieve the former target, but not the latter one. In order to satisfy both of the two targets in the receiver design, one needs to deploy either heterodyne [144] or direct-conversion receivers. Compared to direct-conversion architecture with a zero IF, heterodyne architecture with a near-zero IF is more flexible in the system design with both magnitude and phase detection capability. The magnitude detection in an sub-THz imaging system can be achieved by either heterodyne or direct-conversion architecture, but only heterodyne architecture is able to retain the phase information of sub-THz signal, which is very useful to analyze the complex refractive index of the sample under test. As such, this paper focuses on the design of heterodyne receivers with a near-zero IF.

7.3.1 Architecture and System Specification

The design of CMOS heterodyne receiver in THz has to be conducted in a scenario without any low noise amplifiers (LNA) as illustrated in Fig. 7.2, because hardly can any amplifiers be designed at a frequency close to or above the f_{max} . For example, the f_{max} in a typical CMOS 65nm process is around 300 GHz [145],

The receiver gain (G_{tot}) in such case can be calculated as

$$G_{tot} = G_{ant}G_{mix}G_{pga} \quad (7.4)$$

where G_{ant} , G_{mix} and G_{pga} denote the gain of antenna, mixer and power gain amplifier (PGA), respectively. Also, the total receiver noise figure (NF_{tot}) becomes

$$NF_{tot} = NF_{mix} + \frac{NF_{pga} - 1}{G_{mix}} \quad (7.5)$$

where NF_{mix} and NF_{pga} denote the NF of mixer and VGA, respectively. Equ. (7.4) denotes that the total receiver gain can be improved from antenna, mixer and PGA,

while (7.5) denotes that the noise contributed by each stage decreases as the total gain of preceding stages. The noise contributions from mixer and PGA are no longer negligible without LNA, so the NF of receiver has to be improved by increasing the conversion gain of mixer and minimizing the noise figure of PGA. In following sections, the designs of mixer and PGA in a THz CMOS heterodyne receiver are introduced.

7.3.2 THz Down-conversion Mixer Design

As the first active building block connected to the antenna, the design of the down-conversion mixer largely affects the performance of heterodyne receiver, including conversion gain and NF. Conventionally there are two types of mixers that commonly used in the mm-wave region: Gilbert-cell mixer and single-gate mixer [146,147]. Gilbert-cell mixer [146] has a compact size and low implementation loss, and it generates the cross modulation product of LO and RF signals. However, the conversion gain of gilbert-cell mixer largely depends on the transconductance of transistors in the saturation region, which will be heavily reduced when the signal frequency is approaching f_{max} . On the other hand, single-gate mixer [147] utilized the nonlinearity of transistors when biased in the subthreshold region, which is less frequency dependent compared to Gilbert-cell mixer, and is able to work at a higher frequency in THz. Moreover, compared to the subharmonic mixer working with $1/3$ -LO [135], the conversion loss could be largely reduced when directly mixing the RF and LO signal in fundamental tones. In this work, a single-gate mixer is designed to down-convert the RF signal in 220 ~ 300 GHz to the baseband by the fundamental tone of LO signal.

Figure 7.8: Schematic of THz down-conversion mixer at 280GHz.

Fig. 7.8 shows the schematic of proposed down-conversion mixer design.. A wilkinson combiner implemented by coplanar waveguide (CPW) is deployed to combine the

RF signal from antenna and the LO signal, and also to provide isolation in between. The combined output is connected to the input of common-source stage (M1) biased in the subthreshold region ($0.4\text{-V } V_{GS}$) by a compact composite CPW and lump components matching network. The following common-gate stage is applied to improve the conversion efficiency as well as the reverse isolation. One LC resonator is connected between VDD and the mixer output to filter out the unwanted harmonics of IF signal.

A post-layout simulation is performed to the proposed mixer design with passive devices simulated in EMX environment, including power combiner, matching network and inductors. A maximum conversion gain of -19 dB is demonstrated by proposed mixer in Fig. 7.9 (a) when the power of LO is 0 dBm. Note that the on-chip generation of 0 dBm LO power by 65nm CMOS has been recently demonstrated in [148]. Compared to the subharmonic mixer design in [135], the conversion gain is improved by more than 10 dB. Also, the proposed mixer has a wide operation frequency range with a gain of $-19 \sim -22\text{dB}$ from 220 GHz to 300 GHz. Moreover, a good input matching and LO-RF isolation is achieved with S11, S22 and S12 smaller than -10 dB in $220 \sim 300$ GHz. Note that the conversion gain of proposed mixer is also determined by the available LO power at mixer input. As shown in Fig. 7.9 (a), the conversion gain will drop to -37 dB when LO power is reduced to -20 dBm,

7.3.3 Power Gain Amplifier at 3 GHz

There are two major objectives in the PGA design in a THz heterodyne receiver. First, sufficient gain must be provide to the targeted IF frequency with a low noise figure. Compared to the common-source amplification topology, cascode is more preferred with higher gain and stability. Second, a narrow frequency response is required to increase the selectivity of receiver for the purpose of THz imaging. Generally, a narrow frequency response can be achieved by a resonator tank with a high quality factor, which is mainly determined by the inductor for on-chip implementation. However, since the inductor size is inversely proportional to the resonating frequency for a given Q factor, it will generate large chip area overhead when the resonant frequency is too small. As such, an optimized resonant frequency of 3GHz is selected in the PGA design.

Fig. 7.10 shows the schematic of the proposed PGA, which is implemented by three stages of cascode amplifiers followed by a common-source output buffer. In each cascode stage, both transistors are biased in the saturation region. ($0.6\text{-V } V_G$ for M3, M5 and M7, V_G is connected to VDD for M4, M6 and M8) The resonator is

(a)

(b)

Figure 7.9: Simulation results of proposed mixer: (a) S-parameters and conversion gain when sweeping RF and LO frequencies with $F_{LO} = F_{RF} + 3GHz$; (b) conversion gain at different LO power level when sweeping RF frequency with $F_{IF} = F_{RF} - 280GHz$.

implemented by a 410-fF metal-insulator-metal capacitor and a 3.5nH spiral inductor. A common-source output buffer is used to drive a 50-Ohm output impedance for the purpose of measurement. The post-layout simulation is also performed to the proposed PGA design. As shown in Fig. 7.11, a maximum gain of 33dB is obtained at center frequency of 3 GHz, and the 3-dB bandwidth is 150 MHz. Moreover, the proposed PGA has noise figure lower than 4 dB from 2.5 GHz to 3.5 GHz.

Figure 7.10: Schematic of the three-stage power gain amplifier and the output buffer.

Figure 7.11: Simulation results of the three-stage power gain amplifier with output buffer.

7.3.4 Measurements

The proposed wide band CMOS imager is fabricated in Global Foundries (GF) 65-nm CMOS RF process after integrating the heterodyne receiver with the circular polarized substrate integrated waveguide (SIW) antenna introduced in Sec. 6.2. The die micrograph is shown in Fig. 7.12 with a chip area of 0.99 mm^2 . The fabricated receiver chip is firstly measured alone followed by the applications in the THz imaging system.

Figure 7.12: Chip micro photograph of the proposed CMOS 280-GHz heterodyne receiver with on-chip integrated circular polarized SIW antenna.

Figure 7.13: Equipment setup for 280-GHz receiver measurement.

7.3.4.1 Receiver Measurements

The receiver operates under 0.8-V power supply with overall power consumption of 6.6 mW. As shown in Fig. 7.13, the receiver chip is firstly measured on probe station (CASCADE Microtech Elite-300). A LO-signal (VDI) is directly injected via a waveguide GSG probe with 50- μ m pitch from 220 GHz to 330 GHz, and a RF signal

is emitted by a 20-dB gain horn antenna placed right above the chip under-test by 10-cm distance. The output IF signal is connected to another low-frequency GSG probe with 100- μm pitch.

Figure 7.14: Gain and sensitivity measurement results when sweeping RF and LO frequencies with $F_{LO} = F_{RF} + 3GHz$.

The receiver output power is measured by a spectrum analyzer (Agilent E4408b) when the power of RF source (VDI) is pushed to the maximum power level (~ -10 dBm). The receiver gain in (7.4) can be obtained by $G_{tot}(dBi) = P_{IF} - ERIP_{RF} + L(d)$, where P_{IF} is the output power of receiver in dBm, $ERIP$ is the equivalent isotropically radiated power of signal source in dBm, $L(d)$ is signal propagation loss in dB and d is the distance between the horn antenna and the SIW antenna in the receiver. Note that $ERIP_{RF}$ and $L(d)$ can be obtained by the following equations:

$$\begin{cases} ERIP_{RF} = P_{TX} + G_{TX} \\ L(d) = 20 \log \left(\frac{c}{4\pi f d} \right) + 4.343\alpha d \end{cases} \quad (7.6)$$

where P_{TX} is the source power, G_{TX} is the horn antenna gain, and α is the attenuation factor due to the atmospheric absorption, which is almost negligible for an in-door environment. The wide band gain response is measured by fixing IF output frequency (f_{IF}) at 3GHz and sweeping RF and LO frequencies (f_{RF} and f_{LO}) simultaneously

with $f_{LO} = f_{RF} + f_{IF}$. As shown in Fig. 7.14, the proposed receiver is measured with an operating bandwidth of 42 GHz from 239 GHz to 281 GHz and a maximum conversion gain of -25 dBi. The narrow band selectivity response is measured by fixing f_{LO} at 283GHz and sweeping f_{RF} with $f_{IF} = f_{LO} - f_{RF}$. As shown in Fig. 7.15, a 100-MHz resolution bandwidth is observed. This is slightly lower than simulated bandwidth of PGA because of additional LC resonator in the down-conversion mixer. The best sensitivity (S) is found to be -31.4 dBm at 250 GHz as illustrated in Fig. 7.14, where S is calculated by $PSD_{noise} \cdot B/G$, PSD_{noise} is the measured output noise power spectrum density from spectrum analyzer, B and G are the receiver resolution bandwidth and conversion gain, respectively. Due to the loss of waveguide and probe (~ 15 dB) at LO input, the maximum LO power allowed at the mixer input is about -25dBm, which largely affects the receiver performance in terms of conversion gain and sensitivity. According to relation between conversion gain and LO power illustrated in Fig. 7.9(b), the compensated receiver gain is -2 dB when LO power is increased to 0 dBm. Similarly, the receiver sensitivity in the 0-dBm LO condition is improved to -54.4 dBm as illustrated in Fig. 7.14. Moreover, the receiver sensitivity can be further improved by introducing off-chip filters with even smaller resolution bandwidth. For example, as shown in Fig. 7.16, a -104 dBm sensitivity can be achieved at 250 GHz when the resolution bandwidth is reduced to 1kHz. Note that the maximum imager data rate is determined by the integration time of each pixel, and can be derived from the resolution bandwidth (RBW) based on the selected low pass filtering response. For example, the integration time of a single RC low pass filter is $0.35/RBW(1/Hz)$.

The receiver performance is summarized in TABLE 7.1 and compared to other recent state-of-art CMOS THz image receivers. For the first time, a CMOS based THz image system is demonstrated by the heterodyne receiver with on-chip integrated circular-polarized SIW antenna. The proposed receiver has much smaller detection resolution bandwidth when compared the other detection method. Especially when comparing to the super-regenerative based receiver designs with resonant-type narrow band detection, the resolution bandwidth is further increased by 15 times, while the system bandwidth is improved by 30 times. Moreover, the sensitivity of proposed receiver is comparable to the designs in other receiving topologies [129, 132] when a 0-dBm LO power is applied.

Figure 7.15: Gain measurement results when sweeping RF and LO frequencies with $F_{IF} = F_{LO} + 280GHz$.

TABLE 7.1: STATE-OF-THE-ART CMOS THZ IMAGE RECEIVERS PERFORMANCE COMPARISON

Parameters	Unit	This Work	[24]	[131]	[129]	[113]	[132]
Technology	-	65-nm CMOS	130-nm CMOS	130-nm CMOS	180-nm SiGe BiC-MOS	65-nm CMOS	65-nm CMOS
Frequency	GHz	239 ~ 281	280	280	93-113	183	201
Detection Method	-	Heterodyne	Diode-detection	Diode-detection	Diode-detection	Super-regenerative	Super-regenerative
Detection Polarization	-	Circular	Linear	Linear	Linear	Linear	Linear
System Bandwidth	GHz	42	7	700	20	1.4	1.5
Resolution Bandwidth	GHz	0.1	7	700	20	1.4	1.5
Gain	dB	-25/-2*	-	31	39	-	-
Sensitivity	dBm	-31.4/-54.4*	-26.9	-	-56	-72.5	-59.6
Power	mW/pixel	6.6	2.5	0.1	225	13.5	18.2
Chip Area	mm ² /pixel	0.03	3.8	0.25	0.29	0.45	0.99

*calculated results when 0-dBm LO power is applied to the mixer.

Figure 7.16: Receiver sensitivity at 250 GHz versus receiver resolution bandwidth.

7.3.5 Wide band THz imaging

The THz image system is set up as shown in Fig. 7.17 with samples placed between the horn antenna and receiver chip. The samples under-test are held by X-Y moving stage, controlled by the testing program. The proposed THz image system shown is applied to study Panadol pills and animal skin sample in dry and moisturized conditions at 240 GHz and 280 GHz, respectively. The samples are placed between the antenna of transmitter and receiver chip. The samples are held by X-Y moving stage controlled by the testing program. Fig. 7.18 shows two imaging cases for the biomedical applications. The first case shows that one can differentiate between a moisturized Panadol pill and a dry one because of strong water absorption at THz frequencies. In the second case, the moisturized area in animal skin sample can be clearly identified from the surrounding dry area. Moreover, different images are obtained at 240 GHz and 280 GHz with different absorption ratios.

Figure 7.17: Measurement setup of THz image system.

7.4 280-GHz Reflective Imaging System

Fig. 7.19 shows the block diagram of the proposed CMOS based THz reflective imaging system. The entire system consists of two identical transceivers with on-chip antennas that each of them can either generate or detect a THz signal.

When the transceiver works as a signal source, the signal generated from CMOS local oscillator (LO) is directly fed to the on-chip antenna by a T-line network with high characteristic impedance (high- Z_0). The application of high- Z_0 T-line has two advantages compared to the conventional T-line with $50\ \Omega$ impedance. Firstly, for the signals traveling in the T-line with the same power level, higher Z_0 generates larger voltage magnitude, which can drive the gate of mixer more effectively. Secondly, the signal propagation loss can be effectively reduced due to a smaller current magnitude in the high- Z_0 T-line.

When the transceiver works as a signal detector, the incoming THz signal (RF) is firstly received by the on-chip antenna, then it is down-converted by the single-gate mixers attached to both ends of T-line network. Because both the LO and RF signals travel in the opposite directions with 90° phase shift in the T-line, a differential IF signal is resulted. Note that the LO frequency is usually slightly different from the RF frequency to have a IF output in the mixer. Compared to the single-ended mixer design introduced in Sec. 7.3.2, a differential mixer output in base-band provide stronger immunity to the EM interference as well as common noise rejection. After further amplification by variable gain amplifier (VGA), both magnitude and phase information of the resulting differential IF signal can be detected and processed for

Figure 7.18: Captured THz image results of Panadol pills and skin samples under 240 GHz and 280 GHz radiation.

imaging applications.

In the following sections, the design of each building block in the 280-GHz CMOS transceiver is discussed, including a high power CON based signal source, a high gain 2D CRLH T-line based LWA array and a differential down-conversion mixer with VGA.

7.4.1 280-GHz Injection-locked CON Signal Source

Fig. 7.20 shows the block diagram of the proposed 280-GHz signal source. The input of the proposed 280GHz signal source is a 17.5-GHz reference signal, of which the signal frequency is four times increased by a frequency quadrupler. Based on the 35 GHz to 70 GHz input reference frequency doubler in Sec. 4.5.4, the proposed frequency quadrupler is designed with another push-push frequency doubler with differential signal input at 17.5 GHz. The resulting differential 70-GHz reference signal is amplified by the injection-locking of a 70-GHz zero-phase CON with two oscillator unit-cells, each of which has already been introduced in Sec. 4.5.1. Similar to the 140-GHz signal source in Sec. 4.5, the output signals of two 70-GHz zero-phase oscillator unit-cells are firstly frequency-doubled. Then the resulting two in-phase 140GHz signals

Figure 7.19: The block diagram of CMOS based THz reflective imaging system.

are combined at the center of 70-GHz CON to generate a two times higher 140-GHz output power. The combined single-ended 140-GHz reference signal is further split and converted into two in-phase differential 140-GHz reference signal by transformer based baluns. Each differential 140GHz reference signal is further amplified by injection-lock of a 140-GHz zero-phase CON with four oscillator unit-cells. The output signals of four 140GHz zero-phase oscillator unit-cells in each 140GHz CON are also frequency-doubled. Then the resulting four in-phase 280-GHz signals are combined at the center of each 140-GHz CON to generate a four times higher output power at 280 GHz. Two resulting in-phase 280-GHz output signals are finally combined to have a total eight times higher output power at 280 GHz. The design of each circuit block is shown in following section.

7.4.1.1 Zero-phase Oscillator Unit-cell at 140 GHz

Fig. 7.21 shows the schematic and layout of on-chip 140-GHz MPW based oscillator unit-cell with inter-digital coupled T-line implemented in the top most copper layer (M8) and parasitic capacitances from transistors in 65-nm CMOS process. Compared

Figure 7.20: Block diagram of proposed 280GHz signal source.

to the 70-GHz oscillator unit-cell in Sec. 4.5.1, the size of 140-GHz oscillator unit-cell is about 50% smaller due to a shorter wavelength at 280 GHz. The unit-cell EM-simulation results are shown in Fig. 7.22, where a very small insertion loss of 1 dB is observed in zero-phase mode at 140 GHz. Note that a parasitic capacitance of 32 fF from active devices is also considered in the simulation.

7.4.1.2 140-GHz Zero-phase Coupled Oscillator Network with Output Frequency Doublers

The schematic of the 140-GHz CON is shown in Fig. 7.23(a). Similar to the design of 70-GHz zero-phase CON, four 140GHz MPW based oscillator unit-cells are serially connected in a closed-loop form to generate four in-phase differential output signals at locations A, B, C and D with the same magnitude and frequency, which is injection-locked to the 140-GHz reference signal. The oscillation signal is generated by compensating the energy loss in each unit-cell with a negative resistance formed by

Figure 7.21: Schematic of on-chip MPW base oscillator unit-cell at 140-GHz band with inter-digital coupled T-line.

Figure 7.22: EM simulation results of 140-GHz on-chip MPW unit-cell with inter-digital coupled T-line.

cross-coupled NMOS pair (e.g. M1 and M2). Different from the 70-GHz CON with external push-push frequency doubler, the 2nd harmonics of differential signal in 140-GHz CON are directly extracted from the tail of cross-coupled NMOS pair as shown in Fig. 7.23(b). This method eliminates the parasitic capacitance contributed by the

push-push frequency doubler, and enables the CON to operate at a higher frequency such as 140GHz. However, there are several considerations when applying this method. Firstly, the biasing network such as a one-side shorted $\lambda/4$ T-line is required at the tail of each cross-coupled NMOS pair with small resistance at DC and high impedance at the output frequency. As such, two $\lambda/4$ T-lines at 280 GHz are connected in parallel to increase the power efficiency of the 140-GHz CON. Secondly, the output impedance of each cross-coupled NMOS pair cannot be flexibly adjusted, because the size of NMOS transistors are usually maximized to have a large output current. As such, additional T-line based impedance transformer at 280 GHz is required to convert the output impedance to the targeted value.

Figure 7.23: (a) Schematic of injection-locked 140GHz CON with 4 MPW unit-cells; (b) 2nd harmonic outputs power combining network.

The optimum output impedance at point 'a' is found to be 18Ω by the load-pull analysis in the post-layout simulation. It is further matched to the single size output impedance of 152Ω at point 'b' by a $\lambda/4$ T-line with characteristic impedance of 52.3

Ω . Note that the resulting output impedance for the entire 280-GHz signal source is 76Ω , which is half of the impedance of the single branch due to the parallel in-phase power combination. Compared to the conventional 50Ω system, a 76Ω system has a larger voltage to drive the gate of a CMOS transistor as well as a smaller current to reduce the propagation loss.

7.4.1.3 70-GHz Zero-phase Coupled Oscillator Network with Inter-stage Frequency Doublers

The schematic of the closed-loop 70-GHz CON is shown in Fig. 7.24. Since this 70GHz CON functions as an inter-stage buffer with less stringent output power requirement, a minimum number of two MPW based oscillator unit-cells is deployed to form a loop with a central symmetrical layout. Compared to the 70-GHz CON design with four oscillator unit-cells in Sec. 4.5.1, the output power is reduced by half as well as the power consumption. As such a output power of around 0 dBm can be ensured at 140 GHz.

Figure 7.24: Schematic of injection-locked 70GHz CON with 2 MPW unit-cells.

A 140-GHz power splitter is designed to convert the single-ended 140GHz reference signal output into two identical differential signals. The layout of the proposed power splitter is shown in Fig. 7.25(a). The inlet single-ended 140-GHz signal (P1) is firstly splitted into two path at point 'a', each of the resulting single-ended 140GHz signals is further converted into differential signal by a compact transformer based balun ($25 \times 25 \mu\text{m}^2$) due to the limited inter-stage chip area. The primary and secondary loops of the transformer based baluns are designed in the top most (M8) and second top most (M7) copper layers, respectively. The grounding layer (M1) is removed under the balun to enhance the inductive coupling between the primary and secondary

loops. The phase and magnitude mismatch between P2 and P3 (P4 and P5) can be improved by adjusting the trace length and AC-GND locations of the secondary coils, respectively. A symmetric layout of the whole splitter ensure the magnitude and phase balance among differential port(P2-P3 and P4-P5) with $S_{21} = S_{41}$ and $S_{41} = S_{51}$. As verified by EM simulation in Fig. 7.25(b), the proposed the proposed power splitter has an average intrinsic loss of 9 dB at 140 GHz. The magnitude and phase mismatches at 140 GHz are only 0.7 dB and 5.8 degrees, respectively.

Figure 7.25: 140GHz power splitter from one single-ended input to two differential outputs: (a) On-chip layout, (b) EM-simulation results.

7.4.1.4 17.5-GHz to 70-GHz Input Reference Frequency Quadrupler

Fig. 7.26 shows the schematic of the 17.5 GHz to 70 GHz input reference frequency quadrupler. Based on the frequency doubler discussed in Sec. 4.5.1, the proposed quadrupler is designed with additional push-push frequency doubler from 17.5 GHz to 35 GHz. One transformer based balun is deployed to generate a differential 17.5 GHz reference signal to drive M1 and M2. Fig. 7.27 shows the post-layout simulation results of the entire frequency quadrupler with 0-dBm reference power. The conversion gain is above -20 dB in 16 ~ 19 GHz. Due to the low coupling factor between the primary and the secondary coil in on-chip transformer based balun at 17.5 GHz, the input S11 is only smaller than -6 dB in 16 ~ 17.3 GHz. This can be resolved by the additional matching network in PCB.

Figure 7.26: Input reference frequency quadrupler from 17.5 GHz to 70 GHz.

7.4.1.5 Simulation Results and Discussion

The proposed injection-locked THz signal source is implemented in 65nm CMOS RF process with the Cadence layout shown in Fig. 7.28. It has a total area of $750 \times 550 \mu\text{m}^2$, and a core area of 0.12 mm^2 including CONs at 70 GHz and 140 GHz. A post layout simulation is conducted to evaluate the performance of the proposed 280-GHz signal source. Operating from a 1.2-V power supply, the core of signal source consumes 288 mW, while the input frequency quadrupler consumes another 7.6 mW.

Fig. 7.29 shows the simulation results of output power. By adjusting the reference signal around 17.5 GHz with 0-dBm power level, a 10.5% tuning range centered at 286 GHz is obtained. The maximum output power of 1.9 mW is observed at 276 GHz with a DC-RF efficiency of 0.66%. Moreover, the maximum power density of the proposed

Figure 7.27: Post-layout simulation results of 17.5 GHz to 70 GHz quadrupler.

TABLE 7.2: PERFORMANCE COMPARISON WITH RECENTLY PUBLISHED THZ SIGNAL SOURCES

Parameters	[148]	[149]	[150]	[151]	This Work (Simulated)
Technology	65-nm CMOS	65-nm CMOS	65-nm CMOS	65-nm CMOS	65-nm CMOS
Center Frequency (GHz)	290	294	288	256	286
FTR (%)	4.5	1.1	1.7	6.5	10.5
P_{OUT} (dBm)	-1.2	-3.3	-1.5	4.1	2.8
Integrated DC Bias Network	NO	YES	NO	NO	YES
DC Power (mW)	325	258	275	227	288
Power Efficiency (%)	0.23	0.19	0.25	1.14	0.66
A_{CORE} (μm^2)	600 $\times$ 600	630 $\times$ 330	240 $\times$ 150*	640 $\times$ 470	180 $\times$ 340
P_{OUT}/A_{CORE} (mw/mm 2)	2.1	2.3	19.7	8.2	31.1

*The area of the power combining network is included in the core area calculation.

signal source is 31.1 mW/mm 2 . The performance of the proposed 280 GHz signal source is summarized in TABLE 7.2 with comparison to the recent state-of-the-art THz signal source designs in CMOS process. It can be observed that the proposed 280 GHz signal source has the highest power density as well as the outstanding output

Figure 7.28: Cadence layout of the proposed 280GHz source in CMOS.

power and power efficiency performance. The output power of the proposed design is slightly lower than [151] due to integration of DC bias network, which will at least contribute 1.5-dB loss to the output signal. Note that a signal source with DC Bias Network can be directly used for on-chip integration with following function blocks such as mixer or antenna.

7.4.2 Differential Down-conversion Receiver

7.4.2.1 Differential Down-conversion Mixer with bidirectional hybrid coupler

Fig. 7.30 shows the schematic of the proposed 280 GHz differential down-conversion mixer. A 90° phase shifter is designed for both LO and RF signals with high- Z_0 T-lines implemented by coplanar waveguide (CPW) as well as the parasitics capacitances contributed by M1, M2 and R1. The size of both M1 and M2 is optimized between the down-conversion efficiency and the capacitance loadings to the high- Z_0 T-line. For instance, a smaller size of M1 and M2 helps reduce the loaded capacitance to high- Z_0

Figure 7.29: Simulated output power of the proposed 280GHz source in CMOS.

T-line, but it also reduces the output currents of mixing products. Moreover, they are biased in the subthreshold region ($0.4\text{-V } V_{GS}$) by a diode-connected NMOS transistor (M5) to maximize the nonlinearity. Two $\lambda/4$ T-line open stubs at 280 GHz are connected to the drains of M1 and M2 to improve the down-conversion efficiency by reducing the LO leakage to the following common-gate stages (M3 and M4).

Assuming the frequencies of both RF and LO signals are closed to each other than both of them have 90° phase shift without any signal loss, equal amount of LO or RF voltages are applied to the gate of M1 and M2 with a 90° delayed version of each other that can be expressed as:

$$\begin{cases} V_{M1} = V_{RF} \cdot \cos(\omega_{RF}t + \phi_{RF} + 90^\circ) + V_{LO} \cdot \cos(\omega_{LO}t + \phi_{LO}) \\ V_{M2} = V_{RF} \cdot \cos(\omega_{RF}t + \phi_{RF}) + V_{LO} \cdot \cos(\omega_{LO}t + \phi_{LO} + 90^\circ) \end{cases} \quad (7.7)$$

where $[V_{RF}, \omega_{RF}, \phi_{RF}]$ and $[V_{LO}, \omega_{LO}, \phi_{LO}]$ are the [input voltage magnitude, frequency, initial phase] of RF and LO signal, respectively.

Note that the output voltage of a single-gate mixer can be expressed as

$$\begin{cases} V_{IF}(t) = R_0 g_m(t) V_{RF}(t) \\ g_m(t) = a_0 + \sum_{n=1}^{\infty} a_n \cos(n\omega_{LO}t) \end{cases} \quad (7.8)$$

Figure 7.30: Schematic of proposed differential down-conversion mixer with a input network of 90° high- Z_0 T-line.

where a_n , $n=(0,1,2,\dots)$ are the Fourier coefficients of g_m with respect to ω_{LO} . It can be shown that a_0 and a_1 represent the fundamental transconductance and the first-order mixing product of $(\omega_{RF} - \omega_{LO})$, respectively. As such, by substituting (7.7) into (7.8), the first-order mixing product at IF outputs become:

$$\begin{cases} VP_{IF} = a_1 \cdot V_{RF} R_0 \cos [\omega_{IF} t + (\phi_{RF} - \phi_{LO}) + 90^\circ] \\ VN_{IF} = a_1 \cdot V_{RF} R_0 \cos [\omega_{IF} t + (\phi_{RF} - \phi_{LO}) - 90^\circ] \end{cases} \quad (7.9)$$

where $\omega_{IF} = \omega_{RF} - \omega_{LO}$ is the IF frequency. A differential IF output is observed from (7.9) that VP_{IF} and VN_{IF} have the same magnitude and opposite phase. As a result, the total output voltage of mixer (V_{IF}) is

$$V_{IF} = VP_{IF} - VN_{IF} = 2a_1 \cdot V_{RF} R_0 \cos [\omega_{IF} t + (\phi_{RF} - \phi_{LO}) + 90^\circ] \quad (7.10)$$

Eq. 7.10 indicates that both magnitude and phase information of RF signal can be obtained by the proposed down-conversion mixer, which are very useful in the refractive index measurement by a THz imaging system.

A post-layout simulation is performed to study the proposed differential mixer design. Fig. 7.31 shows the 2-port S-parameters analysis of the high- Z_0 T-line network under 76Ω system impedance. A good input matching is observed with S11 smaller than -10dB in 220-340GHz. The maximum insertion loss (S21) in 220 ~ 340 GHz

Figure 7.31: Post-layout simulation results of the high- Z_0 T-line network with 90° phase delay.

is 1.6dB. A wide band 90° phase shift is observed with ± 10 degrees bandwidth of 55 GHz centered at 280 GHz. Fig. 7.32(a) shows the post-layout simulation results of the 280-GHz differential mixer from 260 GHz to 320 GHz under the following conditions: the power of LO signal is fixed at 0 dBm; the frequency of RF signal is 1GHz above the frequency LO signal. An averaged conversion gain of -4.6 dB is observed at 280GHz for VP_{IF} and VN_{IF} , which have magnitude and phase mismatch of 0.2 dB and -0.7 degrees, respectively. Note that the phase mismatch has already been normalized to 180 degrees.

7.4.2.2 Variable Gain Amplifier

In this work, a modified Cherry-Hooper amplifier based VGA [7] is employed to boost the power of IF signals from mixer outputs with low power consumption, compact design size as well as large gain control range. Fig. 7.32(b) shows the post-layout simulation results of the entire receiving part after integrating mixer and VGA under the following conditions: the power of LO signal is fixed at 0 dBm; the frequency of RF signal is 1GHz above the frequency LO signal. The gain and noise figure(NF) observed at 280GHz is 46.6 dB and 24.6 dB, respectively. And the variations of gain and NF are both less than 2dB in 272 ~ 302 GHz. The 3dB IF bandwidth is 1.1 GHz,

(a)

(b)

Figure 7.32: Post-layout simulation results: (a) the conversion gain and output phase mismatch of the 280-GHz differential mixer; (b) the conversion gain and NF of the 280-GHz receiver with differential mixer and VGA from [7].

which is mainly determined by VGA [7].

Figure 7.33: 2D LWA array with 2×13 unit-cells.

7.4.3 2D On-chip Leaky Wave Antenna Array

A 2D antenna array with on-chip LWA with 2×13 CRLH T-line unit-cells is designed in 65-nm CMOS process as shown in Fig. 7.33. Two 1D CRLH T-line based LWAs introduced in Sec. 6.3 are connected in parallel by a T-junction to further increase the broadside antenna gain as well as reduce the end-fire leakage. The antenna input is matched to the system characteristic impedance of 76Ω by connecting a coplanar waveguide (CPW) with a length of $80 \mu\text{m}$, which is implemented in layer M8. A standard high resistivity silicon layer ($1000 \times 300 \mu\text{m}^2$) with a thickness of $100 \mu\text{m}$ is also placed on top of the antenna surface to enhance the radiation efficiency of antenna array.

The proposed 2D LWA array is verified by a full wave simulation in Ansoft HFSS. Fig. 7.34(a) shows the simulated radiation pattern at 280 GHz. It has a broadside radiation pattern with a directivity of 9.1 dBi and a radiation efficiency of 41%. As a result, the half-power bandwidth (HPBW) is ± 20 degrees in the ZOY plane as illustrated in Fig. 7.34(b); and a broadside radiation gain of 5 dBi (in Z direction) as well as an end-fire leakage of -16 dBi (in X direction) are obtained. Compared to the 1D LWA demonstrated in Sec. 6.3, 2D LWA has 0.9 dB higher gain at 280 GHz. This improvement is lower than the ideal value of 3dB because of the additional loss introduced by the matching network as well as the T-junction. Fig. 7.35 shows the simulation results of input S11 as well as the wide band antenna gain on the broadside direction. The S11 is observed to be smaller than -6 dB from 256 GHz to 310 GHz; a 47-GHz 3-dB bandwidth centered at 279 GHz is observed from 255 GHz to 302 GHz.

7.4.4 Transceiver Integration

The proposed 280-GHz CMOS transceiver Design is implemented in 65-nm CMOS RF process with the Cadence layout shown in Fig. 7.36. It has a total area of $1000 \times 1010 \mu\text{m}^2$. Note that there are no measurements for the proposed 280-GHz transceiver, and the performance of transceiver is verified by post-layout and EM simulation. Operating from a 1.2-V power supply, the whole transceiver consumes 298.6 mW, including 3.5 mW contributed by the 280 GHz differential mixer and VGA.

The performance of the proposed THz source is summarized in TABLE 7.3 with comparison to the recent state-of-the-art THz transmitters and receivers in CMOS. It can be observed that the proposed 280 GHz transmitter has the highest output power and power efficiency. A transmitter with very high equivalent isotropically radiated power (EIRP) density is developed by the integration of the 280GHz CON based signal source with high power density and the compact CRLH T-line based LWA with high gain. Note that the EIRP density is defined by the EIRP power generated in unit chip area of transmitter (P_{EIRP}/A_{TX}). The proposed transmitter has an EIRP density of $4.27 \text{ mW}/\text{mm}^2$, which is more than twice of the best result in the literature [150].

Moreover, the proposed 280 GHz receiver with differential output has a 53 dB higher gain, 13.6 dB smaller NF and a more than twice smaller power consumption when compared to the single-ended receiver proposed in Sec. 7.3. As a result of the compact differential mixer and VGA designs, a high power signal source is integrated in the receiver without increasing the chip size. The sensitivity of the proposed receiver is -57.6 dBm, which is quite close to the result of the super-regenerative receiver design in [152]. Note that the sensitivity can be further improved by additional off-chip IF filters with smaller bandwidth.

7.5 Conclusion

CMOS based imaging systems are demonstrated in this chapter. The proposed 135-GHz SRX is integrated in sub-THz imaging system with various demonstrated imaging diagnosis applications. It has great potential to be utilized for the future large-arrayed transmission-type THz imaging system. In addition, a wide band THz image system based on direct-conversion receiver is demonstrated in CMOS process with wide detection frequency range. The proposed THz image system is able to capture images in $239 \sim 281 \text{ GHz}$ with a resolution bandwidth of 100MHz, which has

TABLE 7.3: PERFORMANCE COMPARISON WITH RECENTLY PUBLISHED THZ TRANSMITTERS AND RECEIVERS

Transmitters					
Parameters	Unit	[23]	[153]	[150]	This Work
Technology	-	CMOS 45nm SOI	CMOS 65nm	CMOS 65nm	CMOS 65nm
Center Frequency	GHz	280	260	288	286
P_{OUT}	dBm/Pixel	-7.2	0.5	-4.1	1.3
Broadside P_{EIRP}	dBm	9.4	15.7	14.2	6.3
FTR	%	3.2	9.5	1.7	10.5
DC Power	mW/Pixel	52.25	800	275	295.6
Power Efficiency	%	0.37	0.14	0.14	0.46
Transmitter Size* (A_{TX})	mm ²	2.7×2.7	78.5	12.56	1×1.01
P_{EIRP}/A_{TX}	mw/mm ²	1.19	0.47	2.09	4.27
Receivers					
Parameters	Unit	[131]	[152]	Receiver in Sec. 7.3	This Work
Technology	-	CMOS 0.13 μ m	CMOS 65nm	CMOS 65nm	CMOS 65nm
Center Frequency	GHz	280	201	260	286
Detection Method	-	Diode-detection	Super-regenerative	Heterodyne with Single-ended output	Heterodyne with Differential output
Gain	dBi	31	-	-2	54
NF	dB	-	-	38.6	25
System Bandwidth	GHz	700	1.5	42	30
Resolution Bandwidth	GHz	700	1.5	0.1	1.1
DC Power	mW/Pixel	0.1	18.2	6.6	3.5
Receiver Size	mm ²	3.8	0.45	0.99	1.01

*The area of silicon lens is included in the transmitter size calculation.

many applications in the detection of tissue with species-specific spectral absorption. An integrated THz CMOS transceiver is demonstrated for 280GHz reflection imaging system with high power transmitter, high sensitivity receiver and high gain on-chip in simulation. The high power transmitter is designed by connecting the 2nd harmonic outputs of two 140GHz CONs in parallel, which are both injection locked with the same phase and magnitude. It is simulated with an output power of +2.8 dBm, a power efficiency of 0.66%, and a frequency tuning range of 10.5% from 272 GHz to 302 GHz. The 2D on-chip high-gain LWA array is designed by connecting two 1D LWAs in parallel with 2×13 unit-cells, and it is simulated with a broadside radiation pattern with 9.1-dBi directivity and 41% radiation efficiency at 280 GHz. The differential

down-conversion receiver is designed by integrating a differential single-gate mixer with one modified Cherry-Hooper amplifier based variable-gain amplifier with compact size, and it is simulated with a conversion gain of 46.6 dB, and a NF of 24.6 dB at 280GHz. The entire transceiver has a compact size of 1 mm², and consumes 298.6 mW power operating under 1.2V power supply. The transmitter is simulated with an equivalent isotropically radiated power (EIRP) power of 6.3 dBm, a EIRP density of 4.27 mW/mm²; the receiver is simulated with a maximum gain of 51dB and a sensitivity of -57.6 dBm.

(a)

(b)

Figure 7.34: HFSS simulated radiation pattern of the 2D LWA array at 280 GHz in: (a) 3D plot, and (b) polar plots in ZOY and ZOX planes.

Figure 7.35: Simulated antenna input S11 and gain at broadside direction (Z-axis).

Figure 7.36: Cadence layout of the proposed 280GHz transceiver in CMOS

Chapter 8

Conclusion and Future Work

8.1 Conclusion

The main contribution of this work is accomplished by the transformation IC designs with high-Q passive structures such as metamaterial. A metamaterial based CMOS transceiver for THz imaging is proposed with significantly improved performance in the THz signal generation, transmission and detection. As explored in this work, on-chip metamaterial structures are demonstrated with several advantages over the conventional approaches based on T-lines or LC-Tanks. Non-resonant type on-chip metamaterials, such as magnetic plasmon waveguide (MPW) or composite right/left handed T-line (CRLH-TL), can create a low loss medium with zero-phase propagation. MPW is mainly explored in the CMOS THz signal source design with improved output power and power efficiency as well as wide frequency tuning range (FTR) and compact size. CRLH-TL is mainly explored in the CMOS THz antenna design with improved radiation gain and efficiency within compact chip area. Resonant type on-chip metamaterials, such as T-line loaded with split ring resonator (TL-SRR) and complementary split ring resonator (TL-CSRR), can effectively increase the energy storage within a very compact area and resulting a much higher Q-factor. They are mainly explored in the CMOS THz receiver design with high sensitivity super-regenerative receiver. With the proposed transmitter and receiver designs, both narrow and wide band THz imaging systems can be demonstrated at 135GHz and 280GHz, respectively.

In the 135GHz transmission-type imaging system, a high power transmitter is designed by MPW based zero-phase CON with four in-phase coupled unit-cells; a high sensitivity super-regenerative receiver (SRX) is designed by quench-controlled metamaterial high-Q oscillator with TL-SRR. The transmitter is measured with 3.5mW peak output power, 2.4% power efficiency, 26.9mW/mm² power density, 9.7% FTR

centered at 133.5GHz with a compact core chip area of 0.13mm². The measured results are: The TL-SRR based SRX is measured with -76.8dBm sensitivity, 9.7dB NF, 0.9fW/ $\sqrt{Hz}$ noise-equivalent power, 6.2mW power consumption with a compact core chip area of 0.0085mm². in addition, various imaging applications are demonstrated by the proposed 135GHz imaging system.

In the wide band 280GHz transmission-type imaging system, a high spectrum resolution and high sensitivity CMOS imager is designed by a 239-281GHz direct-conversion receiver with circular-polarized substrate integrated waveguide (SIW) antenna. The imager is measured with -2dB conversion gain, 42GHz bandwidth, -54.4dBm sensitivity at 100MHz detection resolution bandwidth, 6.6mW power consumption and 0.99mm² chip area. Moreover, frequency dependent biomedical imaging applications are demonstrated by the proposed receiver.

In the wide band 280GHz reflection imaging system, a high power transmitter is designed by connecting the 2nd harmonic outputs of two 140GHz CONs in parallel, which are both injection locked with the same phase and magnitude. It is simulated with an output power of +2.8dBm, a power efficiency of 0.66%, and a frequency tuning range of 10.5% from 272GHz to 302GHz. A 2D on-chip high-gain LWA array is designed by connecting two 1D LWAs in parallel with 2×13 unit-cells. The proposed 2D LWA array is simulated with a broadside radiation pattern with 9.1dBi directivity and 41% radiation efficiency at 280GHz. The differential down-conversion receiver is designed by integrating a differential single-gate mixer with one modified Cherry-Hooper amplifier based variable-gain amplifier with compact size. A bidirectional hybrid coupler with high impedance T-line is proposed to provide 90° phase shift for both RF and LO signals at the mixer input as well as to bypass the LO signal. The proposed receiver is simulated with a conversion gain of 46.6dB and a NF of 24.6dB at 280GHz. The whole transceiver has a compact size of 1mm², and consumes 298.6mW power operating under 1.2V power supply. As verified by post-layout simulations, the transmitter has an equivalent isotropically radiated power (EIRP) power of 6.3 dBm, an EIRP density of 4.27mW/mm²; the receiver has a maximum gain of 51dBi and a sensitivity of -57.6dBm.

8.2 Recommendation for Future Work

Based on above works, there are two recommended future works for this thesis.

The first recommended work is to fabricate and measure the 280GHz transceiver for reflection-type imaging. Moreover, the formation of the 280GHz transceiver into

a 1D or 2D array should be explored. A sensor head can be constructed by multiple transceiver cells formed in a 1D or 2D array to largely increase the imaging speed.

The second recommended work is to further increase transceiver frequency in THz region as well as to reduce the transceiver size. With the rapid scaling of CMOS technology, the f_{MAX} of CMOS transistors could be above 1THz very soon. In such case, a THz transceiver could be designed at a higher frequency with even smaller pixel size. With the help of metamaterial based transformational IC design techniques, a fully integrated THz CMOS image sensor array is becoming possible to be placed into the hand-held devices such as mobile phones or tablet PCs.

Appendix A

Author's Publication

A.1 Journal

1. **Yang Shang**, Hao Yu*, Yuan Liang, Xiaojun Bi, and Muthukumaraswamy Annamalai, High Output Power Millimeter-wave Signal Sources at 60 GHz and 140 GHz by In-phase Coupled Oscillator Network in 65-nm CMOS, *IEEE Journal of Solid-State Circuits (JSSC)*. (Submitted)
2. **Yang Shang**, Hao Yu*, Sanming Hu, Yuan Liang, Xiaojun Bi, and Muthukumaraswamy Annamalai, High-Sensitivity CMOS Super-Regenerative Receiver with Quench-Controlled High- Q Metamaterial Resonator for Millimeter-Wave Imaging at 96 and 135 GHz, *IEEE Transactions on Microwave Theory and Techniques (TMTT)*, vol.PP, no.99, pp.1 - 12, Oct. 2014.
3. **Yang Shang**, Hao Yu*, Haipeng Fu, and Wei Meng Lim, A 239-281GHz CMOS Direct-conversion Receiver with On-chip Circular-polarized Substrate Integrated Waveguide Antenna for Sub-terahertz Imaging, *IEEE Transactions on Terahertz Science and Technology (TTST)*, vol.4, no.6, pp.686 - 695, Sept. 2014.
4. **Yang Shang**, Hao Yu* and Wei Fei, Design and Analysis of CMOS based Terahertz Integrated Circuits by Causal Fractional-order RLGC Transmission Line Model, *IEEE Journal on Emerging and Selected Topics in Circuits and Systems (JETCAS)*, vol.3, no.3, pp.355-366, Sept. 2013.
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9. Wei Fei, Hao Yu*, **Yang Shang**, Deyun Cai, and Junyan Ren, A 96 GHz Oscillator by High-Q Differential Transmission Line loaded with Complementary Split Ring Resonator in 65nm CMOS, *IEEE Transactions on Circuits and Systems II (TCAS-II)*, vol.60, no.3, pp.127-131, March 2013.
10. Wei Fei, Hao Yu*, **Yang Shang**, and Kiat Seng Yeo, A 2D Distributed Power Combining by Metamaterial-based Zero-Phase-Shifter for 60GHz Power Amplifier in 65nm CMOS, *IEEE Transactions on Microwave Theory and Techniques (TMTT)*, vol.61, no.1, pp.505-516, Jan. 2013.
11. Sai Manoj P.D., Hao Yu*, **Yang Shang**, Chuan Seng Tan, and Sung Kyu Lim, Reliable 3D Clock-tree Synthesis Considering Nonlinear Capacitive TSV Model with Electrical-thermal-mechanical Coupling, *IEEE Transactions on Computer-Aided Design of Integrated Circuits and Systems (TCAD)*, vol.32, no.11, pp.1734-1747, Nov. 2013.

A.2 Conference

1. **Yang Shang**, Hao Yu*, Peng Li, Xiaojun Bi, Minkyu Je, A 127-140GHz Injection-locked Signal Source with 3.5mW Peak Output Power by Zero-phase Coupled Oscillator Network in 65nm CMOS, *IEEE Custom Integrated Circuits Conference (CICC)*, Sept. 2014.
2. **Yang Shang**, Hao Yu*, Chang Yang, Yuan Liang and Wei Meng Lim, A 239-281GHz Sub-THz Imager with 100MHz Resolution by CMOS Direct-conversion Receiver with On-chip Circular-polarized SIW Antenna, *IEEE Custom Integrated Circuits Conference (CICC)*, Sept. 2014.
3. **Yang Shang**, Hao Yu, Sanming Hu, Chang Yang, Minkyu Je, A High-Sensitivity 135GHz Millimeter-Wave Imager by Differential Transmission-Line Loaded Split-Ring-Resonator in 65nm CMOS, *European Solid-State Device Research Conference (ESSDERC)*, Sept. 2014.
4. **Yang Shang**, Hao Yu, Peng Li, Yuan Liang, and Chang Yang, A 280GHz CMOS On-chip Composite Right/Left Handed Transmission Line based Leaky Wave Antenna with Broadside Radiation, *IEEE International Microwave Symposium (IMS)*, June 2014.

5. **Yang Shang** and Hao Yu*, Design of High Sensitivity CMOS Sub-THz Image Receiver, *IEEE International Solid-State Circuits Conference (ISSCC) Student Research Preview*, Feb. 2014.
6. **Yang Shang** and Hao Yu*, CMOS on-chip metamaterial: new opportunities in mm-wave and THz, *International Conference on Metamaterials, Photonic Crystals and Plasmonics (META)*, May, 2014. (Invited)
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9. **Yang Shang**, Wei Fei, Hao Yu*, Fractional-order RLGC Model for Terahertz Transmission Line, *IEEE International Microwave Symposium (IMS)*, June 2013.
10. **Yang Shang**, Chun Zhang, Hao Yu*, Chuan Seng Tan, Xin Zhao, and Sung Kyu Lim, Thermal-reliable 3D Clock-tree Synthesis Considering Nonlinear Electrical-thermal-coupled TSV, *IEEE/ACM Asia and South Pacific Design Automation Conference (ASP-DAC)*, Jan. 2013.
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12. **Yang Shang**, Wei Fei, and Hao Yu*, Fast Simulation of Hybrid CMOS and STT-MTJ Circuits with Identified Internal State Variables, *IEEE/ACM Asia and South Pacific Design Automation Conference (ASP-DAC)*, Jan. 2012. (Best Paper Award Nomination)
13. Deyun Cai, **Yang Shang**, Hao Yu*, and Junyan Ren, A 76 GHz Oscillator by High-Q Differential Transmission Line Loaded with Split Ring Resonator in 65-nm CMOS, *IEEE Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems (SIRF)*, Jan. 2013.
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15. Yuhao Wang, **Yang Shang**, and Hao Yu*, Design of Single Saw-tooth-pulse based STT-RAM Readout Circuit by NVM SPICE, *IEEE Non-Volatile Memory Technology Symposium (NVMTS)*, Oct. 2012.
16. Chang Yang, Peng Li, Yuan Liang, **Yang Shang**, Wei Fei, and Hao Yu*, A Compact Fractional-order Model for Composite Right/Left Handed Transmission Line at Terahertz, *IEEE International Union of Radio Science General Assembly and Scientific Symposium (URSI GASS)*, Aug. 2014.
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A.3 Book and Book Chapter

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