Air-gapped microaxial transmission line technology for Si-based microwave integrated circuits

Tian, Yang

2017

Tian, Y. (2017). Air-gapped microaxial transmission line technology for Si-based microwave integrated circuits. Doctoral thesis, Nanyang Technological University, Singapore.

<http://hdl.handle.net/10356/69899>

<https://doi.org/10.32657/10356/69899>

**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**AIR-GAPPED MICROCOAXIAL TRANSMISSION
LINE TECHNOLOGY FOR SI-BASED
MICROWAVE INTEGRATED CIRCUITS**

TIAN YANG

SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING

2016

AIR-GAPPED MICROCOAXIAL TRANSMISSION LINE TECHNOLOGY FOR SI-BASED MICROWAVE INTEGRATED CIRCUITS

TIAN YANG

School of Electrical & Electronic Engineering

**A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirements for the degree of**

Doctor of Philosophy

2016

ACKNOWLEDGEMENT

I would like to express my greatest appreciation and the most sincere gratitude to my advisor, Professor Wang Hong, who gave me lots of guidance, support, and motivation. Without his patient and encouragement, I won't be able to overcome the depression and make this thesis possible. He is also a great example in life and has been always more than an advisor to me.

I also sincerely thank Dr. Lee Kokyan for his guidance, and helpful discussion. His mobile phone would always be available to me. His guidance is essential for the successful development of the proposed technology.

I would also like to thanks the MMIC Technology Design Center (MTDC) for providing the facility and technical support, and the colleagues in MTDC for their friendships. Special thanks to the formal colleague, Dr. Liu Zhihong, who was always willing to help. I would like to thank other colleagues, Mr. Ang Kiansiong, Mr. Louis Nicholas Retnam, Ms. Tng Lihuang, Mrs. Meng Qianqian for their insightful discussions for both work and life.

I would also grateful to my formal roommates, Dr. Liu Wenhui and Dr. Liu Tao, for their genuine friendship; I also want to thank Dr. Xu Weijing, Dr. Liang Guozhen, Mr. Zhu Zhidong, Dr. Zhang Qinfeng, Dr. Jin Chen, for doing sports and dining together.

I am indebted to my parents, Mr. Tian Zhijun and Mrs. Yang Jianhua, for their unconditional love and support; my wife, Mrs. Zhang Yilin, for her great love and steadfast support; and finally my baby, Tian Hanzhe Austin, for the happiness that he brings to the family.

LIST OF CONTENTS

ACKNOWLEDGEMENT	i
ABSTRACT.....	ii
LIST OF TABLES.....	iv
LIST OF FIGURES	v
DICTIONARY OF TERMS AND ABBREVIATIONS	xii
1. Introduction.....	1
1.1. Overview of the Chapter	1
1.2. History of Transmission Lines	1
1.3. Transmission Lines for Si-based MICs.....	12
1.3.1. Isolate the planar passives from the substrate	14
1.3.2. Polyimide-core rectangular coaxial structures	28
1.3.3. Air-gapped rectangular coaxial structures.....	32
1.4. Organization of the Thesis	38
2. Development of Air-gapped MCTL Technology	40
2.1. Overview of the Chapter	40
2.2. Transmission Line Modeling and Wave Propagation Modes	40
2.2.1. Electromagnetic-wave propagation phenomena.....	40
2.2.2. Transmission line modelling	43
2.2.3. Wave propagation modes on transmission lines.....	47
a. Transverse Electric and Magnetic (TEM) Mode:	48
b. Transverse Electric (TE) or Transverse Magnetic (TM) modes:.....	49
2.2.4. Finite element analysis (FEA) design.....	51
2.3. MCTL Fabrication Process Development.....	52
2.3.1. Designed process flow.....	52
2.3.2. Material candidates of the main process steps.....	54
a. Sacrificial photoresist for structure patterning.....	54
b. Dielectric material for support straps.....	57
c. Metals for conductor structure	62
2.3.3 Optimized material combination for the MCTLs	64
2.4. Measurement Setups	65
3. Air-gapped MCTLs Characterization and Integration	67
3.1. Overview of the Chapter	67
3.2. Air-gapped MCTLs	67
3.2.1. Structure design of air-gapped MCTLs	67

3.2.2. Measurement of air-gapped MCTLs	70
3.2.3. Power handling capability of air-gapped MCTLs	79
3.3. Integration of Air-gapped MCTLs	82
3.3.1. Design of microcoax-to-CPW transitions.....	83
3.3.2. Electrical performance of microcoax-to-CPW transitions	87
3.3.3. Mechanical strength of microcoax-to-CPW transitions	92
3.3.4. Hybrid integration of the microcoax-to-CPW transition.....	94
4. Microcoaxial Bends and Ultra-long True-time-delay Lines.....	96
4.1. Overview of the Chapter	96
4.2. Right-angled Microcoaxial Bends.....	96
4.2.1. Introduction of right-angled bends	96
4.2.2. Design of microcoaxial bends	97
4.2.3. Results and measurement of the designed bends.....	101
4.3. Ultra-long Microcoaxial True-Time-Delay (TTD) Lines	110
4.3.1. Introduction of TTD line	110
4.3.2. Design of the ultra-long TTD line	111
4.3.3. Measurement results of the ultra-long TTD lines.....	113
5. Microcoaxial Rat-race Hybrid, Wilkinson Power Divider, Cavity Resonators and Filters	119
5.1. Overview of the Chapter	119
5.2. Microcoaxial Rat-race Hybrid.....	119
5.2.1. Introduction of rat-race hybrid	119
5.2.2. Design of microcoaxial rat-race hybrid	120
5.3. Microcoaxial Wilkinson Power Divider	127
5.3.1. Introduction of Wilkinson power divider	127
5.3.2. Design of microcoaxial Wilkinson power divider.....	127
5.4. Cavity Resonators and Filters	133
5.4.1. Introduction of cavity resonators	133
5.4.2. Design of cavity resonators	134
5.4.3. Single-resonator measurement	136
5.4.4. Design of two-pole bandpass cavity filters.....	141
6. Conclusions and Future Work	147
6.1. Overview of the Chapter	147
6.2. Conclusion of the Thesis	147
6.3. Future Work	151
6.3.1. Patch antenna array with Butler matrix network (BMN)	151
6.3.2. Improvement of WPD	151
6.3.3. Microwave system-on-chip (SOC) with package.....	152
Appendix A.....	154

A.1. Detailed process flow of the air-gapped microcoax fabrication	154
Appendix B	160
B.1. Thick sacrificial resist experiment.....	160
B.1.1. KMPR 1025	160
B.1.1.1. Resist lithography	160
B.1.1.2. Adhesion problem in Au plating solution	161
B.1.1.3. Resist Stripping Problem.....	161
B.1.1.4. Edge bead problem	162
B.1.2. AZ 4620	163
B.1.2.1. Round corner	163
B.1.2.2. Incompatiblity with CMP process and resist developer.....	164
B.1.2.3. Electroplating compatibility	164
B.1.2.4. Microcoaxial structure fabrication experiment	165
B.1.2. JSR THB 151N	168
B.1.2.1. Damascene Cu Compatibility	169
B.2. Support resist experiment	169
B.2.1. B-staged BCB	169
B.2.1.1. Lithography	170
B.2.1.2. Cracking	170
B.2.1.3. Remaining Scum	170
B.2.1.4. Roundcorner	172
B.2.2. SU-8	172
B.2.2.1. Low Temperature Process	173
B.2.2.2. Permanent Material	174
Appendix C	176
C.1. Copper and Gold Electroplating Development	176
C.1.1. Plating Setup	176
C.1.1.1. Copper bath setup.....	176
C.1.1.2. Gold bath setup	176
C.1.1.3. Cu plating solution preparation	177
C.1.1.4. Au plating solution preparation.....	177
C.1.1.5. Plating procedure.....	177
C.1.2. Plating Results	178
Publications.....	181
Journal Papers	181

Conference Papers..... 182

Biography..... 183

Page intentionally left blank

LIST OF TABLES

Table 1. 1. Comparison of common transmission lines..... 11

Table 1. 2. Dimensions of micromachined CPWs [65]20

Table 2. 1. Thick sacrificial photoresist candidates..... 56

Table 2. 2. Dielectric material candidates for the support straps.....58

Table 2. 3. Conductivity, resistivity, and temperature coefficient of conventional metals
..... 63

Table 3. 1. Cross-sectional dimensions of 50 Ω air-gapped MCTLs68

Table 3. 2. Comparison of various improved transmission line techniques 74

Table 3. 3. Dimension of optimized microcoax-to-CPW transitions..... 87

Table 4. 1. Chamfering range of the microcoaxial bends 99

Table 4. 2. Optimized chamfering of the microcoaxial bends..... 102

Table 4. 3. Comparison of reported TTD lines and our work..... 117

Table 5. 1. Microcoax cross-sectional dimensions of rat-race hybrid 121

Table 5. 2. Microcoax cross-sectional dimensions of rat-race hybrid 126

Table 5. 3. Dimensions of the designed cavity resonators..... 136

Table 5. 4. Simulated and measured parameters of two cavities 140

Table 5. 5. Dimensions of the designed cavity filters..... 144

Table B. 1. Baking Steps of the SU-8 2025 174

LIST OF FIGURES

Fig. 1. 1. Early transmission lines configurations: (a) Single-wire with implied ground, (b) Two-wire line, (c) Coaxial cable, and (d) Waveguide.	2
Fig. 1. 2. The evolution of the strip line from coaxial line: (a) Coaxial line, (b) Square coaxial line, (c) Rectangular line, and (d) Flat strip line.	4
Fig. 1. 3. Evolution of micro-strip from two-wire line: (a) Two-wire configuration, (b) Single-wire above ground (with image), (c) Micro-strip configuration.	5
Fig. 1. 4. Planar transmission line configurations in MICs: (a) Stripline, (b) Microstrip, (c) Coupled microstrip, (d) Slotline, and (e) Coplanar waveguide.	7
Fig. 1. 5. Attenuation against frequency [61].	15
Fig. 1. 6. Schematic of the micrometre-sized coplanar MIS transmission line with cross-sectional view and top view [61].	15
Fig. 1. 7. Fabrication and integration for the Cu transmission lines and inductors [64].	16
Fig. 1. 8. SEM photograph of air-gap coaxial line on thick oxide barrier layer and Si substrate [64].	16
Fig. 1. 9. SEM photograph of air-gap strip line on thick oxide barrier layer and Si substrate [64].	17
Fig. 1. 10. SEM photograph of air-gap BMSL on thick oxide barrier layer and Si substrate [64].	17
Fig. 1. 11. SEM photograph of air-gap CPW on thick oxide barrier layer and Si substrate [64].	17
Fig. 1. 12. Measured S -parameters of the $50\ \Omega$ air-gap coaxial lines ($w = 30\ \mu\text{m}$, $g = 35\ \mu\text{m}$) [64].	18
Fig. 1. 13. Measured S -parameters of the $50\ \Omega$ air-gap strip lines [64].	18
Fig. 1. 14. Measured S -parameters of the $50\ \Omega$ air-gap CPW ($w = 35\ \mu\text{m}$, $g = 13\ \mu\text{m}$) [64].	18
Fig. 1. 15. Measured S -parameters of the $50\ \Omega$ air-gap BMSL ($w = 30\ \mu\text{m}$, $g = 25\ \mu\text{m}$) [64].	19
Fig. 1. 16. SU-8 based UV-LIGA micromachining process flow: (a) Spin-coating SU-8 onto a Si wafer and soft bake, (b) UV exposure using a dark field mask, (c) Develop in propylene glycol methyl ether acetate (PMGMEA) and gold evaporation and, (d) Evaporate chrome and gold on the micromachined structure, (e) Spin on positive photoresist, pattern and etch the positive photoresist then chrome/gold and, (f) Strip off positive photoresist [65].	20
Fig. 1. 17. Schematic of the micromachined CPW [65].	20
Fig. 1. 18. Measured return loss of the micromachined CPWs with a width of $280\ \mu\text{m}$ and a gap of $70\ \mu\text{m}$ on $7.2\ \Omega\cdot\text{cm}$ [65].	21
Fig. 1. 19. Measured attenuation of four micromachined CPWs [65].	21
Fig. 1. 20. Schematic of microstrips and CPW embedded in interconnection layers on Si [66].	22

Fig. 1. 21. The variations of characteristic impedance ($ z $), loss and slowing factor (SF) of the microstrip with frequency. Both the width of signal line and the distance between signal line and ground are 2 μm [66].	22
Fig. 1. 22. The variations of characteristic impedance ($ z $), loss and SF of the CPW with frequency. The width of signal line, the space between signal line and ground, the distance from the substrate are 5, 20, and 2 μm , respectively [66].	23
Fig. 1. 23. Fabrication process for the TFMS structure: (a) Ground-plane deposition, (b) pattern BCB, (c) bi-layer photoresist process followed by the resist removal [67].	24
Fig. 1. 24. Schematic diagram of the TFMS structures: (a) TFMS with coplanar accesses, (b) TFMS with sidewall isolation [67].	25
Fig. 1. 25. Attenuation of the fabricated 50 Ω TFMS with different dielectric thickness. Continuous lines are the measurement results and dotted lines are the simulation results [67].	25
Fig. 1. 26. Schematic of the shielded passive for Si-based MMICs (a) cross sectional view of microstrip with floating metal shielding, (b) 3D view of the CPW with floating metal shielding [68].	26
Fig. 1. 27. Attenuation of CPW_REF on alumina, shielded CPW and microstrip on Si [68].	26
Fig. 1. 28. Schematic view of CPW transmission line structure with slot-type grounded shields [69].	27
Fig. 1. 29. Measured attenuation of the transmission lines with different slot-type floating shields [69].	27
Fig. 1. 30. Process flow of the monolithic coaxial transmission line: (a) Gold ground plane deposition, (b) Spin and cure bottom dielectric, (c) Gold centre conductor definition, (d) Spin and cure top dielectric, (e) Define coaxial dimensions with Cr/Au etch mask, and (f) Gold plate outer ground conductor [70].	29
Fig. 1. 31. Cross sectional view of a monolithic coaxial transmission line [70].	29
Fig. 1. 32. Attenuation of the coaxial transmission line with centre conductor width (W) of 4 μm [70].	29
Fig. 1. 33. Fabrication steps of the ground-shielded strip line [72].	30
Fig. 1. 34. SEM photograph of the ground-shielded strip line with SU-8 dielectric [72].	30
Fig. 1. 35. Measured attenuation of the ground-shielded strip line with SU-8 dielectric [72].	30
Fig. 1. 36. Structure of the polyimide core micro-coaxial lines (a) 3D schematic, (b) microscope picture [73].	31
Fig. 1. 37. Simulated and measured S-parameters of the micro-coaxial lines [73].	32
Fig. 1. 38. Structure of the air-core micro-coaxial lines with polyimide pedestals (a) 3D schematic, (b) microscope picture [74].	33
Fig. 1. 39. Schematic examples of various 3-D microwave/millimetre-wave transmission lines, which can be realized by using the demonstrated 3-D microconstruction technology: (a) Microstrip line, (b) Suspended coplanar line, (c) Sidewall-shielded microstrip line, (d) bottom-shielded suspended coplanar line, (e)	

Coplanar microstrip line, (f) Sandwiched microstrip line, (g) Semicoaxial line, (h) Microcoaxial line [76].	34
Fig. 1. 40. SEM of (a) Shield CPW and (b) Rectangular coaxial transmission line [76].	35
Fig. 1. 41. Attenuation of the fabricated microcoaxial transmission line [76].	35
Fig. 1. 42. Comparison of the measured attenuation of the EFAB transmission lines with microstrip and CPW [82].	36
Fig. 1. 43. SEM photograph of rectangular coaxial line built by EFAB technology [82]: (a) Detailed rectangular coaxial structure with CPW probe pad, (b) Bended rectangular coaxial line with short-circuited stubs [82].	36
Fig. 1. 44. SEM photographs of a PolyStrata microcoaxial transmission line: (a) Detailed 3D structure of the microcoax [89], (b) 3D pitch antenna arrays with microcoaxial feeding lines [100].	37
Fig. 1. 45. Simulated and measured insertion loss of PolyStrata microcoaxial transmission lines 1 cm (L1) and 3 cm (L3) long.	37
Fig. 2. 1. Schematic of the fabrication process flow: (i) Resist patterning; (ii) Metal electroplating; (iii) Surface planer and CMP planarization; (iv) Repeat (i)-(iii) to form lower side wall; (v) Deposition of lower signal line layer; (vi) Deposition of dielectric straps as the support of signal line; (vii) Deposition of upper signal line layer; (ix) Deposition of upper side wall (xii) Deposition of top layer; (xiii) Strip resist to release the air-gapped microcoaxial structure; (iv') Step (iv) for the portion of the microcoax with lower release holes; (v') Step (v) for the portion of the microcoax with lower release holes; (viii') Step (viii) for the portion of the microcoax with upper release holes.	53
Fig. 2. 2. Schematic of the fabrication process flow which provides the wafer dicing capability: (i) Pattern thin resist; (ii) Etch substrate to form trenches around the location to fabricate the components, and strip resist; (iii)-(vii) Follow the steps of "(i)-(viii)" in Figure 2.2 to build the microcoax structure; (viii) Stick the UV tape on the front surface, and use lapping or Disco surface planer process to polish the backside of the substrate till reach the etched trenches; (ix) Remove the UV tape and thick resist to get the released and diced components.	54
Fig. 2. 3. Fixed-fixed beam with point load model for dielectric straps supporting the signal line of the air-gapped microcoax.	59
Fig. 2. 4. SEM pictures with the (a) Top-view, and (b) 3D-view, of a MCTL at different intermediary stages of fabrication, which were lower side wall, signal line, upper side wall, and top layer respectively, from left to right.	65
Fig. 2. 6. Schematic of the microwave measurement system.	66
Fig. 2. 5. Microwave measurement setup for two-port passive devices.	66
Fig. 3. 1. Schematic of the air-gapped MCTLs: (a) 3D structure; (b) Cross-sectional schematic.	68
Fig. 3. 2. SEM photograph of the fabricated air-gapped MCTLs.	70
Fig. 3. 3. S-parameters of fabricated air-gapped MCTLs 1 cm long.	71
Fig. 3. 4. ϵ_{eff} of different Si-based transmission line technologies.	73
Fig. 3. 5. Attenuation comparison of different transmission line topologies.	73

Fig. 3. 6. 3D schematic of the test structure for measuring the cross-talk between adjacent lines.....	75
Fig. 3. 7. Cross-talk between two adjacent transmission lines: (a) Simulated and measured S_{11} and S_{21} , (b) Simulated and measured S_{31} and S_{41}	75
Fig. 3. 8. Simulated attenuation of the air-gapped MCTL with different strap density.....	76
Fig. 3. 9. Simulated dielectric loss caused by SU-8 with different volume percentage.	76
Fig. 3. 10. Simulated τ of a $50\ \Omega$ microcoaxial line 1 cm long.	79
Fig. 3. 11. Simulation result showing the distribution of electric field for propagating microwaves with 1 W power in the microcoaxial line with cross-sectional dimensions of $300\ \mu\text{m} \times 300\ \mu\text{m}$	80
Fig. 3. 12. Results of FEA simulations showing the distribution of the electric field for propagating microwaves with 1 W EM power in a microcoaxial line with cross-sectional dimensions of: (a) $400\ \mu\text{m} \times 400\ \mu\text{m}$, (b) $500\ \mu\text{m} \times 500\ \mu\text{m}$, (c) $600\ \mu\text{m} \times 600\ \mu\text{m}$, (d) $700\ \mu\text{m} \times 700\ \mu\text{m}$	82
Fig. 3. 13. Power handling capability of the microcoax with different cross-sectional dimensions, based only of breakdown.	82
Fig. 3. 14. Schematic of a proposal microcoaxial transmission line integrated with the advanced Cu BEOL CMOS. The microcoaxial structure can be fabricated on the top ground plate without changing the existing Cu BEOL process.	83
Fig. 3. 15. Design of the microcoaxial-to-CPW transition for the microcoaxial passive components: (a) Schematic of the microcoax-to-CPW transitions' outer structures; (b) Signal part and support structure of microcoax-to-CPW transition design 1; (c) Signal part and support structure of microcoax-to-CPW transition design 2.	86
Fig. 3. 16. Top view of the designed microcoax-to-CPW transitions, SU-8 polymer is not shown: (a) The signal line part; (b) The support structure of design 1; (c) The support structure of design 2.....	87
Fig. 3. 17. Electric field strength on the supporting structure of the microcoax-to-CPW transition at (a) cylinder shaped ground "seat" without metal removal; (b) tube shaped ground "seat" for transition design 1; (c) partially-metal-removed ground "seat" for transition design 2.	89
Fig. 3. 18. SEM photograph of two microcoax-to-CPW transitions connected back-to-back.	89
Fig. 3. 19. Measured S -parameters of the two microcoax-to-CPW transitions.	90
Fig. 3. 20. S -parameter comparison of the microcoax-to-CPW transition design 1 with the reported transitions: (a) Insertion loss comparison; (b) Return loss comparison. ..	91
Fig. 3. 21. Comparison of the probe pads of PolyStrata technology with our microcoax-to-CPW transition: (a) Compare of the simulated S_{11} ; (b) Compare of the simulated S_{21}	92
Fig. 3. 22. Measured Z_0 of the two microcoax-to-CPW transitions.....	92
Fig. 3. 23. Edge clamped circular plate model for the SU-8 film supporting the signal line metal pillar at the CPW end of design 1 transition.	94

Fig. 3. 24. Schematic of the probe pad connected with CPW on Roger PCB/Si substrate.	95
Fig. 3. 25. <i>S</i> -parameters of the probe pad connected with CPW on Roger PCB/Si substrate.	95
Fig. 4. 1. Systematic architecture of the artificial neural network.	98
Fig. 4. 2. Schematic of the chamfered microcoaxial bends: (a) 3D view of the chamfered straight bend; (b) Top view of the chamfered straight bend signal line layer; (c) 3D view of the chamfered curved bend; (d) Top view of the chamfered curved bend signal line layer.	99
Fig. 4. 3. Output of scalable ANN model of Z_0 for: (a) Straight bend; (b) Curved bend.	100
Fig. 4. 4. Simulated <i>S</i> -parameters of single straight bend and single curved bend. ...	102
Fig. 4. 5. SEM picture of meander lines formed by cascaded straight bends and curved bends.	102
Fig. 4. 6. <i>S</i> -parameters of: (a) Meander line formed by 20 straight bends in cascade; (b) Meander line formed by 20 curved bends in cascade.	105
Fig. 4. 7. Characteristic impedance of: (a) Straight bend; (b) Curved bend.	106
Fig. 4. 8. Attenuation constant of: (a) Straight bend; (b) Curved bend.	107
Fig. 4. 9. Phase constant of: (a) Straight bend; (b) Curved bend.	108
Fig. 4. 10. <i>Q</i> -factor of: (a) Straight bend; (b) Curved bend.	109
Fig. 4. 11. V_p and ϵ_{eff} of the straight and curved bends: (a) V_p ; (b) ϵ_{eff}	110
Fig. 4. 12. SEM picture of ultra-long TTD lines formed by: (a) Transmission lines and straight bends in cascade; (b) Transmission lines and curved bends in cascade.	113
Fig. 4. 13. Measured <i>S</i> -parameters of the ultra-long TTD lines: (a) S_{12} & S_{21} are the insertion loss of the ultra-long TTD line formed by straight bends and transmission lines in cascade; S_{34} & S_{43} are the insertion loss of the ultra-long TTD line formed by curved bends and transmission lines in cascade; S_{11} & S_{22} are the return loss of the ultra-long TTD line formed by straight bends and transmission lines in cascade; S_{33} & S_{44} are the return loss of the ultra-long TTD line formed by curved bends and transmission lines in cascade.	114
Fig. 4. 14. Simulated and measured τ of the fabricated TTD lines.	115
Fig. 4. 15. Measured <i>Q</i> -factors of the fabricated TTD lines.	115
Fig. 5. 1. Schematic structure of a conventional rat-race hybrid.	121
Fig. 5. 2. SEM photo of the fabricated miniature microcoaxial rat-race hybrid.	121
Fig. 5. 3. Schematic of the measurement setup for fabricated rat-race hybrid.	122
Fig. 5. 4. Simulated and measured <i>S</i> -parameters of the microcoaxial rat-race hybrid: (a) Symmetrical mode; (b) Differential mode.	124
Fig. 5. 5. Simulated and measured phase difference between two output ports: upper plot is for symmetrical mode; lower plot is for differential mode.	125
Fig. 5. 6. Schematic structures of the conventional 1-to-2 Wilkinson power dividers formed by microstrips: (a) Square configuration; (b) Circular configuration; (c) Triangle configuration.	128

Fig. 5. 7. 3D representation of the prototype and miniature 1-to-2 microcoaxial WPDs.....	129
Fig. 5. 8. SEM photo of the fabricated 1-to-2 microcoaxial Wilkinson power divider.....	130
Fig. 5. 9. Simulated and measured S -parameters of the 1-to-2 WPD without 100 Ω resistors.....	131
Fig. 5. 10. (a) Measured gain error between two paths, and (b) Phase error between two paths, for our fabricated WPD.....	131
Fig. 5. 11. S -parameter comparison of the developed 1-to-2 Wilkinson Power Divider with PolyStrata wideband WPD presented in [92].....	132
Fig. 5. 12. 3D schematic of the air-gapped cavity resonator with weak coupling.....	134
Fig. 5. 13. Fabrication of the 3D microcoax, microcoax-to-CPW transition, and cavity.....	137
Fig. 5. 14. SEM photographs of the fabricated cavity resonators: (a) for 36 GHz applications; (b) For 60 GHz applications.....	137
Fig. 5. 15. Simulated and measured S_{12} & S_{21} of the cavity resonators with f_r value of: (a) 36 GHz; (b) 60 GHz.....	138
Fig. 5. 16. Design of the bandpass filters using two cavity resonators in cascade: (a) 3D representation of the filter design; (b) Top view of the filter design.....	141
Fig. 5. 17. Circuit model of the two coupled identical square cavities: (a) Fully circuit model; (b) Simplified circuit model; (c) Simplified circuit with coupling reactance divided into two parts.....	142
Fig. 5. 18. SEM photographs of the fabricated cavity filters for: (a) 36 GHz applications; (b) 60 GHz applications.....	144
Fig. 5. 19. Simulated and measured S -parameters of the cavity filters for: (a) 36 GHz applications; (b) 60 GHz applications.....	145
Fig. B. 1. SEM photos of photolithographically patterned KMPR resist by lithography process.....	160
Fig. B. 2. Patterned KMPR 1025 film: (a) before Au plating; (b) after Au plating. ..	161
Fig. B. 3. Incomplete strip of KMPR 1025 after multilayer depositions.....	162
Fig. B. 4. Resist thickness from center to edge of a 3" wafer.....	163
Fig. B. 5. SEM photograph of the AZ 4620 for single, double and triple layers deposition.....	164
Fig. B. 6. SEM photograph of plated gold using AZ 4620 as the mold.	165
Fig. B. 7. Microscope photographs for the fabrication steps of the microcoaxial structure using AZ 4620 lithography and Au plating.	167
Fig. B. 8. SEM photographs for the fabricated microcoaxial structures using AZ 4620 lithography and Au plating: (a) Overall view of the microcoaxial structures on Si wafer; (b) Zoom-in view of a microcoaxial structure.	168
Fig. B. 9. Patterned THB 151N film with 50 μm thickness: (a) Microscope photograph, (b) SEM photograph.	168
Fig. B. 10. SEM photograph of Cu structures after CMP.....	168
Fig. B. 11. Microscope photograph of BCB.	171

Fig. B. 12. Microscope photographs of BCB before descum: (a) BCB coated on bare wafer; (b) BCB coated on thick resist and plated metal. 171

Fig. B. 13. Microscope photographs of single layer BCB patterned with: (a) Negative, and (b) Positive patterns, respectively. 172

Fig. B. 14. Microscope photographs of 25 μm thick single layer SU-8 patterned on: (a) A bare Si wafer, (b) A thick resist layer with electroplated Cu. 173

Fig. B. 15. SU-8 2025 on metals after sacrificial resist stripping..... 175

Fig. C. 1. SEM photographs of plated Cu using: (a) Positive patterns, (b) Negative patterns..... 179

Fig. C. 2. SEM photographs of the plated Cu after CMP in: (a) Positive patterns, (b) Negative patterns 180

Fig. C. 3. SEM photographs of the plated Au: (a) With humps and voids, (b) With relative smooth surface. 180

DICTIONARY OF TERMS AND ABBREVIATIONS

Term or Abbreviation	Definition
A	Elongation
ADS	Advanced Design System
Al	Alumina
ANN	Artificial Neural Networks
ASD	Ampere Per Square Decimeter
Au	Gold
BCB	Benzocyclobutene
BEOL	Back End of Line
BMN	Butler Matrix Network
BNC	Bayonet Neill-Concelman
BMSL	Buried Microstrip Line
CBCPW	Conductor Backed Coplanar Waveguide
CMOS	Complementary Metal Oxide Semiconductor
CMP	Chemical Mechanical Polish
CPS	Coplanar Strip
CPW	Coplanar Waveguide
CRLH	Composite Right Left Hand
CTE	Coefficient of Thermal Expansion
Cu	Copper
CVD	Chemical Vapor Deposition

DC	Direct Current
DI water	Deionized Water
DNQ	Diazonaphthoquinone
DUT	Device Under Test
EBR	Edge Bead Remover
EM	Electromagnetic
FOM	Figure of Merit
GaAs	Gallium Arsenide
GaN	Gallium Nitride
GRC	Gold Recovery Carbons
GSG	Ground Signal Ground
HCl	Hydrochloric Acid
HFSS	High Frequency Structure Simulator
HMIC	Hybrid Microwave Integrated Circuit
H2SO4	Sulfuric Acid
IC	Integrated Circuit
InP	Indium Phosphide
IPA	Isopropyl Alcohol
KAIST	Korea Advanced Institute of Science and Technology
LCP	Liquid Crystal Polymer
LIGA	<small>German Acronym</small> Electroplating, Lithography, and Modeling

LSI	Large Scale Integration
LTCC	Low Temperature Co-fired Ceramic
MCTL	Microcoaxial Transmission Line
MEMS	Microelectromechanical Systems
MESD	Multi Exposure Single Development
MIC	Microwave Integrated Circuit
MIS	Metal Insulator Semiconductor
MLP	Multilayer Perceptron
MMIC	Monolithic Microwave Integrated Circuit
MPC	Microwave Printed Circuit
MSI	Medium-Scale Integration
MTDC	MMIC Technology Design Center
Ni	Nickel
NMP	N-Methyl Pyrrolidone
PAC	Photo Active Compound
PCB	Printed Circuit Board
PDS	Plasma Descumming System
PEB	Post Exposure Bake
PEC	Perfect Electric Conductor
PLL	Phase Lock Loop
PNA	Performance Network Analyzer
Pt	Platinum
Q-factor	Quality Factor

RF	Radio Frequency
RIE	Reactive Ion Etch
RVWT	Resistivity Variation with Temperature
SEM	Scanning Electron Microscope
Si	Silicon
SIP	System in Package
SOC	System on Chip
SOG	Spin on Glass
SOLT	Short-Open-Load-Through
SSI	Small-Scale Integration
TE	Transverse Electric
Teflon	TFE Fluorocarbon
TEM	Transverse Electromagnetic
TFMS	Thin Film Microstrip
Ti	Titanium
TM	Transverse Magnetic
TNC	Threaded Neill-Concelman
TTD	True Time Delay
UHF	Ultra-High Frequency
ULSI	Ultra-Large-Scale Integration
UV	Ultraviolet
VHF	Very-High Frequency
VLSI	Very-Large-Scale Integration

VNA	Vector Network Analyzer
W	Tungsten
WPD	Wilkinson Power Divider
WSI	Wafer-Scale Integration
c	Speed of Light
f	Wave Frequency
R_s	Surface Resistivity
α	Attenuation Constant
β	Phase constant
γ	Propagation Constant
δ_s	Skin Depth
λ	Wavelength
ρ	Resistivity
σ	Conductivity
τ	Group Delay
ϵ_{eff}	Effective Permittivity
ϵ_r	Relative Permittivity

1. Introduction

1.1. Overview of the Chapter

Any medium that confines and transmits electromagnetic energy from one point to another can be defined as a transmission line. Transmission lines are the fundamental elements of the microwave circuits and can also be used together with other passive devices. The evolution of the transmission lines and their associated technologies has been going on for more than 100 years. The semi-insulating III-V compound semiconductor substrates are the most popular substrates for the current microwave applications, due to the high power and high cutoff frequency of III-V active devices made on them, as well as the low substrate loss for the passive components. A brief history of transmission lines is introduced at the beginning of this chapter. The literature of the advanced transmission line techniques for Si-based microwave applications is reviewed in the latter part of this chapter. The organization of the thesis is summarized at the last part of this chapter.

1.2. History of Transmission Lines

The first batch of transmission lines can be traced back to the electricity power distribution lines and electric telegraph lines in the 19th century. During 1880 to 1900, Hertz built first batch of radio frequency (RF) transmission lines for his experimental setup which verified Maxwell's electromagnetic theory, including single-wire line with implied ground, two-wire line, and coaxial cable [1], although their configurations are very primitive. The waveguide structure was mathematically approved to be able to propagate microwaves by Rayleigh in 1897. The configurations of the early transmission lines are shown in Figure 1.1.

Fig. 1. 1. Early transmission lines configurations: (a) Single-wire with implied ground, (b) Two-wire line, (c) Coaxial cable, and (d) Waveguide.

From the 20th century, the growing need for telephone, radar, and television systems drove the development of transmission lines supporting electromagnetic signals with higher frequencies. In 1930's, fully-shielded coaxial line with sizes from 0.5 inch to 2.4 inch were widely used for telephone and telegraph applications. The air-dielectric coaxial lines with bead supports between inner conductor and outer conductor were also developed for a low dielectric loss. The bead supports were formed by different materials including rubber, porcelain, and ceramics, also by different configurations such as webbed and star configurations. Air occupies more than 99% of total space volume between the signal and ground conductors. The air-dielectric coaxial lines were further developed and improved for the applications in the transmitters for radio broadcasting as well as the ultra-high frequency (UHF) radars.

The heavy demand of the radar systems during the WW II period also boosted the research on transmission lines. Two-wire line was used in radar at frequencies of 200 MHz and below throughout that period. Coaxial lines were continued used for the S-

band and X-band radar systems. Components such as filters, resonators, directional couplers, attenuators, detectors, and mixers were also synchronized by the coaxial lines. Meanwhile, flexible coaxial lines were invented to provide flexing capability as all the previous coaxial lines had rigid structures. Polyethylene developed in 1940's as the dielectric material was one important breakthrough for the development of flexi coaxial lines. The polyethylene has a dielectric constant of 2.25 and loss tangent of 0.0004 hence is suitable for developing low loss coaxial lines. Waveguides with hollow tube structure provide advantages such as ultra-high power handling capability, low attenuation, compatibility with other RF components and test equipment, and ease of fabrication. Hence they were more commonly used than coaxial lines during WW II [1].

The RF and microwave system requirements change after WW II. Circuits and systems developed towards smaller size, wider bandwidth, lower power consumption, greater functional complexity. Following this trend, coaxial transmission line and two-conductor transmission line with new configurations were widely used. TFE fluorocarbon (Teflon) support bead or support stub was used in air-dielectric coaxial lines, especially for radio and television broadcasting, as well as high power very-high frequency (VHF) and UHF communications.

The early microwave circuits and systems developed were heavy and bulky until the advent of microwave printed circuits (MPC) in 1950s, which was an important milestone for the microwave industry. The embryo of the MPC was the planar transmission line circuit developed during WW II [2]. The planar transmission lines are evolved results from conventional transmission lines in progressive steps: The evolution of the strip line from coaxial cables is shown in Figure 1.2; the evolution of

the microstrip line from the two-wire configuration transmission cables is shown in Figure 1.3.

Fig. 1. 2. The evolution of the strip line from coaxial line: (a) Coaxial line, (b) Square coaxial line, (c) Rectangular line, and (d) Flat strip line.

The evolution of the transmission lines to planar geometry enabled them to adapt to the conventional printed-circuit technology. Besides the stripline and microstrip lines, other planar transmission line configurations, such as coplanar waveguide (CPW) and slot lines, were also used in the MPCs. The planar transmission line technology can be used to provide most of the passive components on the MPCs, such as power splitters, filters, antennas, hybrid junctions, directional couplers, terminations and attenuators, etc. The MPC provides a platform for the design and fabrication of simple, fast, and inexpensive RF and microwave circuits. Therefore, the planar transmission lines became the more popular choice compared with the large, three-dimensional waveguides and coaxial

lines in traditional, bulky, microwave circuits, except for certain high-power applications.

Fig. 1. 3. Evolution of micro-strip from two-wire line: (a) Two-wire configuration, (b) Single-wire above ground (with image), (c) Micro-strip configuration.

In 1960s, the growing market drove the development of the transmission line technologies. Before the launch of the commercial integrated circuits (IC) on the market, the RF systems were all built by a conventional component assembly method. In this assembly method, discrete passive devices formed by coaxial lines, waveguides, and strip lines were packaged as individual stand-alone components. Then these individual components were assembled by coaxial connectors such as Bayonet Neill–Concelman

(BNC) connectors and Threaded Neill-Concelman (TNC) connectors to form RF systems. The active dies were mounted and bonded to the passive circuits in individual packages. The conventional assembly method provided good solutions for many RF applications. The electrical, thermal, and mechanical problems commonly roused by the solid-state devices were also solved by this method. However, the package assembly method encountered many problems for high frequency applications. The size of the entire packaged system was usually much bigger than the active dies inside. The discontinuities between the devices also limited the performance. Moreover, the reliability and mechanical integrity issues also limited its usage for space and military applications.

The launch of the first IC on the market inaugurated the era of modern electronics. Problems such as interface parasitic reactance, high cost, and big size can be solved by utilizing the IC technologies to the MPC. The first ICs operated only at low frequencies due to drastic limitations imposed by the quality and performance of the incorporated devices as well as the significant parasitic components inherent to the system. However, higher frequencies had always been an important objective from the beginning of communication industry. The frequency increase provides smaller beam width and higher gain for a given size of antenna. A smaller beam width also increases angular measurements accuracy, and a higher gain improves the detection sensitivity or range for the same transmitter power and receiver sensitivity. The frequency increase also enables the reduction of equipment size for a given system function, which is very important for portable and airborne communication systems. The frequency increase of ICs utilized the microwave integrated circuits (MICs). Two forms of MICs were developed, namely the hybrid microwave integrated circuits (HMICs) and the monolithic microwave integrated circuits (MMICs).

In HMICs, the transmission lines on non-semiconductor substrates provide a platform to connect the discrete active and passive components to form the circuits and system [3]. As shown in Figure 1.4, transmission lines in HMICs are mostly in configurations of stripline, microstrip [4], slotline, CPW [5], finline, and many other types of related geometries. These

Fig. 1. 4. Planar transmission line configurations in MICs: (a) Stripline, (b) Microstrip, (c) Coupled microstrip, (d) Slotline, and (e) Coplanar waveguide.

transmission lines provide easy integration with the semiconductor active devices. The sealed enclosures are required for the HMICs thus drive a rapid growth of the impedance matched hermetic-seal connectors in many different configurations.

With the development of the IC technology, all the parts including the active devices and passive circuits can be completely integrated on single semiconductor chips. This development of IC technology played an extremely important role in MICs development, giving birth to the MMICs. The MMICs provide much greater integration and much smaller sizes than HMICs, as well as the capability for mass production. Hence the MMICs become preferred choice instead of the HMIC for any applications that require high volume [6].

The research on MMICs has started in the 1960s. First, silicon (Si) substrates were tried in MMIC experiments, but they provided poor performance. This is mainly due to the fact that the Si substrate cannot provide enough power for low voltage active devices, as well as having a high substrate loss for transmission lines and other passive devices. In the 1970s, the first commercial MMIC was built on a GaAs substrate, which enables the fabrication of high power and high frequency active devices [7]. Additionally, the low loss tangent of GaAs provides a perfect platform for transmission lines and passive circuits. Other compound semiconductors, such as GaN, InP, are also used for further development of the MMICs. The MMICs provide greater integration than HMIC as all the active devices and passive devices are fabricated on the same chip. The MMICs have many performance advantages such as low loss, low noise, large bandwidth, high power, smaller volume and weight, all of which are very important for both defense and commercial applications. However, transmission lines and their synchronized passive circuits occupy more than 95% of the substrate area. Hence the MMIC doesn't offer any cost advantage, and it is difficult to obtain high packing density for MMICs.

In the recent few decades, there is a large expansion of consumer-based wireless communication systems, including cellular phones, direct satellite communications and high data-rate secure digital networking. These systems demand faster transceiver speed, higher variety of carrier frequencies, more functions and lower cost in lighter and smaller formats. There is a strong trend of integrating the RF and microwave systems to realize system-on-chip (SOC) or system-in-package (SIP). Singular discrete and surface-mounted passive components tend to be replaced by more integrated solutions. The RF and microwave systems are expanding by miniaturizing electronics components and increasing the total number of electronic devices. Planar transmission line

technologies are used widely in fabrication of passive components for the integrated RF and microwave system.

However, cost reduction has recently reached a plateau indicating that the conventional planar technology has reached its limitations, such as low power handling capabilities, inability to support the pure transverse-electromagnetic (TEM) mode, unwanted radiation and dispersion, high loss, poor field confinement, and low signal isolation. In order to address the needs of the next generation of communication systems and microwave or millimeter wave radars, micromachining techniques utilized by microelectromechanical systems (MEMS) are studied to provide three-dimensional (3D) transmission line geometries to reduce or eliminate these limitations of planar transmission lines. These 3D transmission lines will constitute the next generation of wireless communication systems.

The above description briefly introduced the development history of the transmission lines, or rather, of the microwave circuits and systems. The transmission lines are the fundamental components of, and have evolved together with the microwave circuits and systems. Furthermore, many important passive devices can be connected using transmission lines, including distributed filters, the branch-line and rat-race hybrids, power combiners, etc. [8]. Lumped components, such as spiral inductors and parallel-plate capacitors, can usually be fabricated with the same technology as the transmission lines. The comparison between different transmission line topologies is listed in Table 1.1. These topologies were invented in different historical periods and because they have gone through a continuous development, their modern forms still play important roles in nowadays microwave systems. The major difference is that some topologies are used more for chip level applications while others may be more used for

system level applications. The coaxial transmission line is conventionally used for system level applications due to the high difficulty of fabrication and integration. However, the coaxial topology provides absolute performance advantages such as ultra-low loss, ultra-low crosstalk, real TEM mode, and high power handling capability.

In practice, waveguides act as the equivalent of cables for super high frequency (SHF) systems. For these applications, it is desired to operate waveguides with only one propagation mode. Practical waveguides usually have rectangle cross-section with length to width ratio of 2:1. Assume the lower cutoff frequency (f_c) of the lowest order mode is f_{c1} . Then the recommend operation frequency band ranges $1.25f_{c1} \sim 1.75f_{c1}$. And the cutoff frequency of the next order mode is about $2f_{c1}$. The standard waveguides have inner dimensions ranges from $23 \times 11.5 \text{ inch}^2$ (WR2300) to $0.034 \times 0.017 \text{ inch}^2$ (WR3), which is corresponding to the operation band from 0.32~0.45 GHz to 220~330 GHz. Coaxial cables are widely used to connect RF components. Their practical operation frequencies for the RF circuits and systems are below 3 GHz. The losses of the coaxial cables for long distance transmission will be too excessive for frequencies above 3 GHz. A higher operation frequency would also cause the wave propagation in higher order modes. However, microcoaxial lines developed could be used with operation frequencies up to hundreds of GHz in applications of MICs, which are formed by miniaturized components and short-distance connection.

Table 1. 1. Comparison of common transmission lines

Type of transmi -ssion line Characteristic	Waveguide	Coaxial	Strip line	Microstrip	CPW	Slot line
Modes: Preferred Other	TE ₁₀ TM, TE	TEM TM, TE	TEM TM, TE	Quasi-TEM Hybrid TM, TE	Quasi-TEM Hybrid TM, TE	Hybrid TM, TE
Dispersion	Medium	None	None	Low	Low	Low
Bandwidth (GHz)	Low	High	High	High	High	High
Loss	Low	Medium	High	High	High	High
Power capacity	High	Medium	Low	Low	Low	Low
Ease of fabrication	Medium	Medium	Easy	Easy	Easy	Easy
Integration	Hard	Hard	Fair	Easy	Easy	Easy

1.3. Transmission Lines for Si-based MICs

The previous section described the history of the transmission line and how it was constantly innovated and modified for different generations of microwave circuits. The 3D topologies have been more often used for system-level microwave applications while the planar topologies are preferred for chip-level microwave applications. There is still high demand and big room for improving the chip-level microwave circuits such as the MICs, including both HMICs and MMICs. The conventional MICs are using the planar transmission line topologies such as microstrips and CPWs. However, microstrips and CPWs have the drawbacks of dispersion [9-11], high attenuation [10, 12], radiation loss and cross-talk between devices.

To provide low loss HMICs for microwave applications, Si bulk micromachining and various different substrates were applied to develop new topologies for passive components. Previously reported works include microstrip and CPW on liquid crystal polymer (LCP) [13], microstrip and conductor-backed CPW (CBCPW) on metallized glass substrate coated with SU-8 [14, 15], cavity resonator formed by metallized SU-8 on glass substrate [16], cavity resonator on low-temperature co-fired ceramic (LTCC) [17-19], passive components fabricated by bulk micromachining on the Si substrates [20-27], cavities formed by bulk micromachining and wafer bonding of the Si substrates [28-36], shielded and membrane-supported planar passives fabricated by bulk micromachining and wafer bonding of the Si substrates [37-41], rectangular microcoax cable assembled by microfabrication, peel off Si substrate and, bonding [42-45], membrane supported CPW mounted with active devices for integration [46], various inductors using CMOS-compatible technologies [47-60], and monolithic half-coaxial structure using surface micromachining process on a quartz substrate [61, 62].

As discussed in the previous section, the commercial MMIC circuits are dominated by III-V compound semiconductor substrates, although with extremely high cost. Si is a low cost, widely used and well developed substrate material in IC industry. However, the development of the Si-based MMICs was suspended from the early stage due to the low power density for active devices and high loss for the passive devices. Providing high performance active and passive devices, Si would still be the most desirable material for MMICs industry to achieve a tremendous cost reduction.

Since the first commercialization, silicon complementary metal-oxide-semiconductor (CMOS) has evolved rapidly and attracted billions of dollars for further development. The CMOS ICs have progressed generations of processes from small-scale integration (SSI), medium-scale integration (MSI), large-scale integration (LSI), very-large-scale integration (VLSI), and ultra-large-scale integration (ULSI). Recently, wafer-scale integration (WSI), SOC, and 3D ICs are under development for the new generations of CMOS technology. The modern CMOS technology has improved the power performance of active devices and pushed their cut-off frequency to hundreds of GHz. The development of Si-based MICs is only retarded by the absence of high-quality and area-efficient on-chip passive components.

Extensive studies were conducted to develop low-loss passive components on Si based on various approaches that were proposed in the past few decades. One approach is to modify the conventional planar passives to reduce the substrate loss by either adding isolative membranes on the substrate or elevating the whole passive structure [63-69]. Another approach is using the rectangular coaxial structure with ground shielding to confine the electromagnetic wave. Devices of this category include a polyimide-core 3D rectangular coaxial structure fabricated using a modified thin layer process [70-73],

microcoaxial air-lines realized using a multi-layer sacrificial resist process [74], a copper (Cu) coaxial structure implemented using a multi-exposure-single-development (MESD) lithography process [75-77], a rectangular coaxial transmission line made using a Cu-nickel (Ni) EFAB process with more than 12 layers [78-83], and a rectangular coaxial transmission line manufactured with PolyStrata technology by Nuvotronics, Inc [84-102]. This section will give a detailed review of the published work for developing advanced transmission lines for Si-based MMICs.

1.3.1. Isolate the planar passives from the substrate

One approach is to modify the conventional planar passives to reduce the substrate loss, by either adding isolative membranes on the substrate or elevating the whole passive structure [63-69].

Vincent M. Hietala *et al* developed coplanar metal-insulator-semiconductor (MIS) transmission lines on a heavily doped N^+ silicon substrate [63], as shown in Figure 1.5. Figure 1.6 shows the attenuation of this MIS transmission line. The thin layer of insulator can effectively block the direct current (DC) from the transmission line conductor to the substrate. However, the RF current can still be transmitted through the capacitor formed by the conductor, insulator layer, and the Si substrate, especially at high frequencies. Hence, the coplanar MIS transmission line shows a relatively high attenuation of less than 1 dB/mm in the frequency range 1 to 10 GHz.

Fig. 1. 6. Schematic of the micrometre-sized coplanar MIS transmission line with cross-sectional view and top view [61].

Fig. 1. 5. Attenuation against frequency [61].

Inho Jeong *et al* developed air-gap transmission lines on a 1- μm -thick thermal oxide barrier layer or a 50- μm -thick spin-coated resist barrier layer, which separate the transmission lines from the Si substrate to reduce the substrate loss [64, 75, 103]. Figure 1.7 shows the fabrication process for the air-gap transmission lines. Figure 1.8 to Figure 1.11 show the scanning electron microscope (SEM) photographs of air-gap Cu coaxial line, stripline, CPW, and buried microstrip line (BMSL), respectively, fabricated using the multilayer process developed by Inho Jeong *et al*, while Figure 1.12 to Figure 1.15 show their measured *S*-parameters. The attenuations are less than 0.08 dB/mm, 0.15

dB/mm, 0.07 dB/mm and 0.13 dB/mm for the fabricated air-gap coaxial line, stripline, CPW, and BMSL, respectively, in the frequency range 0.1 to 40 GHz.

Fig. 1. 7. Fabrication and integration for the Cu transmission lines and inductors [64].

Fig. 1. 8. SEM photograph of air-gap coaxial line on thick oxide barrier layer and Si substrate [64].

Fig. 1. 9. SEM photograph of air-gap strip line on thick oxide barrier layer and Si substrate [64].

Fig. 1. 10. SEM photograph of air-gap BMSL on thick oxide barrier layer and Si substrate [64].

Fig. 1. 11. SEM photograph of air-gap CPW on thick oxide barrier layer and Si substrate [64].

Fig. 1. 12. Measured S -parameters of the $50\ \Omega$ air-gap coaxial lines ($w = 30\ \mu\text{m}$, $g = 35\ \mu\text{m}$) [64].

Fig. 1. 13. Measured S -parameters of the $50\ \Omega$ air-gap strip lines [64].

Fig. 1. 14. Measured S -parameters of the $50\ \Omega$ air-gap CPW ($w = 35\ \mu\text{m}$, $g = 13\ \mu\text{m}$) [64].

Fig. 1. 15. Measured S -parameters of the $50\ \Omega$ air-gap BMSL ($w = 30\ \mu\text{m}$, $g = 25\ \mu\text{m}$) [64].

Daniel P. Newlin *et al* developed a micromachined CPW on $200\text{-}\mu\text{m}$ -thick spin-coated SU-8 barrier layers, which separates the transmission lines from the Si substrate to reduce the substrate loss, by using the LIGA (a German acronym for electroplating, lithography, and molding) micromachining process [65]. Figure 1.16 shows the fabrication process for the SU-8 based ultraviolet (UV)-LIGA micromachining. Figure 1.17 shows schematically the micromachined CPW. Table 1.2 shows the dimension of the fabricated micromachined CPW on Si with SU-8 barrier layer. Figure 1.18 and Figure 1.19 show the measured return loss and attenuation of the fabricated micromachined CPW. The conductor-backed micromachined CPW on Si ($7.2\ \Omega\cdot\text{cm}$) achieved a measured return loss less than $-20\ \text{dB}$ and attenuation less than $0.18\ \text{dB/cm}$ in the frequency ranges $0.1\ \text{GHz}$ to $20\ \text{GHz}$.

Fig. 1. 16. SU-8 based UV-LIGA micromachining process flow: (a) Spin-coating SU-8 onto a Si wafer and soft bake, (b) UV exposure using a dark field mask, (c) Develop in propylene glycol methyl ether acetate (PMGMEA) and gold evaporation and, (d) Evaporate chrome and gold on the micromachined structure, (e) Spin on positive photoresist, pattern and etch the positive photoresist then chrome/gold and, (f) Strip off positive photoresist [65].

Fig. 1. 17. Schematic of the micromachined CPW [65].

Table 1. 2. Dimensions of micromachined CPWs [65]

	Signal Width W (μm)	Gap S (μm)	Si Thickness (μm)	SU-8 Thickness (μm)
SU-8/Si CPW1 SU-8 CPW1	280	70	400	200
SU-8/Si CPW2 SU-8 CPW2	90	30	400	200

Fig. 1. 18. Measured return loss of the micromachined CPWs with a width of 280 μm and a gap of 70 μm on 7.2 $\Omega\cdot\text{cm}$ [65].

Fig. 1. 19. Measured attenuation of four micromachined CPWs [65].

Guo Lihui *et al* developed CMOS-compatible microstrip and CPW on the Si substrate with 2- μm -thick thermally-grown SiO_2 barrier layer and 2- μm -thick Cu as conductor material using state-of-the-art back-end-of-line (BEOL) technology [66]. Figure 1.20 shows the schematic of the microstrips and CPW embedded in interconnection layers on Si substrate, while Figure 1.21 and Figure 1.22 show their measured attenuation. It can be seen that the attenuation of the microstrips had values less than 1 dB/mm in the frequency ranges of 0.1 GHz to 30 GHz. And the CPW embedded in interconnection

layers on Si substrate has a measured attenuation less than 0.7 dB/mm in the frequency ranges of 0.1 GHz to 40 GHz.

Fig. 1. 20. Schematic of microstrips and CPW embedded in interconnection layers on Si [66].

Fig. 1. 21. The variations of characteristic impedance ($|Z|$), loss and slowing factor (SF) of the microstrip with frequency. Both the width of signal line and the distance between signal line and ground are 2 μm [66].

Fig. 1.22. The variations of characteristic impedance ($|Z|$), loss and SF of the CPW with frequency. The width of signal line, the space between signal line and ground, the distance from the substrate are 5, 20, and 2 μm , respectively [66].

Gonzague Six *et al* developed the thin-film microstrip (TFMS) on the thick benzocyclobutene (BCB) layer, which separates the TFMS from the silicon substrate to reduce the substrate loss [67]. Figure 1.23 shows the fabrication process of the TFMS. Figure 1.24 shows the schematic of the fabricated TFMS structure and Figure 1.25 shows its measured attenuation vs. frequency. The fabricated TFMS obtained very low loss, regardless of the substrate type. As shown in Figure 1.25, the attenuation is less than 0.6 dB/mm, 0.3 dB/mm, and 0.2 dB/mm for TFMS with polymer film of 6 μm , 10 μm , and 20 μm in the frequency range 0.1 to 50 GHz.

Fig. 1. 23. Fabrication process for the TFMS structure: (a) Ground-plane deposition, (b) pattern BCB, (c) bi-layer photoresist process followed by the resist removal [67].

Fig. 1. 24. Schematic diagram of the TFMS structures: (a) TFMS with coplanar accesses, (b) TFMS with sidewall isolation [67].

Fig. 1. 25. Attenuation of the fabricated 50 Ω TFMS with different dielectric thickness. Continuous lines are the measurement results and dotted lines are the simulation results [67].

Tak Shun Dickson Cheung *et al* developed shielded passive devices for Si-based MMICs using CMOS or BiCMOS technologies [68]. The shielded passive devices use floating metal lines (typically Al and Cu) to shield the passives in order to avoid the loss from the Si substrate. Figure 1.26 shows schematically the microstrip and CPW with floating shield for Si-based MMICs. As shown in Figure 1.27, the attenuation is less than 0.4 dB/mm, 0.5 dB/mm, 0.6 dB/mm, and 0.7 dB/mm for the fabricated shielded microstrip 2, microstrip 1, CPW 1, and CPW 2 in the frequency range of 0.1 to 40 GHz.

Fig. 1. 26. Schematic of the shielded passive for Si-based MMICs (a) cross sectional view of microstrip with floating metal shielding, (b) 3D view of the CPW with floating metal shielding [68].

Fig. 1. 27. Attenuation of CPW_REF on alumina, shielded CPW and microstrip on Si [68].

Cho Hsiu Ying *et al* developed a transmission line with optimized slot-type floating shields [69], which is very similar to the shielded passive devices developed by Tak Shun Dickson Cheung *et al*. Figure 1.28 shows the schematic of a CPW with slot-type grounded shields, while Figure 1.29 shows the measured attenuations of the transmission lines with different slot-type floating shields.

Fig. 1. 28. Schematic view of CPW transmission line structure with slot-type grounded shields [69].

Fig. 1. 29. Measured attenuation of the transmission lines with different slot-type floating shields [69].

1.3.2. Polyimide-core rectangular coaxial structures

Besides reducing the substrate loss, many research groups proposed the approach, namely fabricating rectangular coaxial structures by deposition of metal conductors inside and around the polyimide core [70-73].

In 1991, J. A. Bishop proposed a monolithic coaxial transmission line structure fully shielded by a gold-plated ground plane [70], whose fabrication process flow is shown in Figure 1.30. The coaxial transmission line uses Au evaporation and lift-off process for the metallization and uses probimide as the dielectric. Figure 1.31 shows the cross-sectional schematic of the monolithic coaxial transmission line. The attenuation of the transmission line is about 1.6 dB/mm at 40 GHz, as shown in Figure 1.32.

Fig. 1. 30. Process flow of the monolithic coaxial transmission line: (a) Gold ground plane deposition, (b) Spin and cure bottom dielectric, (c) Gold centre conductor definition, (d) Spin and cure top dielectric, (e) Define coaxial dimensions with Cr/Au etch mask, and (f) Gold plate outer ground conductor [70].

Fig. 1. 31. Cross sectional view of a monolithic coaxial transmission line [70].

Fig. 1. 32. Attenuation of the coaxial transmission line with centre conductor width (W) of $4\text{ }\mu\text{m}$ [70].

In 2003, Ricardo Osorio developed a 3D ground-shielded strip line using MEMS technologies [71, 72]. Figure 1.33 shows the fabrication process flow of the 3D ground shielded strip line. Figure 1.34 shows the SEM picture of the fabricated 3D ground-shielded strip line. The dielectric material of the strip lines is formed by photosensitive polymer SU-8, which has a high loss tangent of 0.043. This causes the strip lines to have high attenuations up to 0.8 dB/mm at 50 GHz, as shown in Figure 1.35.

Fig. 1. 33. Fabrication steps of the ground-shielded strip line [72].

Fig. 1. 34. SEM photograph of the ground-shielded strip line with SU-8 dielectric [72].

Fig. 1. 35. Measured attenuation of the ground-shielded strip line with SU-8 dielectric [72].

In 2010, Saravana P. Natarajan *et al* presented a rectangular polyimide core micro coaxial transmission line [73], whose structure is shown in Figure 1.36. Figure 1.37 shows its simulated and measured *S*-parameters.

Fig. 1. 36. Structure of the polyimide core micro-coaxial lines (a) 3D schematic, (b) microscope picture [73].

Fig. 1.37. Simulated and measured S -parameters of the micro-coaxial lines [73].

1.3.3. Air-gapped rectangular coaxial structures

The polyimide-core rectangular coaxial line is fabricated using a simple surface micromachining process. However, passives fabricated by this approach do not offer a significant on device attenuation due to the dielectric loss of the polyimide core. Many researches also focused on developing air-gapped rectangular coaxial structures [74-82, 84-102].

Besides the polyimide core micro-coaxial transmission line, Saravana P. Natarajan *et al* also developed an air core micro-coaxial transmission line partially filled with polyimide as the pedestals used to support the signal line [74]. The structure of the developed air core microcoaxial transmission line is shown in Figure 1.38.

Fig. 1. 38. Structure of the air-core micro-coaxial lines with polyimide pedestals (a) 3D schematic, (b) microscope picture [74].

In 2002 and 2003, Korea advanced institute of science and technology (KAIST) proposed various CMOS-compatible thick-metal surface micromachining technologies which can be used to build 3D passive components [75-77]. The fabrication process of a thick-metal surface micromachining technology is shown in Figure 1.39. The SEM pictures of the air-suspended CPW and rectangular coaxial transmission lines fabricated by this 3D micromachining technology are in shown Figure 1.40. The measured attenuation of the fabricated rectangular coaxial transmission lines on both Si and glass substrates are plotted in Figure 1.41. The air suspended structures eliminate the dielectric loss of conventional planar transmission lines. The measured attenuation of the shielded CPW is 0.035 dB/mm at 25 GHz and the measured attenuation of the

rectangular coaxial transmission line is 0.24 dB/mm at 10 GHz. The transmission line's test pads suffer high substrate loss due to the intimate contact with the lossy Si substrate.

Fig. 1. 39. Schematic examples of various 3-D microwave/millimetre-wave transmission lines, which can be realized by using the demonstrated 3-D microconstruction technology: (a) Microstrip line, (b) Suspended coplanar line, (c) Sidewall-shielded microstrip line, (d) bottom-shielded suspended coplanar line, (e) Coplanar microstrip line, (f) Sandwiched microstrip line, (g) Semicoaxial line, (h) Microcoaxial line [76].

Fig. 1. 40. SEM of (a) Shield CPW and (b) Rectangular coaxial transmission line [76].

Fig. 1. 41. Attenuation of the fabricated microcoaxial transmission line [76].

The commercial EFAB foundry provides a solution for the fabrication of miniature rectangular coaxial transmission lines [78-83], as shown in Figure 1.42, which have the advantage of high isolation, small turning radius, and lossless air-dielectric. Figure 1.43 compares the attenuation of the EFAB transmission lines with those of the conventional microstrip and CPW. However, the fabrication process of this EFAB technology is too complex considering the 41 layers of Cu-Ni depositions, polishing, and etching. And Ni is a magnetic metal which has permeability dispersion along frequency, which may result signal distortion and is not desired for wideband applications. Another drawback of the EFAB approach is the lack of dielectric support for the center conductor. The

Fig. 1. 43. SEM photograph of rectangular coaxial line built by EFAB technology [82]: (a) Detailed rectangular coaxial structure with CPW probe pad, (b) Bended rectangular coaxial line with short-circuited stubs [82].

Fig. 1. 42. Comparison of the measured attenuation of the EFAB transmission lines with microstrip and CPW [82].

center conductor is supported by the short-circuited quarter wavelength stubs thus limiting the transmission lines with narrow bandwidths.

From 2004 to 2010, miniature, dense and very complex 3D metallic-dielectric components were demonstrated using the so-called PolyStrata technology developed by Nuvotronics and the University of Colorado, Boulder [84-102]. The shielded, air-suspended rectangular coaxial transmission lines and patch antennas are shown in Figure

1.44. The rectangular coaxial transmission line supports the TEM propagation mode up to hundreds of GHz. Figure 1.45 shows both the simulated and measured losses of the fabricated microcoaxial transmission lines 1 cm and 3 cm long. The attenuation of the transmission line is below 0.01 dB/mm up to 40 GHz.

Fig. 1. 44. SEM photographs of a PolyStrata microcoaxial transmission line: (a) Detailed 3D structure of the microcoax [89], (b) 3D pitch antenna arrays with microcoaxial feeding lines [100].

Fig. 1. 45. Simulated and measured insertion loss of PolyStrata microcoaxial transmission lines 1 cm (L1) and 3 cm (L3) long.

1.4. Organization of the Thesis

The previous section reviewed the literature to present the state-of-the-art for Si-based MICs. From this literature review, it can be concluded that three main approaches were used for developing low loss passive components on Si substrates.

The 1st approach added insulating layers on the substrate or elevated the whole passive structure to reduce the substrate loss, in a certain degree. The 2nd approach used a polyimide-core 3D rectangular coaxial structure that eliminated the substrate loss by confining the electromagnetic field within the ground shield. However, the polyimide core also introduces a non-negligible dielectric loss for the coaxial structure. Finally, in the 3rd approach, the air-gapped microcoaxial structure is obviously a more implementable approach due to the advantages it offers in performance, such as low loss, low cross-talk, and real TEM propagation mode. The major limitations of this approach include high cost and complexity of fabrication, and the challenge of device integration with the planar circuit platform, which dominates the current circuits and systems.

In this thesis, the air-gapped microcoaxial transmission line (MCTL) technology for Si-based MICs is demonstrated. The design, fabrication, and measurement for the air-gapped MCTL technology are presented in Chapter 2. The fabrication process of transmission lines using the developed MCTL technology is presented in Chapter 3. The integration of the fabricated microcoaxial components is also discussed. A novel Microcoax-to-CPW transition for device measurement and integration is also demonstrated in this chapter. The microcoaxial bends for miniaturization of the microcoaxial passive devices are demonstrated in Chapter 4. The true-time-delay lines developed by combining the coaxial transmission lines and microcoaxial bends are also

presented in this chapter. Chapter 5 presents the developed microcoaxial rat-race hybrids and other passive components including microcoaxial Wilkinson power divider, cavity resonator, and cavity filters. Finally, Chapter 6 gives the conclusions and discussions the results of the thesis, as well as introduces the future work.

2. Development of Air-gapped MCTL Technology

2.1. Overview of the Chapter

The previous chapter introduced the history of transmission lines and various transmission line technologies for Si-based microwave ICs. This chapter describes the detailed procedures for the MCTL technology development, including microwave modeling and simulation, fabrication process development, and device measurement. The modeling and simulation of the MCTLs is introduced at the beginning of this chapter. The fabrication process development is discussed at the later part of this chapter. The measurement setup is briefly introduced at the last part of this chapter.

2.2. Transmission Line Modeling and Wave Propagation

Modes

2.2.1. Electromagnetic-wave propagation phenomena

In order to study the transmission line and other microwave passive devices, the electromagnetic (EM) wave propagation phenomena has to be understood. The microwave fundamental theory starts from the Maxwell Equations:

$$\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t}$$

$$\text{(Faraday's Law of Induction)} \quad (2.1),$$

$$\nabla \times \vec{H} = \frac{\partial \vec{D}}{\partial t} + \vec{J}$$

$$\text{(Ampere's Circular Law)} \quad (2.2),$$

$$\nabla \cdot \vec{D} = \rho_v \quad \text{(Gauss's Law)} \quad (2.3),$$

$$\nabla \cdot \vec{B} = 0 \quad \text{(Gauss's Law of Magnetism)} \quad (2.4),$$

where E is the electric field intensity, H is the magnetic field intensity, D is the flux density or electric displacement vector, B is the magnetic flux density vector, J is the current density vector which equals σE , and ρ_v is the volume charge density. In a single uniform, homogeneous and isotropic material, the following simple relations hold between the electric and magnetic field intensities and flux densities:

$$\vec{D} = \epsilon \vec{E} \quad (2.5),$$

$$\vec{B} = \mu \vec{H} \quad (2.6).$$

In the constitutive equations, ϵ is the electrical permittivity or dielectric constant. It describes how much electric flux is generated per unit charge or a material's ability to transmit an electric field. It is opposite related to the ability of a material to polarize in response to the field. Similarly, μ is the magnetic permeability which describes the magnetization of a material in response to a magnetic field, and σ is the conductivity of the medium.

The foregoing equations are valid for arbitrary time dependence. For dealing with time-harmonic fields, the phase notations can be used to convert phasor quantities to real time-varying quantities by multiplying the phasor by $e^{j\omega t}$ and taking the real part. Then a sinusoidal electric field can be expressed as:

$$\vec{e}(x, y, z, t) = \text{Re}[\vec{E}(x, y, z)e^{j\omega t}] \quad (2.7).$$

Assuming an $e^{j\omega t}$ time dependence, Maxwell equations for a source-free ($\rho_v = 0$) medium which are linear and isotropic becomes:

$$\nabla \times \vec{E} = -j\omega\mu\vec{H} \quad (2.8),$$

$$\nabla \times \vec{H} = j\omega\epsilon\vec{E} + \sigma\vec{E} \quad (2.9),$$

$$\nabla \cdot \epsilon \vec{E} = 0 \quad (2.10),$$

$$\nabla \cdot \mu \vec{H} = 0 \quad (2.11).$$

Through the use of vector identity $\nabla \times (\nabla \times \vec{A}) = \nabla (\nabla \cdot \vec{A}) - \nabla^2 \vec{A}$, the Telegrapher equation for E and H can be derived by the curl of (2.8) and (2.9):

$$\nabla^2 \vec{E} = j\omega\sigma\mu\vec{E} - \omega^2\epsilon\mu\vec{E} \quad (2.12),$$

$$\nabla^2 \vec{H} = j\omega\sigma\mu\vec{H} - \omega^2\epsilon\mu\vec{H} \quad (2.13).$$

The solution of the Telegrapher equation for $\vec{E}$ is exponentials having the general form:

$$E_x(z) = E_+ e^{-\gamma z} + E_- e^{+\gamma z} \quad (2.14),$$

which represents a superposition of forward and backward propagating waves propagate in direction Z. They are characterized by the propagation constant γ , which is composed of the attenuation constant α and the phase constant β , in the following form:

$$\gamma = \alpha + j\beta = \sqrt{j\omega\mu(\sigma + j\omega\epsilon)} \quad (2.15).$$

For the case of a good conducting media such as metals, we have $\sigma \gg \omega\epsilon$. Then (2.15) can be expanded as follows:

$$\gamma = \alpha + j\beta \approx j\omega\sqrt{\mu\epsilon} \sqrt{\frac{\sigma}{j\omega\epsilon}} = (1 + j) \sqrt{\frac{\omega\mu\sigma}{2}} \quad (2.16).$$

The skin depth (δ_s) or characteristic depth of penetration is defined as:

$$\delta_s = \frac{1}{\alpha} = \sqrt{\frac{2}{\omega\mu\sigma}} \quad (2.17),$$

where $\delta_s = 1/\alpha$ represents the distance for the amplitude of a wave decreases by a factor of $e \approx 2.718$ when the wave travels into a material. The microwave engineer

Harold A. Wheeler developed the Wheeler’s incremental inductance rule which is a novel method to calculate the self-inductance and the resistance of a line [104]. Based on Wheeler’s incremental rule, the skin effect approximation is valid when δ_s of the conductor is much less than the conductor dimensions. With the lossless case, the wave impedance can be defined to relate the electric and magnetic fields:

$$\eta = \frac{j\omega\mu}{\gamma} \tag{ 2.18).}$$

Then the associated magnetic field can be calculated as:

$$H_y = \frac{j}{\omega\mu} \frac{\partial E_x}{\partial z} = \frac{1}{\eta} (E_+ e^{-\gamma z} - E_- e^{\gamma z}) \tag{ 2.19).}$$

2.2.2. Transmission line modelling

In many ways the transmission line theory bridges the gap between field analysis and basic circuit theory. The phenomenon of electromagnetic (EM) wave propagation along transmission lines can be considered as a specialization of the Maxwell’s equations, or

Figure 2-1: Circuit model of a transmission line.

Telegrapher’s equations as discussed previously, or it can be treated as an extension of circuit theory. The circuit model of a transmission line is shown in Figure 2.1. R represents the series resistance due to the finite conductivity of the conductor, L represents the series self-inductance of the two conductor, G is the shunt conductance

of the dielectric between the two conductors, and C is the shunt capacitance per unit length along the line. The line voltage $V(z)$ and the current $I(z)$ can be expressed in the frequency domain as:

$$\frac{\partial V(z)}{\partial z} = -(R + j\omega L)I(z) \quad (2.20),$$

$$\frac{\partial I(z)}{\partial z} = -(G + j\omega C)V(z) \quad (2.21),$$

The second order steady-state Telegrapher's equations for $V(z)$ and $I(z)$ are similar to (2.12) and (2.13):

$$\frac{\partial^2 V(z)}{\partial z^2} = -(R + j\omega L)(G + j\omega C) \cdot V(z) \quad (2.22),$$

$$\frac{\partial^2 I(z)}{\partial z^2} = -(R + j\omega L)(G + j\omega C) \cdot I(z) \quad (2.23),$$

The solution of the Telegrapher's equations for $V(z)$ and $I(z)$ are given as:

$$V(z) = V^+ e^{-\gamma z} + V^- e^{\gamma z} \quad (2.24),$$

$$I(z) = \frac{1}{Z_0} (V^+ e^{-\gamma z} - V^- e^{\gamma z}) \quad (2.25),$$

where γ and Z_0 are the propagation constant and characteristic impedance of transmission line, which are given as:

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (2.26),$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (2.27),$$

where α is the attenuation constant, and β is the phase constant. For a lossless transmission line, there is no serious resistance or dielectric conductance. Then $\alpha = 0$.

For $R = 0$ and $G = 0$, γ and Z_0 can be simplified as:

$$\gamma = \sqrt{(0 + j\omega L)(0 + j\omega C)} = j\beta \quad (2.28),$$

$$Z_0 = \sqrt{\frac{0 + j\omega L}{0 + j\omega C}} = \sqrt{\frac{L}{C}} \quad (2.29).$$

In real practice, all transmission lines have loss. The loss of the transmission line is composed of three parts: conductor loss, dielectric loss, and radiation loss. In the case when the conductor's resistivity and the dielectric conductivity cannot be ignored, γ can be rearranged as:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = j\omega\sqrt{LC} \sqrt{1 - j\left(\frac{R}{\omega L} + \frac{G}{\omega C}\right) - \frac{RG}{\omega^2 LC}} \quad (2.30).$$

For a low loss transmission line, the conductor loss and dielectric loss are small, $R \ll \omega L$, $G \ll \omega C$ and $RG \ll \omega^2 LC$, then (2.30) is reduced to:

$$\gamma \approx j\omega\sqrt{LC} \sqrt{1 - j\left(\frac{R}{\omega L} + \frac{G}{\omega C}\right)} \approx j\omega\sqrt{LC} \left[1 - j\left(\frac{R}{\omega L} + \frac{G}{\omega C}\right)\right] \quad (2.31).$$

So α and β become:

$$\alpha \approx \frac{1}{2} \left(R \sqrt{\frac{C}{L}} + G \sqrt{\frac{L}{C}} \right) = \frac{1}{2} \left(\frac{R}{Z_0} + GZ_0 \right) \quad (2.32),$$

$$\beta \approx \omega\sqrt{LC} \quad (2.33).$$

The attenuation can be split into conductor loss (α_c) and dielectric loss (α_d), which are given by:

$$\alpha_c \approx \frac{R}{2Z_0} \quad (2.34),$$

$$\alpha_d \approx \frac{GZ_0}{2} \quad (2.35).$$

Then the wavelength and phase velocity for a wave to propagate along the transmission lines is:

$$\lambda = 2\pi/\beta \quad (2.36),$$

$$v_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}} = \lambda f \quad (2.37).$$

The characteristic impedance of the line can be approximated as:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \approx \sqrt{\frac{L}{C}} = \frac{1}{v_p C} \quad (2.38).$$

The standard Z_0 of transmission line is 50 Ω and 75 Ω , which are chosen for few reasons. For a coaxial cable with air as the dielectric, the minimum attenuation occurs at around 75-77 Ω . As other dielectric materials are used, the lowest attenuation drops to a value between 52-64 Ω . The second consideration is power-handling capability, which is maximized at about 30 Ω regardless of the dielectric used. The characteristic impedance of most microwave systems is standardized to 50 Ω as it is a good compromise between lowest loss and highest power handling for a given cable size. For a lossless transmission line connected to a load with impedance of Z_L , then the input impedance (Z_{in}) is given by formula:

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan \beta l}{Z_0 + jZ_L \tan \beta l} \quad (2.39).$$

When Z_0 of the transmission line and Z_L are perfectly matched, then Z_{in} will be a constant value of Z_0 . When the two are not perfectly matched, then the load termination reflects some of the energy, and Z_{in} will also fluctuate with respect to the value of βl .

In this case, the voltage reflection coefficient is given by:

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (2.40).$$

2.2.3. Wave propagation modes on transmission lines

The term of propagation mode is used to describe the field configuration along the transmission lines. Assume a lossless EM wave propagates in z direction as $e^{-\beta z}$, proceeding from the Maxwell curl Equations $\nabla \times \vec{E} = -j\omega\mu\vec{H}$, it is easy to obtain:

$$\frac{\partial E_z}{\partial y} - \frac{\partial E_y}{\partial z} = -j\omega\mu H_x \quad (2.41),$$

$$\frac{\partial E_z}{\partial x} - \frac{\partial E_x}{\partial z} = j\omega\mu H_y \quad (2.42),$$

$$\frac{\partial E_y}{\partial x} - \frac{\partial E_x}{\partial y} = -j\omega\mu H_z \quad (2.43).$$

And the spatial variation in z direction is known so that:

$$\frac{\partial(e^{-\gamma z})}{\partial z} = -\gamma(e^{-\gamma z}) \quad (2.44).$$

Consequently, these curl equations can be simplified as:

$$\frac{\partial E_z}{\partial y} + \gamma E_y = -j\omega\mu H_x \quad (2.45),$$

$$\frac{\partial E_z}{\partial x} + \gamma E_x = j\omega\mu H_y \quad (2.46),$$

$$\frac{\partial E_y}{\partial x} - \frac{\partial E_x}{\partial y} = -j\omega\mu H_z \quad (2.47).$$

By performing a similar expansion of Ampere's Equation $\nabla \times \vec{H} = j\omega\epsilon\vec{E}$, these equations can also be obtained:

$$\frac{\partial H_z}{\partial y} + \gamma H_y = j\omega\mu E_x \quad (2.48),$$

$$\frac{\partial H_z}{\partial x} + \gamma H_x = -j\omega\mu E_y \quad (2.49),$$

$$\frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} = j\omega\mu E_z \quad (2.50).$$

Combine (2.45) and (2.49), as well as (2.46) and (2.48) for H_x and E_x , (2.47) and (2.50)

for H_y and E_y , respectively:

$$H_x = \frac{-\gamma}{k_c^2} \frac{\partial H_z}{\partial x} + \frac{j\omega\epsilon}{k_c^2} \frac{\partial E_z}{\partial y} \quad (2.51),$$

$$H_y = \frac{-\gamma}{k_c^2} \frac{\partial H_z}{\partial y} - \frac{j\omega\epsilon}{k_c^2} \frac{\partial E_z}{\partial x} \quad (2.52),$$

$$E_x = \frac{-\gamma}{k_c^2} \frac{\partial E_z}{\partial x} - \frac{j\omega\mu}{k_c^2} \frac{\partial H_z}{\partial y} \quad (2.53),$$

$$E_y = \frac{-\gamma}{k_c^2} \frac{\partial E_z}{\partial y} + \frac{j\omega\mu}{k_c^2} \frac{\partial H_z}{\partial x} \quad (2.54),$$

where $k_c = \sqrt{k^2 + \gamma^2}$, which is the cutoff wavenumber. And $k = \omega\sqrt{\mu\epsilon} = 2\pi/\lambda$, which is the wavenumber of the materials that fill the transmission line or waveguide region.

a. Transverse Electric and Magnetic (TEM) Mode:

TEM mode waves have both the electric and magnetic fields in the plane transverse to the direction of propagation along the line ($E_z = 0$, and $H_z = 0$). For TEM mode, the only solution for transverse fields to be non-zero with $E_z = 0$, and $H_z = 0$ is for $k_c = 0$ and $f_c = 0$, which means there is no lower cutoff frequency for TEM mode. From $k_c = \sqrt{k^2 + \gamma^2}$, we can get:

$$\gamma = \sqrt{-k^2} = jk = \alpha + j\beta \quad (2.55).$$

This requires $\alpha = 0$ and $\beta = k$. Hence, the TEM mode requires unguided waves to move through a lossless medium or guided waves to propagating on an ideal transmission line. And when the two-conductor transmission line structure is homogeneous and the conductor is lossless, EM waves can propagate in TEM mode. The propagation velocity v_p of TEM waves propagating along the transmission line is

determined by the physical properties of the region between the conductors. The propagation velocity is determined by:

$$v_p = \frac{1}{\sqrt{\mu\epsilon}} \quad (2.56)$$

The TEM mode is one possible solution of Maxwell's equations that describes the propagation of electromagnetic fields with conductors. But it requires the media to have at least two distinct conductors. TEM mode has the benefit of no cutoff frequency and low signal distortion.

b. Transverse Electric (TE) or Transverse Magnetic (TM) modes:

There are other solutions of Maxwell's equation exist besides TEM mode. An EM wave in TE mode has $E_z = 0$, and $H_z \neq 0$. While an EM wave in TM mode has $E_z \neq 0$, and $H_z = 0$. The propagation constant (γ) of an EM wave in TE, TM, or TE-TM hybrid modes has very different characteristics than γ of an EM wave in TEM mode. For the high order modes, γ can be written as:

$$\gamma = \sqrt{k_c^2 - k^2} = jk \sqrt{1 - \left(\frac{k_c}{k}\right)^2} \quad (2.57)$$

The ratio of k_c/k can be written in terms of the cutoff frequency (f_c) as follows:

$$f_c = \frac{k_c}{k} \times f = \frac{k_c}{2\pi\sqrt{\mu\epsilon}} \quad (2.58)$$

γ can be rewritten in terms of the waveguide cutoff frequency as:

$$\gamma = \sqrt{k_c^2 - k^2} = jk \sqrt{1 - \left(\frac{f_c}{f}\right)^2} \quad (2.59)$$

The propagation constant equation reveals the cutoff frequency behavior of the high order propagation modes such as TE, TM, or TE-TM hybrid modes. When $f < f_c$, we

have $\gamma = \alpha$, which is a purely real value. The EM wave is attenuated along z in the form of $e^{-\gamma z} = e^{-\alpha z}$, which is an evanescent mode. When $f > f_c$, we have $\gamma = \beta$, which is a purely imaginary value. The EM wave is not attenuated when propagates along z direction in the form of $e^{-\gamma z} = e^{-\beta z}$, which is a propagating mode.

Hence when an EM wave has frequency higher than the cut-off frequencies (f_c), the high order propagation modes including TE, TM, and TE-TM hybrid modes will be involved. The propagation velocity of TE and TM waves is a complex function of frequency. For lossy transmission line, the phase term β in (2.26) is generally a complicated function of angular frequency (ω). The phase velocity $v_p = \omega/\beta$ will also vary with ω if the function is not linear. If the propagation velocity of the waves change with frequency, then the individual frequency components of the signals will not maintain their original phase relationship as they propagate along the transmission line, different phase velocities allow the “faster” waves to lead in phase relative to the “slower” waves. Then signal distortion will occur and such an effect is called dispersion. When the transmission line structure is inhomogeneous, which means the fields extend in more than one medium, the effective relative permittivity ϵ_{eff} is used. Microwaves propagate in a hybrid TE-TM mode along transmission lines with mixed dielectrics. The effective permittivity ϵ_{eff} of the inhomogeneous transmission lines changes with frequency as the proportion of energy stored in the different region changes.

The single-conductor mediums such as hollow metal waveguides can only support TE mode, TM mode, and hybrid TE-TM mode. Thus they cannot be used to propagate EM waves at frequencies lower than f_c . The TEM propagation mode is preferred for low loss, low distortion, and wideband applications. And it can support wave propagation at frequencies all the way down to 0^+ Hz. The transverse dimension of interconnects should

be kept small enough to avoid the high order propagation mode. The planar transmission lines such as microstrip and CPW support a quasi-TEM mode and hybrid TE and TM mode. Coaxial transmission lines support TEM mode propagation from 0 Hz to the f_c of high order modes. The TE mode, TM mode, hybrid TE-TM mode and lossy transmission lines are undesirable situations as different frequency components of the signals travel at different speeds, which causes severe signal distortion. For a dispersive transmission line, there is no single phase velocity that can be attributed to the signal as a whole. However, if the bandwidth of the signal is relatively small, or if the dispersion is not too severe, a group velocity can be defined in a meaningful way to describe the speed at which the signal propagates.

2.2.4. Finite element analysis (FEA) design

Ansoft high frequency structure simulator (HFSS) is a commercial high-performance full-wave FEM EM field solver for passive devices by Ansys, Inc. It provides a Microsoft Windows graphical user interface and integrates simulation, visualization, solid modeling, and automation for the quick and accurate design of arbitrary 3D volumetric passive devices.

The air-gapped microcoaxial components in this thesis are all designed using the HFSS software. The 3D structures of the microwave components are drawn with specified geometries. Then the proper materials are assigned for the objects, followed by defining the boundary, input/output ports, and the simulation frequency range for the desired microwave analysis. The HFSS will generate an appropriate, efficient, and accurate mesh for solving the problem using the selected solution technology. As a high-performance full-wave EM field solver, HFSS simulations provided results that were always in good agreement with the measurement results.

2.3. MCTL Fabrication Process Development

2.3.1. Designed process flow

The air-gapped microcoax technology development started with the design of the fabrication process flow, and was followed by the material recipe experimental optimization, including various key processes of lithography, sputtering, e-beam evaporation, electroplating, surface polish by the Disco surface planer with diamond bit, chemical-mechanical-polish (CMP), stripping, and surface planarization. The realization of the 3D structured air-gapped microcoax can be briefly described as the process of thick metal deposition on wafer layer-by-layer. Each metal layer deposition is realized by sequential steps composed of thick resist lithography to form the electroplating mold as well as the stop layer for CMP, metal electroplating to fill the thick resist trench, and use Disco surface planer followed by CMP to remove the over-plated metal and flatten the metal surface. During the deposition of the signal line layer, UV-sensitive polymer is deposited and patterned to form polymer straps embedded in between the signal line and the side wall of the ground for the signal line support. The sacrificial resist is stripped away through the release holes on the ground metals after the deposition of all the layers. The details of all the designed process flow steps are described in Appendix A. A brief description of the process flow is shown in Figure 2.2. To dice out the fabricated components from the substrate, few more steps can be added to the process flow, as shown in Figure 2.3. This dicing method can cut chips into arbitrary shapes.

Fig. 2. 1. Schematic of the fabrication process flow: (i) Resist patterning; (ii) Metal electroplating; (iii) Surface planer and CMP planarization; (iv) Repeat (i)-(iii) to form lower side wall; (v) Deposition of lower signal line layer; (vi) Deposition of dielectric straps as the support of signal line; (vii) Deposition of upper signal line layer; (ix) Deposition of upper side wall (xii) Deposition of top layer; (xiii) Strip resist to release the air-gapped microcoaxial structure; (iv') Step (iv) for the portion of the microcoax with lower release holes; (v') Step (v) for the portion of the microcoax with lower release holes; (viii'') Step (viii) for the portion of the microcoax with upper release holes.

Fig. 2. 2. Schematic of the fabrication process flow which provides the wafer dicing capability: (i) Pattern thin resist; (ii) Etch substrate to form trenches around the location to fabricate the components, and strip resist; (iii)-(vii) Follow the steps of “(i)-(viii)” in Figure 2.2 to build the microcoax structure; (viii) Stick the UV tape on the front surface, and use lapping or Disco surface planer process to polish the backside of the substrate till reach the etched trenches; (ix) Remove the UV tape and thick resist to get the released and diced components.

2.3.2 Material candidates of the main process steps

a. Sacrificial photoresist for structure patterning

Thick photoresist lithography is a key process step of the air-gapped microcoax fabrication technology. The thick resist forms the patterned mold for metal electroplating as well as the stop layer for metal CMP. For this purpose, the thick resist has to meet certain requirements, such as high thickness, vertical side wall, compatibility with metal electroplating and CMP processes as well as with the support layer

deposition, and the easily stripped away. Based on the literature survey, possible candidates include AZ resist [105-110], KMPR [111-113], Shipley BPR-100N [114, 115], JSR THB [116, 117], and DiaPlate 133 [118, 119]. The suggested features of all these resist candidates from the literatures are listed in Table 2.1.

SU-8 cannot be stripped fully hence not suitable to be used as the sacrificial resist. However, the difficulty of stripping makes SU-8 a possible candidate for the support material. DiaPlate 133 is not commercially available. Hence only KMPR, Shipley BPR and BPN, JSR THB and AZ resist are possible candidates to be used as the sacrificial layer for the fabrication process of the air-gapped microcoaxial devices. Experiments were done on KMPR 1025 as the sacrificial resist for the development of the fabrication process. However, the experimental results showed that KMPR 1025 was difficult to strip, and incompatible with the Au plating process. Experiments using AZ 4620 as the sacrificial resist show it was not compatible with Cu plating and SU-8 process. The positive-tone resist such as AZ 4620 dissolved in the Cu plating solution. However, it is compatible with Au plating process.

From the experiments and literature study, resists such as SU-8, KMPR 1025, AZ 4620, and DiaPlate 133 listed in Table 2.1 are not qualified for the application as the sacrificial photoresist. This left only the Shipley BPR and BPN, and JSR THB resists as the possible candidates. The compatibility with the other two processes, the metal deposition and permanent resist deposition respectively, should also be considered for choosing the most suitable sacrificial resist. The selection of the materials for the deposition of sacrificial resist, permanent resist and conductor are based on an overall consideration of all the three key processes, which will be discussed in the later part of this Section.

Table 2. 1. Thick sacrificial photoresist candidates

Resists	Material Features	Problem
KMPR	Single-spin film thickness up to 100 μm . High aspect ratio with vertical sidewalls. Suitable for Cu electroplating. Bake temperature of 100 $^{\circ}\text{C}$	Long bake time (~1 hour). Edge bead effect.
	Aqueous developer compatible (TMAH & KOH). Swell, lift and strip with Remover PG. Negative-tone photoresist.	Dewetting and bubbles. Poor adhesion to substrates. Strong agitation for stripping.
SU-8	Widely used in MEMS applications. Single-spin film thickness up 200 μm . Vertical sidewalls with aspect ratio of ~1. Bake temperature of 150 -200 $^{\circ}\text{C}$.	Difficult to strip.
	Develop with SU-8 developer, ethyl lactate, and diacetone alcohol. Negative-tone photoresist. Swell and lift off with Remover PG.	
Shipley BPR-100	Single-spin film thickness up to 100 μm . Vertical sidewalls. Excellent adhesion to all substrates. Aqueous development without swelling.	Edge bead effect. Bubbles.
	Compatible with Cu and Ni plating. Strip with BPR stripper. Negative-tone photoresist.	
Shipley BPN	Single-spin film thickness 30 - 140 μm . Suitable for Cu electroplating. Bake temperature of 120 $^{\circ}\text{C}$	Edge bead effect. Dewetting and bubbles.
	Alkaline developer compatible. Strip without image swelling.	Aggressive rinsing with IPA.

	Negative-tone photoresist.	
	Single-spin film thickness up to 700 μm .	Long bake time (few hours).
	Promoting replacement of SU-8.	Residue at corner due to overdose.
DiaPlate	Vertical sidewalls with aspect ratio of ~ 1 .	Rounding corner due to under dose.
133	Develop with Na_2CO_3 at 30°C .	Unable to strip.
	Negative-tone photoresist.	Commercially unavailable.
	Single-spin film thickness up 100 μm .	
JSR	8 μm resolution in the gh-line	Angled sidewalls with tiptop shape.
THB-521	10 μm resolution in the i-line.	Incompatible with CMP process.
	Positive-tone photoresist.	
	Single-spin film thickness 8 - 12 μm .	Angled sidewalls with tiptop shape.
AZ4620	High aspect ratio.	Incompatible with CMP process.
	Compatible with Au plating process.	Incompatible with Cu plating.
	Positive-tone photoresist.	
	Single-spin film thickness up 80 μm .	
JSR	Low temperature process $< 120^\circ\text{C}$.	Edge bead effect.
THB-	Vertical sidewalls with aspect ratio of ~ 1 .	Bubbles.
151N	Compatible with Cu plating and CMP.	
	Easy to strip.	
	Negative-tone photoresist.	

b. Dielectric material for support straps

Dielectric straps need to be deposited in between the ground shield and the signal line to form support for the signal lines. For this purpose, the suitable dielectric material needs to have certain characteristics, including the capability to be deposited into very thick layers as well as be easily patterned, have high mechanical strength, low stress and excellent electrical insulation in the operation frequency range, and withstand high temperature. Based on the literature survey, dielectric materials such as Si_3N_4 [120-123], SiO_2 [124-127], spin-on-glass (SOG) [128-132], BCB [132-136], and SU-8 [14, 15, 46,

65, 72, 106, 119] are possible candidates. The features of all the dielectric material candidates for support straps are compared in Table 2.2.

Table 2. 2. Dielectric material candidates for the support straps

Resists	Advanced Features	Problem
Si ₃ N ₄	Low loss dielectric. Easy to deposit with CVD/PECVD. High mechanical strength.	High temperature for deposition. High stress and fragile film. Difficult for thickness > 1 μm. Easy to be polished by CMP.
SiO ₂	Low loss dielectric. Easy to deposit with CVD/PECVD. High mechanical strength.	High temperature for deposition. High stress film and can easily crack. Difficult for thickness > 1.5 μm. Easy to be polished by CMP High annealing temperature ~ 400°C
Spin-on-glass	Good mechanical property Can achieve 5 μm layer by layer	Micro cracking Crack line propagation Layer delamination Edge bulging
SU-8	Single-spin film thickness up 200 μm. Negative-tone photoresist. High resistance to cracking. Low temperature process <120 °C. Break down field > 4×10 ⁷ V/m	High loss tangent Aggressive developer
BCB	Low loss tangent. Break down field ~ 5.3×10 ⁶ V/m.	High curing temperature ~ 250 °C Can easily crack.

Fig. 2. 3. Fixed-fixed beam with point load model for dielectric straps supporting the signal line of the air-gapped microcoax.

The signal lines are supported by dielectric straps that fixed on the sidewalls of the ground shielding. The dielectric strap can be modeled as a “fixed-fixed beam with a point load”, as shown in Figure 2.4. The dielectric straps’ deflection (δ) is given as:

$$\delta = \frac{Gg^3}{24EI} \quad (2.61),$$

where G is the load of signal line and any other external force, g is the gap between signal line and the ground, E is the Young’s Modulus, and I is the moment of inertia of the beam’s cross-section, which is given by:

$$I = \frac{Jt^3}{12} \quad (2.62),$$

where J and t are the straps’ width and thickness respectively. Hence δ of the straps are given as:

$$\delta = \frac{Gg^3}{2EJt^3} \quad (2.63).$$

The elongation (A) and δ of the straps have the relationship of:

$$\delta = \sqrt{(1+A)^2 g^2 - g^2} = \sqrt{2A + A^2} g \quad (2.64).$$

The straps used to support the signal lines should have enough mechanical strength so it won't break physically. Hence A should have value less than the material's maximum elongation before breaking (A_{max}) for the supported straps, with the load of signal line and even external forces. Hence large values of the mechanical parameters such as E , I , and A_{max} are desired for the dielectric materials to be used as the straps. And δ of the straps is affected by t most notably, as it is a highest order parameter of the denominator in (2.63).

The parameter g also has a notably effect on δ . The MCTL structure's cross-sectional dimension is designed based on a targeted power handling capability, which will be discussed in Chapter 3. Hence g is a fixed parameter for the straps in the microcoaxial lines with given Z_0 .

Thick resists such as SU-8, Diaplate 133, KMPR 1025, and BCB are good candidates to provide dielectric straps with good mechanical strength, in term of high values of I , E and A_{max} . As discussed in thesis, Diaplate 133 is not commercially available. KMPR 1025 can be stripped by resist stripper, and BCB requires a high curing temperature. Hence SU-8 is the best candidate. A_{max} of SU-8 is 6.5%, which can be used to calculate the maximum deflection (δ_{max}) of the SU-8 strap:

$$\delta_{max} = \sqrt{(1.065g)^2 - g^2} = 0.3664g \quad (2.65).$$

The standard deposition process of SU-8 is under a temperature < 100 °C. And the curing of SU-8 is not necessary. The support straps only occupy a small volume percentage of the space between signal line and ground and minimized the dielectric loss, which will be discussed in Chapter 3. SU-8 has been widely used to deposit crack-free thick films in fabrication of MEMS and other devices [14, 15, 44, 46, 65, 72, 137]. SU-8 straps in the air-gapped MCTL will be suspended in air hence avoided any stress

between SU-8 and substrate. And the mechanical properties of SU-8 vary widely under different processing conditions [138].

Both Si_3N_4 and SiO_2 are grown by the CVD or PECVD, which is conducted in a chamber with high temperature. The plasma and high temperature for deposition will change the chemical property of the sacrificial resist and make the resist stripping very difficult. And the Si_3N_4 and SiO_2 films usually suffer a high stress when the deposition is thicker than $1.5\text{ }\mu\text{m}$. For conventional RF MEMS such as the capacitive membranes and cantilever beams, each layer of the structures has thickness ranges from submicron ($0.1\text{ }\mu\text{m}$) to a few microns (usually $< 10\text{ }\mu\text{m}$). Si_3N_4 and SiO_2 deposited by CVD or thermal growth have a similar scale of thickness hence can be used as the dielectrics in RF MEMS. However, for the air-gapped MCTL structure, each metal layer has thickness of $100\text{ }\mu\text{m}$, which is much larger than the thickness of Si_3N_4 and SiO_2 deposited by CVD or thermal growth. Hence the small thickness values of Si_3N_4 and SiO_2 also make the two materials not suitable for straps support.

Furthermore, Si_3N_4 and SiO_2 can be easily polished by CMP process or etched by the CMP slurry and oxide etchant. This also makes them unfavorable options as the supporting straps. To the-best-of-our-knowledge, A_{max} values of Si_3N_4 and SiO_2 were not reported by any groups yet. We can assume that Si_3N_4 and SiO_2 have similar A_{max} values as other hard ceramic materials such as Al_2O_3 , SiC , whose A_{max} values are usually less than 0.5. The small values of maximum A will cause the propagation of the films' crack lines, which could be formed during thermal expansion, CMP process, and resist stripping process. The polymers such as SU-8 and BCB have much larger A_{max} values of about 6.5% and 11.2% respectively.

Composite structures formed by metals covered with thin layers of dielectrics, such as Si_3N_4 and SiO_2 , were used as supporting structures in some micro-fabricated devices. However, this kind of composite structures are more suitable for low frequency operations, as the thin dielectrics cannot block RF waves or microwaves to pass through efficiently. If such composite structures were used as the supporting straps in the air-gapped MCTL technology, metal-dielectric-metal parasitic capacitors will be formed between the signal line and supporting straps, and between ground and supporting straps. This will cause a severe power leakage and limited bandwidth of the MCTLs.

Other materials such as the spin-on-glass and BCB also have the problem of high temperature curing process and easy to crack, which go against the stripping of sacrificial resist and mechanical strength of the dielectric.

In conclusion, SU-8 is the best dielectric candidate to be used to as the signal line support. The experiments carried out to verify the most suitable thick resist for the sacrificial layer and the dielectric for the support layer are shown in Appendix B. The mechanical properties of the SU-8 straps will be further discussed in Chapter 3.

c. Metals for conductor structure

The conventional metals used in the MMIC fabrication foundry include Cu, gold (Au), alumina (Al), tantalum (Ta), titanium (Ti), tungsten (W), Ni, and Platinum (Pt). Other metals which are not commonly used in the MMIC fabrication foundry are not studied here. The development of passive components with low loss requires suitable conductor materials. Table 2.3 lists the material properties of the conventional metals, including conductivity (σ), resistivity (ρ), resistivity variation with temperature (RVWT), and coefficient of thermal expansion (CTE).

Cu and Au are the best choices for microwave and millimeter wave application as they provide the highest σ among the metal candidates. Damascene Cu plating is widely used in CMOS BEOL process, while Au sputtering is widely used in the fabrication of MMICs on III-V semiconductor substrates. The proposed 3D microcoaxial structure is targeting an ultra-high metal thickness $> 300 \mu\text{m}$ for ultra-low loss and high power handling capability. Hence the sputtering or evaporation processes for metal deposition is not suitable in this case. Both damascene Cu plating and Au electroplating process have been investigated, as described in Appendix C. Cu CMP is a standard process used for deposition of the metal layers in semiconductor industry, which enables the construction of multilayer Cu structures with uniform and controllable thickness. As an inert metal, Au is quite difficult to be polished by the conventional CMP process, although some literatures reported the possibility of conducting Au CMP [139, 140]. And the high cost also limits Au to be used massively as metal layers.

Table 2. 3. Conductivity, resistivity, and temperature coefficient of conventional metals

<i>Material</i>	σ (S/m)	ρ ($\Omega\cdot\text{m}$)	<i>RVWT</i> (K^{-1})	<i>CTE</i> m/(mK)	<i>Skin Depth</i> (μm)	
	at 20 °C	at 20 °C			0.1 GHz	40 GHz
Cu [141]	5.96×10^7	1.68×10^{-8}	0.003862	16.6×10^{-6}	6.523	0.326
Au [142]	4.10×10^7	2.44×10^{-8}	0.0034	14.2×10^{-6}	7.862	0.393
Al [142]	3.50×10^7	2.82×10^{-8}	0.0039	22.2×10^{-6}	8.452	0.427
W [142]	1.79×10^7	5.60×10^{-8}	0.0045	4.3×10^{-6}	11.91	0.602
Ni [143]	1.43×10^7	6.99×10^{-8}	0.006	13×10^{-6}	0.543~1.331	0.027~0.067
Pt [142]	9.43×10^6	1.06×10^{-7}	0.00392	9×10^{-6}	16.386	0.828
Ta [144]	7.7×10^6	1.3×10^{-7}	0.0038	6.5×10^{-6}	18.146	0.917
Ti [143]	2.38×10^6	4.20×10^{-7}	0.0038	8.6×10^{-6}	32.62	1.648

2.3.3 Optimized material combination for the MCTLs

The selection of materials for each process step is based on their suitability for the desired functionality, for the respective process steps, and on their compatibility with the materials used in subsequent steps. As discussed previously and in Appendix A, B, and C, most of the materials studied have major disadvantages that make them not suitable candidates to be used as the conductor material, thick sacrificial resist, or support straps. Cu was selected as the material to form the conductors in the MCTL structure, due to its high conductivity and compatibility with the CMP process. JSR THB 151N was selected as the sacrificial resist, as it can be deposited with large thickness, with vertical side walls, low temperature processing, easy stripping, and compatible with Cu plating and CMP process. SU-8 was selected as the polymer material to form the support straps, due to its low temperature deposition process, and high chemical resistance. Furthermore, SU-8 can be deposited with large thickness to provide sufficient mechanical strength to support the signal lines. In a previous study, SU-8 was not recommended to be used as the dielectric material for low loss passives due to its high loss tangent [72]. However, the support straps occupy less than 0.5 % of the total gap volume in the MCTL devices, which enables SU-8 to be used in the MCTL devices for low loss passives and high frequency applications.

The fabrication process steps for MCTL structures are shown in Figure 2.5. From the figure, we can see that the thick air-gapped Cu structure with SU-8 support straps is demonstrated on Si substrate. Because it benefits from the advanced thick resist deposition and patterning, and the well-developed CMP process, the developed MCTL technology can provide novel 3D structures with large thickness, vertical sidewalls, and smooth metal surface. Passives devices with arbitrary 3D structure can also be fabricated with the same developed MCTL technology.

Fig. 2. 4. SEM pictures with the (a) Top-view, and (b) 3D-view, of a MCTL at different intermediary stages of fabrication, which were lower side wall, signal line, upper side wall, and top layer respectively, from left to right.

2.4. Measurement Setups

The Agilent E8363B and PNA-X performance network analyzer (PNA) and Cascade microwave ground-signal-ground (GSG) probes were used for the device measurement, as shown in Figure 2.6. As this thesis just focuses on passive components, only the *S*-parameters were measured for device characterization. The measured *S*-parameters were then imported to the Agilent advanced design system (ADS) for analysis, which was used to derive other characteristic parameters of the tested passive devices.

The hardware setup was done by following the standard procedures for fixing the probes, adapters and bending cables, bias tees, DC cables, and RF cables. The system’s schematic setup is shown in Figure 2.7. It is recommended to use ACP probes instead of Infinity probes for measuring the fabricated MCTL devices, as these structures have suspended signal lines.

System calibration needs to be done before any measurement. The calibration started with the probe tip planarization and stage angel checking to flatten the probe tip on the probe station and then reset the stage angle to its original value (0). Finally, the calibration frequency was defined, and the standard calibration kits were used to carry out the short-open-load-through (SOLT) calibration for the single path two-port devices.

Fig. 2. 6. Microwave measurement setup for two-port passive devices.

Fig. 2. 5. Schematic of the microwave measurement system.

3. Air-gapped MCTLs Characterization and Integration

3.1. Overview of the Chapter

The design, fabrication, and measurement setup of the air-gapped MCTL technology has been discussed in the previous chapter. The studies on the characteristics of the fabricated air-gapped MCTLs will be presented in this chapter. The characteristics of the fabricated air-gapped MCTLs are also compared with those of other transmission line technologies. This chapter also discusses the integration technologies of the developed air-gapped MCTLs.

3.2. Air-gapped MCTLs

3.2.1. Structure design of air-gapped MCTLs

The schematic architecture of the air-gapped MCTLs is shown in Figure 3.1. The signal line is supported by SU-8 straps inserted periodically along the line. The ground shielding has release holes along the line to strip away sacrificial resist after the multilayer fabrication process. The air-gapped MCTLs with different cross-sectional dimensions were designed and simulated by using Ansoft HFSS.

The designed cross-sectional dimensions of the air-gapped MCTLs with $50\ \Omega$ based on 7-layer process are listed in Table 3.1. The SU-8 straps have thickness of $18\ \mu\text{m}$, width of $100\ \mu\text{m}$, and the periodicity of $1/\text{mm}$. The space between signal line and shielded ground of the microcoaxial structure is mostly filled with air, which is an ideal dielectric for low loss and non-distortion microwave propagation.

Fig. 3. 1. Schematic of the air-gapped MCTLs: (a) 3D structure; (b) Cross-sectional schematic.

Table 3. 1. Cross-sectional dimensions of 50 Ω air-gapped MCTLs

Symbol	a	b	G	W
Size	100 μm	100 μm	80 μm	140 μm

The periodic SU-8 support straps occupy less than 0.5% of the total space between the signal line and the ground shield. Hence they will have a very minor effect on the microwave performance of the transmission line. A study of the effect of the polymer straps is discussed in the later part. The effective permittivity (ϵ_{eff}) of the air-gapped MCTL, which has the dielectric formed by the mixture of air and SU-8 straps, is given as:

$$\epsilon_{eff} = 1 + q(\epsilon_r - 1) \tag{3.1},$$

where ϵ_r is the relative permittivity of the polymer support; and q is the filling factor, which can be approximated by the ratio of the volume of polymer support V_{SU-8} and the total volume of air and polymer support $V_{air} + V_{SU-8}$ in between the centre conductor and shielding structure:

$$q \approx \frac{V_{SU-8}}{V_{air} + V_{SU-8}} \quad (3.2).$$

The dielectric of the transmission line can be approximated by a homogenous material with properties very close to those of the air. In general, the attenuation of transmission lines is due to the fact that conductors are not infinitely conductive and dielectric materials are lossy. The attenuation due to the conductor loss of the designed transmission line can be approximated by a homogenous dielectric-filled rectangular coaxial cable model [145]:

$$\alpha_c = \frac{\beta \Delta Z_0}{2Z_0} = \frac{4R_s \sqrt{\epsilon_{eff}}}{\eta_0 Z_0} \left[\frac{\partial Z_0}{\partial a} + \frac{\partial Z_0}{\partial b} - \frac{\partial Z_0}{\partial w} - \frac{\partial Z_0}{\partial t} \right] \quad (3.3),$$

where η_0 is the intrinsic impedance of free space; ϵ_{eff} is given in equation (3.1); R_s is the surface resistivity, which equals $\frac{1}{\sigma \delta_s}$, where δ_s is the skin depth as defined by (2.17); and Z_0 is determined by the cross-sectional dimension. The formula for the characteristic impedance for rectangular coaxial transmission lines is given in Chen's paper [146]:

$$Z_0 = \sqrt{\frac{L_0}{C_0}} = 1/cC_0 = \frac{376.62}{\frac{4t}{b-t} + \frac{4w}{a-w} + \frac{C_{frig}}{\epsilon_{eff}}} \quad (3.4),$$

where L_0 is the inductance of the conductor per unit length; C_0 is the capacitance between center conductor and ground shield per unit length; c is the wave propagation speed; and C_{frig} is the total fringing capacitance between the center conductor and the ground shield at the four corners.

3.2.2. Measurement of air-gapped MCTLs

The SEM picture of the fabricated air-gapped MCTLs is shown in Figure 3.2. The on-wafer measurement was conducted for the fabricated microcoaxial structures in the frequency range 0.1 GHz to 40 GHz. The probe pads connected at the two ends of the transmission line can be de-embedded through various techniques [147-150]. Figure 3.3 shows the measured S -parameters of the air-gapped 1 cm long MCTLs designed to have $Z_0 = 50 \Omega$ after de-embedding the probe pads [151]. Other parameters of the MCTLs can be extracted from the measured S -parameters.

Fig. 3. 2. SEM photograph of the fabricated air-gapped MCTLs.

Fig. 3. 3. *S*-parameters of fabricated air-gapped MCTLs 1 cm long.

The propagation constant (γ) of the transmission line can be extracted by converting measured *S*-parameters into ABCD matrix using the well-known conversion equations:

$$\gamma = \frac{1}{l} \cdot \text{arccosh}(A) = \alpha + j\beta \quad (3.5),$$

where l is the length of the transmission line, α is the attenuation constant of the transmission line, and β is the phase constant of the transmission line. The attenuation of the transmission line can also be derived from the *S*-parameters directly:

$$\text{Attenuation} = \sqrt{\left| \frac{G-1}{G+1} \right|} \quad (3.6),$$

$$G = \frac{1 - S_{11}^2 + S_{21}^2}{\sqrt{[(1 + S_{11})^2 - S_{21}^2][(1 - S_{11})^2 - S_{21}^2]}} \quad (3.7).$$

Figure 3.4 compares α of the air-gapped MCTLs with other topologies fabricated on Si substrates. The air-gapped MCTLs have ultra-low α (< 0.02 dB/mm up to 40 GHz) while occupying relatively small foot print (width ≤ 500 μm). The ϵ_{eff} of the fabricated transmission line can be calculated by:

$$\epsilon_{eff} = \left(\frac{c}{V_p} \right)^2 = \left(\frac{c\beta}{2\pi f} \right)^2$$

$$(3.8),$$

where c is the speed of light, V_p is the phase velocity of the transmission line, β is the phase constant of the transmission line, and f is the wave frequency. Figure 3.5 compares ϵ_{eff} of the air-gapped MCTLs with that of other transmission line topologies. As expected, ϵ_{eff} of the fabricated transmission line is approximately 1.02, which is close to the value of air due to the low volume percentage of dielectric materials in the gap between the signal line and shielded ground. Thus, the Cu-based air-gapped microcoaxial structure provides the advantages of extremely low dielectric loss and conductor loss, both of which are desired for high quality factor (Q -factor) passives, particularly at high frequencies. As seen from the measurements, ϵ_{eff} is insensitive to the operating frequency indicating that the fabricated air-gapped MCTLs have low dispersion. The SU-8 straps occupy less than 0.5% of the total space between the central conductor and the shielded ground, and have a negligible effect on the RF performance of the transmission lines. Table 3.2 lists the important parameters of the novel transmission line topologies and our work.

The air-gapped MCTLs can be closely routed together, as the ground shielding eliminates the cross-talk between the adjacent lines effectively. Negligible electromagnetic signal will leak out through release holes on the ground shielding when the dimension of the holes is less than $1/10$ of the wavelength. In our current microcoaxial design, the dimension of release holes is set as $200\text{ }\mu\text{m} \times 200\text{ }\mu\text{m}$, which is much smaller than $\lambda/10$ for propagating electromagnetic waves with frequency less than 100 GHz. There would be negligible signal leakage from the release holes until at

Fig. 3. 4. Attenuation comparison of different transmission line topologies

Fig. 3. 5. ϵ_{eff} of different Si-based transmission line technologies.

frequencies of hundreds of GHz. A test structure formed by two 1 cm long transmission lines with a shared wall is fabricated to characterize the cross-talk of the air-gapped MCTLs, as shown in Figure 3.6. The simulated S -parameters are plotted in Figure 3.7. The crosstalk between two adjacent lines with 1 cm length is less than -60 dB at

frequency up to 20 GHz. The crosstalk performance is quite similar to the microcoax structures fabricated by PolyStrata process [152].

SU-8 has a relatively high loss tangent of about 0.043 [71, 72]. The attenuation of the air-gapped MCTL with different SU-8 strap densities is simulated and shown in Figure 3.8. And the dielectric loss caused by SU-8 straps with different volume percentage of the gaps between signal line and ground is shown in Figure 3.9.

Table 3. 2. Comparison of various improved transmission line techniques

Reference	Structure	Metal	Support Material	Line Width	Attenuation at 40 GHz	ϵ_{eff}
Pham et al [24, 153]	Microstrip	Al	SiO ₂ +Si ₃ N ₄	n.a.	1.5 dB/mm	n.a.
Takahashi et al [154]	Microstrip	Au	BCB	300 μ m	0.028 dB/mm	13.5
Six et al [67]	Microstrip	Au	BCB	200 μ m	0.133 dB/mm	2.3
Cho et al [69]	CPW	Au	SiO ₂	120 μ m	1 dB/mm	40
Todd et al [27]	CPW	Au	SiO ₂	300 μ m	0.19 dB/mm	2.3
Cheung et al [68]	CPW	Cu	SiO ₂	56 μ m	0.42 dB/mm	7.5
Jeong et al [64]	Microcoax	Cu	n.a.	100 μ m	0.08 dB/mm	n.a.
Yoon et al [76]	Microcoax	Cu	n.a.	200 μ m	n.a.	1.6
Brown et al [78, 82]	Microcoax	Ni	Ni	330 μ m	0.025 dB/mm	n.a.
Lukic et al [87]	Microcoax	Cu	n.a.	250 μ m	0.028 dB/mm	n.a.
Filipovic et al [88]	Microcoax	Cu	n.a.	630 μ m	0.012 dB/mm	n.a.
Our Work	Microcoax	Cu	SU-8	300 μ m	0.02 dB/mm	1.02

Fig. 3. 6. 3D schematic of the test structure for measuring the cross-talk between adjacent lines.

Fig. 3. 7. Cross-talk between two adjacent transmission lines: (a) Simulated and measured S_{11} and S_{21} , (b) Simulated and measured S_{31} and S_{41} .

Fig. 3. 8. Simulated attenuation of the air-gapped MCTL with different strap density.

Fig. 3. 9. Simulated dielectric loss caused by SU-8 with different volume percentage.

In the air-gapped microcoax, partially filled SU-8 strings instead of fully filled SU-8 are used to reduce the dielectric loss. The mechanical strength of the support is proportional to the percentage of SU-8 filling. Hence the improvement of the dielectric loss by SU-8 reduction is always in a tradeoff with the mechanical strength of the supports.

The signal line was formed by two layers with the same layer thickness. The SU-8 bridges are stiffened from the two sidewalls and cross over the signal line to provide signal line support. As shown in Figure 2.4, the fixed-fixed beam with a point load model can be used to analyze the supporting mechanism of the SU-8 bridges. As discussed in section 2.3.2 of Chapter 2, the deflection of the strap (δ) is given by:

$$\delta = \frac{Gg^3}{24EI} \quad (3.9),$$

And the moment of inertia (I) can be given as:

$$I = \frac{Jt^3}{12} \quad (3.10).$$

where J and t are the width and thickness of the SU-8 bridges. For the SU-8 material, the elongation (A) before break is 6.5%. Then the maximum deflection (δ_{max}) of the beam before break can be given by:

$$\delta_{max} = \sqrt{(1.065g)^2 - g^2} = 0.3664g \quad (3.11).$$

The maximum load of the microcoax signal line that each SU-8 string can support is given as:

$$F_{max} = \frac{0.7327EJt^3}{g^2} \quad (3.12).$$

The typical probing forces ranges from 5 to 10 grams (0.05N ~ 0.1N) [155]. The probing forces applied on the probe pad may transfer to other part of the signal line. So it is necessary to design the SU-8 straps along the lines to have enough mechanical strength to support the force applied on it. The Young's Modulus (E) of the SU-8 is about 2 Gpa. In order to achieve F_{max} larger than 0.1 N, SU-8 straps with 100 μm width should have t at least 18 μm .

The air-gapped MCTLs can support wave with pure TEM propagation mode. Transmission lines could be dispersive due to the departure of β from a linear function when it is lossy or has an inhomogeneous dielectric structure. The transmission line built by the microcoaxial technology can be considered as a lossless homogeneous structure in the low frequency range. The departure of β will not be significant until the frequency reaches a threshold value. The high-order propagation mode of transmission lines starts when the distance between conductors (d) are about $\lambda/2$ or more apart. The ground shielding of our designed microcoaxial structure has a cross-sectional dimension of $300 \mu m \times 300 \mu m$. Microwaves propagate in the air-gapped microcoaxial line at the speed of about c_0 . The ϵ_{eff} of the air-gapped MCTLs is about 1.02 [151]. Therefore, the microcoaxial line will be able to support pure TEM mode at frequency up to:

$$f_{max} = \frac{c_0}{2\sqrt{\epsilon_{eff}}d} \approx 500 \text{ GHz} \quad (3.13).$$

Hence the group delay (τ) of the transmission line will have a very small variance until the frequency is high enough for the departure of β or for the high-order propagation mode involved. A simulation was conducted using the HFSS simulator for τ of a 1 cm long microcoaxial transmission line. The simulated τ from 0.1 GHz up to 600 GHz is shown in Figure 3.11. τ is defined as:

$$\tau = -\partial\phi/\partial\omega \quad (3.14),$$

where $\partial\phi$ is the change of phase and $\partial\omega$ is the change of frequency. τ of the 1 cm long microcoaxial transmission lines is at a constant value of 34 ps in the simulated frequency range from 0.1 GHz to 520 GHz. The sharply changing behaviors are numerical errors generated when the software unwrap the phase around the value of $\pm \pi$. τ only starts to

Fig. 3. 10. Simulated τ of a $50\ \Omega$ microcoaxial line 1 cm long.

drift from 520 GHz and onwards. So the microcoaxial structure can sustain an ultra-low dispersion in an ultra-wide bandwidth.

3.2.3. Power handling capability of air-gapped MCTLs

Sufficient DC handling capability of the air-gapped MCTLs is required for biasing the active components. The cross-sectional dimension and conductor material of the inner conductors determine the maximum DC current. The ohmic loss calculations show that a DC current of 2.5 A necessitates an inner conductor cross-section $\geq 65\ \mu\text{m}^2$ for Cu lines. Thus, the inner conductor with height of $100\ \mu\text{m}$ shown in Figure 3.1 was simulated which includes a reasonable margin of safety. The RF power handling is defined as the maximum power of the electromagnetic waves that the air-gapped MCTLs can handle before the electric field breakdown, which would most probably happen at the sharp corners of the rectangular coaxial structures.

Fig. 3. 11. Simulation result showing the distribution of electric field for propagating microwaves with 1 W power in the microcoaxial line with cross-sectional dimensions of $300 \mu\text{m} \times 300 \mu\text{m}$.

The HFSS simulator was used to simulate the electric field distribution in the cross-section of the air-gapped MCTLs, which can be used to estimate the breakdown field. Figure 3.12 shows the simulated electric field distribution in the cross-sectional area of a 50Ω transmission lines with dimensions of $300 \mu\text{m} \times 300 \mu\text{m}$ at 40 GHz. The maximum electric field of the electromagnetic waves with power of 1 W input from a matched generator is $2.5466 \times 10^5 \text{ V/m}$. This maximum electric field for 1 W EM power can be defined as E_{1W} . As the microcoaxial structure is using air and SU-8 straps as the dielectric, the breakdown field is determined by the electric field strength of air (E_{air}). The maximum power handling capability (P_{max}) of the microcoaxial structure can thus be calculated as:

$$P_{max} = (E_{air}/E_{1W})^2 \times 1 \text{ W} \quad (3.15).$$

As the E_{air} is about $3 \times 10^6 \text{ V/m}$, the E_{1W} value of $2.5466 \times 10^5 \text{ V/m}$ of the microcoax with $300 \mu\text{m} \times 300 \mu\text{m}$ cross-sectional dimension is corresponding to a maximum power of 139 W. If higher power handling capability is desired for a microcoax-structured passive, larger cross-sectional dimensions are required. This can be realized by using

more metal layers and wider width, with the tradeoff of increased difficulty in fabrication and bigger footprints.

In order to understand the effect of cross-sectional dimension increase on power handling capability, simulations were carried out on microcoaxial transmission lines with different cross-sectional dimensions. The E_{1W} for 50 Ω transmission lines with cross-sectional dimension of 400 $\mu\text{m} \times 400 \mu\text{m}$, 500 $\mu\text{m} \times 500 \mu\text{m}$, 600 $\mu\text{m} \times 600 \mu\text{m}$, and 700 $\mu\text{m} \times 700 \mu\text{m}$ are plotted in Figure 3.13, (a)-(d) respectively. The maximum E_{air} of the microcoax transmission lines reduces from $2.5466 \times 10^5 \text{ V/m}$ to $1.2272 \times 10^5 \text{ V/m}$ when the cross-sectional dimension increase from 300 $\mu\text{m} \times 300 \mu\text{m}$ to 700 $\mu\text{m} \times 700 \mu\text{m}$, which is corresponding to an increase of the maximum power handling capability from 139 W to 600 W, as shown in Figure 3.14.

The maximum power handing capability of the microcoaxial transmission lines can also be increased by coating the microcoax inner surface with a thin layer of dielectric which supports a higher electrical break down field. Possible solutions include using a thin layer of parylene, SiO_2 or Si_3N_4 .

Fig. 3. 12. Results of FEA simulations showing the distribution of the electric field for propagating microwaves with 1 W EM power in a microcoaxial line with cross-sectional dimensions of: (a) 400 $\mu\text{m} \times 400 \mu\text{m}$, (b) 500 $\mu\text{m} \times 500 \mu\text{m}$, (c) 600 $\mu\text{m} \times 600 \mu\text{m}$, (d) 700 $\mu\text{m} \times 700 \mu\text{m}$.

Fig. 3. 13. Power handling capability of the microcoax with different cross-sectional dimensions, based only of breakdown.

3.3. Integration of Air-gapped MCTLs

Figure 3.15 shows a conceptual integration of the microcoaxial transmission lines with typical Cu BEOL interconnects giving an overview of the main features of the structure. Unlike the bulk and surface micromachined processes, the proposed structures allow the integration of the high performance passives without any changes to the existing CMOS process.

3.3.1. Design of microcoax-to-CPW transitions

Since the current microwave systems and measurement setups are mostly based on planar platforms, coaxial-to-planar transitions are required for integration and measurement of microcoaxial passives. The requirement of low loss and high impedance matched coaxial-to-planar transitions has been intensively studied, such as coax-to-CPW transition for SU-8 technology [45], coax-to-coplanar transition for SMA connectors [156], coaxial-to-microstrip transition for SMA connectors [157], and silicon-based microcoax-to-CPW transition [158].

Fig. 3. 14. Schematic of a proposal microcoaxial transmission line integrated with the advanced Cu BEOL CMOS. The microcoaxial structure can be fabricated on the top ground plate without changing the existing Cu BEOL process.

Two types of microcoax-to-CPW transitions were designed by using the HFSS simulator and fabricated together with the microcoaxial passive devices without any process modification. The outer structures of transition designs are identical, as illustrated in Figure 3.16 (a). The top-left end of both transitions has a CPW topology, which is suitable for measurement or assembly with planar passive devices. The right end of the designed transition has a coaxial topology, which can connect with microcoaxial passives seamlessly. The designed transition can be used as a probe pad

for RF testing as well as an interface for integrating microcoaxial passives with planar circuits.

The signal lines and its supporting structures of the two transition designs are shown in Figure 3.16 (b) and Figure 3.16 (c). The signal parts of both transition designs are also identical. Two ends of the transition with different orientations are connected with a vertical bend. The bending region is chamfered to compensate the excessive capacitance and avoid current accumulation from the bending. The detailed design and chamfering techniques for the bending region of the microcoaxial passives are presented in Chapter 4.

The CPW end and the coaxial end are linked by the humped metal on both ends. The supporting structure of the signal line is formed by a cylinder-shaped metal pillar, a layer of SU-8 polymer, and a metal “seat”. A parasitic capacitor is associated with the support structure, which has the metal pillar and metal “seat” as the two electrodes and the SU-8 polymer as the dielectric. The main difference between two transition designs is for the supporting structures formed by the cylinder-shaped metal pillar and the metal “seat”.

The SU-8 polymer between the metal pillar of the signal line and the metal “seat” on the ground provides the blocking of DC current to avoid short-circuit of the signal line. However, the parasitic capacitor forms a particular RF path that may result in RF loss. The path impedance (Z_p) and the parasitic capacitance (C_p) have a reverse linear relationship of $Z_p = 1/j\omega C_p$. The RF attenuation caused by the signal leakage is proportional to the RF frequency and C_p . Hence C_p should be minimized to reduce the attenuation, especially for propagating high frequency signals. In order to minimize the transition loss, the center part of the metal “seat” has been removed. The parallel plate

capacitance was eliminated and only the fringing capacitance remained, which results in a much smaller C_p . The cylinder-shaped metal pillar is designed to a minimum dimension to reduce the fringing capacitance. The holes are designed on the side walls of the ground “seat” to assist the removal of thick resist by resist stripper. The circular-shaped SU-8 film claws on the edge of the hollow shaped metal “seat” which provides strong support to the metal pillar of the signal line.

Figure 3.17 (a) shows the top view of the signal line portion of the CPW-to-coaxial transition, including the CPW end, coaxial end, and the metal link. To ease the fabrication, the transition utilizes the same design rules as the air-gapped microcoaxial transmission lines reported in [151, 159]. The supporting structures of the two transitions are shown in Figure 3.17 (b) and Figure 3.17 (c). For clarity, the SU-8 polymer layer is hidden in the drawing. Optimization of the design to minimize the parasitic capacitance is critical for the performance of the transitions. The fringing capacitance of the optimized transition is reduced to a negligible value (~ 0.4 pF) for impedance matching and loss reduction. The dimension of the remainder of the signal line was optimized to match the impedance to $50\ \Omega$ at different sections. Detailed dimensions of the signal line and its supporting structure of the transition design are listed in Table 3.3.

Fig. 3. 15. Design of the microcoaxial-to-CPW transition for the microcoaxial passive components: (a) Schematic of the microcoax-to-CPW transitions' outer structures; (b) Signal part and support structure of microcoax-to-CPW transition design 1; (c) Signal part and support structure of microcoax-to-CPW transition design 2.

Fig. 3. 16. Top view of the designed microcoax-to-CPW transitions, SU-8 polymer is not shown: (a) The signal line part; (b) The support structure of design 1; (c) The support structure of design 2.

Table 3. 3. Dimension of optimized microcoax-to-CPW transitions

Signal line	Value	Support 1	Value	Support 2	Value
a	80 μm	a	50 μm	a	80 μm
b	140 μm	b	80 μm	b	64 μm
c	20 μm	c	50 μm	c	160 μm
d	40 μm	d	40 μm	d	40 μm
e	85 μm	e	140 μm	e	22 μm
				f	30 μm

3.3.2. Electrical performance of microcoax-to-CPW transitions

The electric field plot of the signal line part and supporting structure of the microcoax-to-CPW transition is shown in Figure 3.18. The maximum electric field in the transition without ground “seat” metal removal is 900 kV/m, when thru it propagates a 1 W electromagnetic wave. However, the transition design 1 reduced the maximum electric field to 500 kV/m by removing the ground “seat” metal underneath and near the signal line metal pillar. The transition design 2 has a larger signal line metal pillar and ground

“seat” with smaller portion of metal removed. The maximum electric field is 750 kV/m when electromagnetic wave of 1 W propagates. From the electric field plot, it can be seen that both design 1 and design 2 have effectively eliminated C_p and avoided the overcrowding of the electrical charge.

In order to characterize the designed transition, testing structures which consisted of two symmetric transitions in back-to-back configuration were fabricated by using a multi-layer Cu damascene process, as shown in Figure 3.19.

The RF measurement of the testing structures was carried out using an Agilent E8363B PNA network analyzer. The measured S -parameters of the two microcoax-to-CPW transition designs are shown in Figure 3.20. One single transition of design 1 has an insertion loss less than 0.25 dB and a return loss better than 30 dB in the frequency range of 0.1 to 40 GHz, while one single transition of design 2 has an insertion loss less than 0.29 dB and a return loss better than 25 dB in the same frequency range. The measured S -parameters of our microcoax-to-CPW transition and of other reports in the literature are compared and shown in Figure 3.21. Jaimes’ work [45] is not included in the plot as it works in a different frequency range. The measured results clearly show that the transition design in this work has much improved insertion and return losses. The microcoaxial-to-CPW transition is also compared with the coaxial-to-coaxial probe pads used by PolyStrata technology [160, 161], as shown in Figure 3.22.

Fig. 3. 17. Electric field strength on the supporting structure of the microcoax-to-CPW transition at (a) cylinder shaped ground "seat" without metal removal; (b) tube shaped ground "seat" for transition design 1; (c) partially-metal-removed ground "seat" for transition design 2.

Fig. 3. 18. SEM photograph of two microcoax-to-CPW transitions connected back-to-back.

The Z_0 of the fabricated transition can be extracted from the measured S -parameters by:

$$Z_0 = 50 \times \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (3.16).$$

The real part and imaginary part of Z_0 of the fabricated transition are shown in Figure 3.23. Z_0 of the overall transition is well matched to $50 \, \Omega$ for the transition design. Although the CPW end, the vertical bending, and the coaxial end of the transition have different topologies, the Z_0 close to $50 \, \Omega$ suggests good impedance matching of the design. As the air-gapped microcoaxial structure has very low attenuation, the conductor loss and dielectric loss are negligible in a short distance [151, 159]. The loss at the microcoax-to-CPW transition could be mainly from the parasitic capacitor and power reflection due to the different structure topologies at different sections of the transition. The measurement results of low loss indicate that C_p of the support structure is minimized.

Fig. 3. 19. Measured S -parameters of the two microcoax-to-CPW transitions.

Fig. 3. 20. *S*-parameter comparison of the microcoax-to-CPW transition design 1 with the reported transitions: (a) Insertion loss comparison; (b) Return loss comparison.

(b)

Fig. 3. 21. Comparison of the probe pads of PolyStrata technology with our microcoax-to-CPW transition: (a) Compare of the simulated S_{11} ; (b) Compare of the simulated S_{21} .

Fig. 3. 22. Measured Z_0 of the two microcoax-to-CPW transitions.

It should be pointed that the designed microcoax-to-CPW transitions can be directly integrated with other microcoaxial passive devices without any process modification. The designed microcoax-to-CPW transition can be fabricated upside down on top of the conventional planar MMICs for a monolithic integration. It can also be used as a bonding pad for wire bonding or flip-chip solder-reflow bonding, to realize a system level packaging.

3.3.3. Mechanical strength of microcoax-to-CPW transitions

An external mechanical load may be applied on the CPW end of the transitions for device measurement and integration. The supporting structure was formed by the combination of the ground “seat”, the SU-8 film, and the metal pillar on the signal line. The designed transition needs to have enough mechanical strength for the intended load on it.

In the design 1 probe pad, the hollow ground “seat” has an inner radius of b . The circular SU-8 film used to load the signal line will bend at the inner part when there is a heavy load from the signal line metal pillar. Benefitting from its mechanical properties, SU-8 has been widely used in micromachining and MEMS application. It was demonstrated that SU-8 can be used to create microstrip on a thin film [15], CPW on a thin film [65], and membrane-supported CPW lines [14, 46]. Using SU-8 as the dielectric of coaxial transmission lines was also investigated by Ricardo [71, 72]. The edge-clamped circular plate model can be used to analyze the supporting mechanism of the circular SU-8 film, as shown in Figure 3.24. The deflection (δ) of the SU-8 film at the point of the sitting tip is given by:

$$\delta = \frac{Fb^2}{16\pi D} \quad (3.17),$$

$$D = \frac{Et^3}{12(1-\nu^2)} \quad (3.18),$$

where F is the concentrated center load; E is Young’s Modulus, t is the thickness of the circular plate, and ν is the Poisson’s coefficient. As discussed previously, SU-8’s maximum elongation before break is 6.5%. The maximum load (F_{max}) of metal pillar can be calculated as:

$$F_{max} = \frac{16\pi D \delta_{max}}{b^2} \quad (3.19),$$

$$\delta_{max} = \sqrt{(1.065b)^2 - b^2} = 0.3664b \quad (3.20).$$

Considering the radius (b) of the ground bottom equals to 80 μm . And the poisson’s coefficient (ν) of SU-8 equals to 0.22, the maximum load (F_{max}) on the edge clamped SU-8 plate results to be approximately 0.3225 N. This transition can be used as a probe

pad for the RF testing, as the typical probing force applied through the probe tip ranges from 5 to 10 grams ($0.05\text{N} \sim 0.1\text{N}$) [155]. However, it is not strong enough for

Fig. 3. 23. Edge clamped circular plate model for the SU-8 film supporting the signal line metal pillar at the CPW end of design 1 transition.

integration processes such as wire bonding and solder reflow bonding, which requires bonding pressures higher than 8 N based on experimental studies [162, 163].

As shown in Figure 3.16 (c) and Figure 3.17 (c), the second type of the microcoax-to-CPW transition has the signal metal pillar with a larger diameter than the hole on the ground “seat”, which can support the metal pillar from its edge. As the SU-8 plate is supported by the ground “seat”, there won’t be any deflection when it is loaded with force from the signal line metal pillar. The SU-8 plate won’t break for a vertical load on the metal pillar and is suitable for the wire bonding and solder reflow bonding.

3.3.4. Hybrid integration of the microcoax-to-CPW transition

Besides the direct fabrication and integration of the air-gapped MCTLs with CMOS Si wafer processing, as shown in Figure 3.15, the hybrid integration of the air-gapped MCTLs to different substrates and systems can be realized by using flip-chip solder-reflow bonding. Figure 3.25 shows the schematic of the microcoax-to-CPW transition connected with CPW on a Roger printed circuit board (PCB)/Si substrate by flip-chip solder-reflow bonding, whose simulated S -parameters are shown in Figure 3.26. The bonding structure has a high insertion loss for Si substrate (~ 1 dB at 40 GHz), which is

mainly caused by the dielectric loss at the CPW part. The bonding structure has a high return loss for the Roger PCB structure as the CPW structure has much bigger critical dimensions of the metal line width and gaps, thus cause a bigger dimension mismatch between the microcoax and CPW at the bonding interface.

Fig. 3. 24. Schematic of the probe pad connected with CPW on Roger PCB/Si substrate.

Fig. 3. 25. S-parameters of the probe pad connected with CPW on Roger PCB/Si substrate.

4. Microcoaxial Bends and Ultra-long True-time-delay Lines

4.1. Overview of the Chapter

The previous chapter discussed the characteristics of the fabricated air-gapped MCTLs and the integration technologies. This chapter discusses about the miniaturization of the microcoaxial passive devices using right-angled microcoaxial bends. The detailed design, fabrication, and measurement of the right-angled microcoaxial straight and curved bends are presented. This chapter also presents the ultra-long miniature true-time-delay (TTD) lines fabricated by parallel arranged lines linked by the designed straight and curved bends.

4.2. Right-angled Microcoaxial Bends

4.2.1. Introduction of right-angled bends

In RF/microwave ICs, bends are important parts in passive components for interconnection and changing orientation. The bends for planar transmission lines such as microstrip and CPW could be problematic as the bending region can lead to unfavorable current crowding effects at the corners and imbalance in ground potential, especially for high frequency applications. The optimization of the bends for planar passive components is comprehensively studied [164-170]. However, the design and optimization of bends for the rectangular coaxial passive components is not comprehensively studied. The modeling of coaxial passive components is of great technical challenge as their 3D structures bring much more complexity as compared to the planar passive components. The artificial neural networks (ANN) modeling approach provides a well-known accurate and efficient modeling for complex and

arbitrary structures [171]. The difficulty for design and optimization of the complex 3D microwave components could be resolved by using the ANN models.

4.2.2. Design of microcoaxial bends

In order to obtain the optimal dimensions of a bend structure, the model of bend structure should be set up such that it has scalable dimensions. The bend structure can be modeled as a two-port transmission line network. Its optimal dimensions should provide Z_0 that ensures matching to the system impedance (normally $50\ \Omega$). Conventionally, analytical models based on conformal mapping are typically applied to characterize the transmission line structures, such as a microstrip or a CPW. However, only specific dimension range can be obtained after using this kind of models, and which are not scalable to different sizes, as required/imposed by various conditions, such as the power needed to be transmitted by the lines as the fabrication constraints. ANNs are information processing systems that have been widely applied in the RF/microwave modeling tasks as an unconventional alternative. Through a process called training, ANNs have the ability of fitting any nonlinear behavior of passive and active component/circuit from experimental data and generating an ANN model function that can be used to describe the port-characters of that component/circuit [172, 173]. The multiplayer perceptron (MLP) is a popularly applied neural network structure. The basic processing unit/element called neuron is grouped into different layers in an MLP structure. The layers in an MLP structure are typically defined as input, output and hidden layers. The hidden layers are the core part of a neural network, and their complexity determines the complexity of the nonlinear behavior that the neural network can be trained to learn [174]. In general, the more complicated the nonlinear behavior is, the more hidden layers and neurons will be added. A widely used MLP structure

Fig. 4. 1. Systematic architecture of the artificial neural network.

consists of one input layer, one output layer and one hidden layer, referred to as a three-layer MLP (or 3LP), which is shown in Figure 4.1.

In this thesis, a 3LP ANN is applied to construct scalable models for both bend structures. As for each ANN-based scalable bend model, the inputs are dimensions and the output is Z_0 . The training samples are collected from EM simulations done using HFSS. The training process is an optimization procedure using a quasi-Newton algorithm, whose weight values and bias values in the neural network are adjusted in order to make the outputs of neural network fit to target impedance value.

Two forms of right-angled bends for microcoaxial devices are analyzed in this paper, the straight bend and the curved bend. The schematic of the two bends and the top view of their signal line layers are shown in Figure 4.2. The two ends of both bends have the same cross-sectional dimensions as the microcoaxial passive components to be connected to them for a perfect impedance match. Both the straight and curved bends occupy a footprint of $600\text{ }\mu\text{m} \times 600\text{ }\mu\text{m}$. The excessive capacitance and current crowding can be compensated by a chamfering of the signal line at the bending area. As

Fig. 4. 2. Schematic of the chamfered microcoaxial bends: (a) 3D view of the chamfered straight bend; (b) Top view of the chamfered straight bend signal line layer; (c) 3D view of the chamfered curved bend; (d) Top view of the chamfered curved bend signal line layer.

Table 4. 1. Chamfering range of the microcoaxial bends

Parameters	Minimum Value	Maximum Value
frequency	1 GHz	40 GHz
c	0 μm	280 μm
d	0 μm	300 μm
e	0 μm	280 μm
f	0 μm	280 μm

shown in Figure 4.2 (b), the straight bend chamfering can be modeled by the two parameters c and d . As shown in Figure 4.2 (d), the curved bend chamfering is following a shape of two circles with the centers at (e, e) and $(-f, -f)$ and the radius of $R1$ and $R2$ respectively. The curved bend has a gradual increased chamfering from two ends to the center. The straight and curved bends chamfering ranges are listed in Table 4.1.

Fig. 4. 3. Output of scalable ANN model of Z_0 for: (a) Straight bend; (b) Curved bend.

The outputs of the scalable ANN models of Z_0 for straight and curved bend structures are shown in Figure 4.3. The training error ($E_{training}$) is calculated by mean squared error of Z_0 from between trained ANN model (Z_0^{ANN}) to trained data (Z_0^{train}):

$$E_{training} = \frac{1}{n} \sum_{k=1}^n (Z_0^{train} - Z_0^{ANN})^2 \quad (4.1).$$

During every training procedure, the Z_0^{ANN} is compared with Z_0^{train} refer to related dimension by using (4.1) until the $E_{training}$ reach to 0.1%. The trained ANN models are

tested by EM simulation results, and the testing error (E_{testing}) is also based on mean square error, except the Z_0^{train} is changed to Z_0 from testing data (Z_0^{test}):

$$E_{\text{testing}} = \frac{1}{n} \sum_{k=1}^n (Z_0^{\text{test}} - Z_0^{\text{ANN}})^2 \quad (4.2).$$

The average E_{testing} is 0.9%. Z_0 of the designed straight bend topology ranges from 30 to 60 Ω while Z_0 of the designed curved bend topology ranges from 20 to 80 Ω . It is easy to determine the dimension parameters from the trained and tested ANN models of the microcoaxial bends for desired Z_0 .

4.2.3. Results and measurement of the designed bends

Figure 4.4 shows the simulated S -parameters of the optimized straight and curved bends. The discrepancy of the S -parameters between single straight bend and single curved bend is extremely small, and it is difficult to compare the measured results if single bend is used.

In order to characterize the designed straight and curved bends, test structures were fabricated by connecting identical bends in cascade. As shown in Figure 4.5, a straight meander line was formed by 20 right-angled straight bends connected in cascade, while the curved meander line was formed by 20 right-angled curved bends connected in cascade. Both bends have dimensions selected from the equivalent Z_0 lines of 50 Ω plotted in Figure 4.3. The detailed dimensions are listed in Table 4.2. The typical S -parameters of the fabricated meander lines are shown in Figure 4.6. The other parameters can be derived from them. From Figure 4.6, it can be seen that both the meander lines formed by 20 straight or curved bends have insertion losses less than 0.8

dB at frequency from 0.1 GHz to 40 GHz, which results in an attenuation less than 0.04 dB for each straight and curved bend in the same frequency range.

Fig. 4. 4. Simulated S-parameters of single straight bend and single curved bend.

Fig. 4. 5. SEM picture of meander lines formed by cascaded straight bends and curved bends.

Table 4. 2. Optimized chamfering of the microcoaxial bends

Parameters	c	d	e	f
Value	126 µm	176 µm	20 µm	20 µm

Z_0 is the most important parameter for the optimization of bend designs, and its value can be extracted from the S -parameters using equation (3.16). Z_0 of the straight and curved bends forming the testing structures were plotted in Figure 4.7. The real part of Z_0 is close to 50Ω while the imaginary part of Z_0 is close to 0.

The fluctuation of Z_0 along frequency is due to the impedance mismatch induced by different architectures between the CPW and microcoaxial configurations. The almost perfect Z_0 (50Ω) of the meander lines suggests that the chamfering on the signal lines of the straight and curved bends provides an accurate adjustment to the capacitance of the bending region for desired Z_0 . The optimization of different bends by the ANN models is of high accuracy.

γ of the microcoaxial bends can be determined from the S -parameters by:

$$\gamma = -\frac{1}{2l} \log \left| \frac{G-1}{G+1} \right| \quad (4.3),$$

where G is given by equation (3.7). The α and β of the microcoaxial bends can be extracted from γ as equation (3.5). The values of both α and β of the fabricated microcoaxial bends are plotted in Figure 4.8 and Figure 4.9. Both straight and curved bends have α less than 0.1 dB/mm in the frequency range 0.1 GHz to 40 GHz. As to β , both bends show a linear relation with the frequency. The Q -factors of the microcoaxial bends can be estimated by [63, 68, 175]:

$$Q = \beta/2\alpha \quad (4.4),$$

if α is expressed in units of [Neper/m], and β is expressed in units of [rad/m]. As shown in Figure 4.10, Q -factors higher than 400 are obtained for both straight and curved meander lines in the measured frequency range. Compared to the simulation results, the

higher Q -factors for the measured results could be due to the inaccuracy of the measured attenuation in the low frequency range. The high Q -factors of the meander lines indicate a low loss per degree of phase shift and also a low loss per wavelength in the designed structures.

The phase velocity (V_p) and ϵ_{eff} of the bends can be extracted from β value. The two designs have slightly different V_p values, as shown in Figure 4.11 (a). The small variation of V_p for both bends suggests a pure TEM propagation mode in the frequency range 0.1 to 40 GHz. It could be expected that the pure TEM propagation mode is supported to a much broader bandwidth beyond 40 GHz from the trend of the plot in Figure 4.11 (b). Both the microcoaxial straight and curved bends have an air-gapped structure with SU-8 straps. SU-8 has dielectric constant of ~ 4.1 . In order to provide sufficient mechanical strength, two SU-8 straps are inserted at two ends of each straight bend, which occupy about 1.83% of the total space between signal line and ground shielding. However, for each curved bend, one extra SU-8 strap is needed, which is placed at the center of the bend.

This only occupies about 0.5% of the total space between the signal line and ground shielding. As expected, the curved bend design gives a slightly smaller ϵ_{eff} shown in Figure 4.11 (b). Nevertheless, both structures have fairly small values of ϵ_{eff} , which are very close to that of air.

Although having different chamfering and bending topologies, the 50 Ω straight and curved bends show similar microwave propagation and attenuation characteristics, which are determined by the microwave characteristics of the air-gapped microcoaxial structures. The results suggest that the characteristics of the microcoaxial structures are immune to bending, if the bends have been optimized to match Z_0 perfectly. Hence, by

using such bends, the microcoaxial technology achieved a unique advantage: the passive components realized using it can be folded for size reduction while still maintaining a low attenuation, matched impedance, constant ϵ_{eff} , and extremely low cross-talk.

Fig. 4. 6. S-parameters of: (a) Meander line formed by 20 straight bends in cascade; (b) Meander line formed by 20 curved bends in cascade.

Fig. 4. 7. Characteristic impedance of: (a) Straight bend; (b) Curved bend.

(a)

(b)

Fig. 4. 8. Attenuation constant of: (a) Straight bend; (b) Curved bend.

(a)

(b)

Fig. 4. 9. Phase constant of: (a) Straight bend; (b) Curved bend.

Fig. 4. 10. Q -factor of: (a) Straight bend; (b) Curved bend.

Fig. 4. 11. V_p and ϵ_{eff} of the straight and curved bends: (a) V_p ; (b) ϵ_{eff} .

4.3. Ultra-long Microcoaxial True-Time-Delay (TTD) Lines

4.3.1. Introduction of TTD line

Delay lines have many important applications for microwave circuits and systems, such as feed-forward amplifiers, phase-lock-loops (PLLs), and phase arrays. The ability to provide a broadband TTD has always been a desired feature of the delay line to avoid

the beam squint effect. In modern portable communication systems, many desirable features are required for the delay lines, such as compact size, integration with microwave IC circuits, constant and large delay, low insertion loss, and zero or low control power or voltage. Various technologies were explored for the development of the advanced TTD lines recently, including RF MEMS with microstrip line on GaAs, alumina, and LTCC substrates [176-179], ferroelectric varactor with coplanar strip (CPS) line on high resistivity Si substrate [180], superconducting CPW line and coplanar double-spiral meander line on LaAlO₃ substrate [181], right/left-handed transmission line on PCB substrate [182], spiral microstrip line on InP substrate [183], and shielded CPW line on GaAs substrate [184].

4.3.2. Design of the ultra-long TTD line

τ of a delay line is given as:

$$\tau = L/v_g = L\sqrt{\epsilon_{eff}}/c \quad (4.5),$$

where L is the total length of the delay line, v_g is the group velocity of the wave propagation, and c is the speed of light in vacuum. It can be seen that the total τ is proportional to the length of the delay line, which needs to have a constant ϵ_{eff} at different frequencies in order to achieve a TTD feature.

The ultra-long microcoaxial TTD lines can be formed by using the designed straight and curved bends to connect parallel transmission lines together, as shown in Figure 4.12 [159]. The TTD lines are folded into 11 straight sections linked by either straight bends or curved bends, with dimensions of 1.14 cm $\times$ 0.75 cm. The electromagnetic waves are confined in their fully shielded lines. There is no crosstalk between the meandered sections adherent to each other. This unique feature is not available for planar

transmission lines such as the microstrip, CPS line, and CPW line. In addition, the microcoaxial TTD line doesn't have any substrate loss. This feature facilitates miniaturization of the ultra-long TTD line with L of 11.5 cm for chip-level device fabrication.

The straight bend and curved bend have the same cross-sectional dimensions at the two ends as the straight line parts. Since the SU-8 pillars occupy less than 0.3% of the gap volume between signal line and shielded ground, they have a negligible effect on the microwave propagation of the TTD line. Hence the TTD line has ϵ_{eff} close-to-air and provides TTD from DC to 40 GHz.

The fabrication of the microcoaxial TTD lines are realized by the same air-gapped MCTL technology. Release holes are designed along the TTD lines to strip away the thick resist after metallization for last layer. The fabrication process has a low thermal budget, and can be easily integrated with the advanced Cu BEOL CMOS process. The electromagnetic waves are fully confined within the volume delimited by the lateral grounded metallic shields of the microcoaxial structures. Hence the TTD lines can be fabricated on any type of substrate including CMOS-class silicon wafers, without degrading the microwave performance.

(a)

(b)

Fig. 4. 12. SEM picture of ultra-long TTD lines formed by: (a) Transmission lines and straight bends in cascade; (b) Transmission lines and curved bends in cascade.

4.3.3. Measurement results of the ultra-long TTD lines

The fabricated TTD line is measured on wafer using the Agilent E8363B PNA network analyzer. The measured *S*-parameters of the two TTD lines are shown in Figure 4.13. It

can be seen that both TTD lines have insertion loss less than 2.5 dB and return loss better than 15 dB at frequency from 0.1 to 40 GHz.

The ultra-long microcoaxial TTD lines achieved constant τ of 390 ps from 0.1 to 40 GHz, as shown in Figure 4.14. The measured τ of the TTD line is well matched with the simulation result. The measured Q -factors of the TTD lines are shown in Figure 4.15.

Fig. 4. 13. Measured S -parameters of the ultra-long TTD lines: (a) S_{12} & S_{21} are the insertion loss of the ultra-long TTD line formed by straight bends and transmission lines in cascade; S_{34} & S_{43} are the insertion loss of the ultra-long TTD line formed by curved bends and transmission lines in cascade; S_{11} & S_{22} are the return loss of the ultra-long TTD line formed by straight bends and transmission lines in cascade; S_{33} & S_{44} are the return loss of the ultra-long TTD line formed by curved bends and transmission lines in cascade.

Fig. 4. 14. Simulated and measured τ of the fabricated TTD lines.

Fig. 4. 15. Measured Q -factors of the fabricated TTD lines.

The TTD line performance can be characterized by its figure-of-merit (FOM) value, which is defined as the ratio of the insertion loss to τ of the TTD line. The fabricated TTD line provides a low insertion loss and a relative high τ and has a FOM value around 0.00654 dB/ps at 40 GHz. A comparison between other TTD lines previously reported in literature and our work is listed in Table 4.3. The TTD lines fabricated on GaAs and Si substrates [176, 177, 180, 184] provide a very small τ within an acceptable insertion

loss, in other words, they have a high FOM value. The TTD line on InP provides a larger τ by using an Archimedean spiral layout [183]. In order to realize a real TTD, the layout needs to be able to spread the crosstalk noise uniformly in time to minimize the cross talk penalty on both the time delay and signal waveform. This is by no means easy, as the layout design and optimization is time consuming, and requires specially-trained skills. The FOM value at 40 GHz is used for the InP TTD line to have a fair comparison with the other TTD lines, although FOM values at higher frequencies are also available for it. The TTD lines built on alumina and LTCC [178, 179] have a lower FOM value due to a lower substrate loss. In addition, the alumina and LTCC are substrates used for the system-level passive device fabrication and system integration instead of for chip-level device fabrication. The other system-level TTD lines such as superconductor on LaAlO_3 substrate [181], and composite right/left hand (CRLH) transmission line [182] can provide very large τ up to ns range. However, the superconductor TTD line requires a low temperature environment while the CRLH transmission line has a very narrow operational bandwidth; hence it is not suitable for broadband applications.

Table 4. 3. Comparison of reported TTD lines and our work

Reference	Bandwidth	Group Delay	Control Power or Voltage	Size	Return Loss	Delay Line Technology and substrate	FOM at maximum frequency (Insertion Loss/ τ)
[176]	DC to 60 GHz	4.5 ps	10 to 23 V	5.2 mm (length)	> 10 dB	MEMS NLTL on GaAs	444.4×10^{-3} dB/ps at 60 GHz
[177]	DC to 40 GHz	61 to 86 ps	0 and 75 V	3.5×2.6 mm ²	> 15 dB	MEMS NLTL on GaAs	29.07×10^{-3} dB/ps at 40 GHz
[178]	DC to 10 GHz	394 ps	N.A.	27×14 mm ²	> 10 dB	MEMS NLTL on Alumina	6.35×10^{-3} dB/ps at 10 GHz
[179]	DC to 40 GHz	530 ps	20 to 50 V	Large	> 10 dB	MEMS NLTL on LTCC	9.85×10^{-3} dB/ps at 40 GHz
[180]	DC to 20 GHz	70 to 110 ps	~ 2 μ W	2×0.33 mm ²	> 12 dB	Ferroelectric on Si/SiO ₂	30×10^{-3} dB/ps at 40 GHz
[181]	2 to 18 GHz	25 ns	High, Watts	18×15 mm ²	> 20 dB	Superconductor on LaAlO ₃	0.06×10^{-3} dB/ps at 18 GHz
[182]	2.2 to 3.8 GHz	1.45 ns	0 mW	30×15 mm ²	> 15 dB	CRLH TL on PCB	0.69×10^{-3} dB/ps at 3.75 GHz
[183]	DC to 60 GHz	200 ps	0 mW	3.9×3.2 mm ²	> 15 dB	Spiral microstrip on InP	12.5×10^{-3} dB/ps at 40 GHz
[184]	DC to 60 GHz	85 ps	0 mW	1.2×1.1 mm ²	> 10 dB	Shielded CPW on GaAs	85.88×10^{-3} dB/ps at 26 GHz
This Work	DC to 40 GHz	390 ps	0 mW	22.7×2.3 mm ²	> 15 dB	Microcoax on Si	6.41×10^{-3} dB/ps at 40 GHz

Our microcoaxial structure has the advantage of supporting pure TEM propagation mode within a wide frequency range. Higher order modes will be presented when the half wavelength of the propagation wave is shorter than the space between the conductors of the TTD line. For the microcoaxial structure, the space between the two sidewalls of shielded ground is 300 μm , which corresponds to a pure TEM mode up to 500 GHz [185]. The microcoaxial TTD line presented in this work has the lowest FOM value for on-wafer TTD lines. Moreover, it is compact, able to integrate with microwave IC circuits, supports constant and large delay, has low insertion loss, and requires no control power or control voltage. It clearly provides the best combination of desired features for microwave integrated circuits and systems. Most importantly, the microcoaxial TTD lines are simple to design and can be fabricated on any type of substrates including the low-resistivity ($\rho = 10 \Omega \cdot \text{cm}$) Si without degrading their microwave performance.

5. Microcoaxial Rat-race Hybrid, Wilkinson Power Divider, Cavity Resonators and Filters

5.1. Overview of the Chapter

The previous chapter discussed about the detailed design, fabrication, and measurement of the right-angled microcoaxial straight and curved bends. The ultra-long miniature TTD lines fabricated by using the designed straight and curved bends to link parallel arranged transmission lines were also presented in the previous chapter. In this chapter, other miniaturized microcoaxial components formed by the air-gapped MCTLs and microcoaxial bends, such as the rat-race hybrids and power combiners, will also be presented. Other microwave components, such as cavity resonators and filters, can also be fabricated by the same air-gapped MCTL technology without any process modification.

5.2. Microcoaxial Rat-race Hybrid

5.2.1. Introduction of rat-race hybrid

The rat-race hybrid is an important component used to form mixers, frequency multipliers, amplifiers, and antennas in MICs. The conventional configuration of a rat-race hybrid is formed by one $3\lambda/4$ section and three $\lambda/4$ sections of microstrips cascaded in a ring shape for a desired operational frequency. The conventional configuration occupies a large chip area, which results in a high fabrication cost. Various approaches were proposed to develop low loss miniature rat-race hybrids for on-chip microwave applications. One approach used phase inverters to shorten the $3\lambda/4$ section to a length of $\lambda/4$ [186-190]. Another popular approach added lump components to the distributed components to reduce or eliminate certain distributed portions [191-193].

Folding the transmission lines to meander shapes to realize compact hybrids was also reported [194, 195].

As discussed in previous chapter, the air-gapped MCTL technology provides the unique advantage that the transmission lines can be folded to closely routed meander structure, while still provide an ultra-low crosstalk, insertion loss, and return loss. This chapter demonstrates a novel rat-race hybrid using the developed air-gapped MCTLs and microcoaxial straight bends.

5.2.2. Design of microcoaxial rat-race hybrid

Figure 5.1 shows the schematic structure of a conventional rat-race hybrid formed by microstrip lines. The cascaded one $3\lambda/4$ section and three $\lambda/4$ sections form a circular shape to maximize the distance and isolation between each section. The proposed rat-race hybrid was developed by replacing the microstrip lines of the conventional design with MCTLs, which provides ultra-low loss, dispersion, and crosstalk for developed passive devices. By using the MCTL technology, the rat-race hybrid folds and squeezes its sections together, resulting in size reduction. The high frequency structure simulator (HFSS) was again used to design the microcoaxial rat-race hybrid for the Ku-band with a central frequency (f_0) at 16 GHz. The cross-sectional schematic of the air-gapped microcoax is illustrated in Figure 3.1 (b) and the dimensions of $50\ \Omega$ and $70.7\ \Omega$ lines used in the hybrid are listed in Table 5.1.

A scanning electron microscope (SEM) photograph of the fabricated Ku-band microcoaxial rat-race hybrid is shown in Figure 4.1. The folded lines are used in all sections for miniaturization. The size of the miniaturized design is only about 23% of the prototype ring-shaped rat-race hybrid fabricated using the same air-gapped MCTL technology.

Fig. 5. 1. Schematic structure of a conventional rat-race hybrid.

Fig. 5. 2. SEM photo of the fabricated miniature microcoaxial rat-race hybrid.

Table 5. 1. Microcoax cross-sectional dimensions of rat-race hybrid

Parameter	a	b	G	W
50 Ω Line	100 μm	100 μm	80 μm	140 μm
70.7 Ω Line	100 μm	100 μm	114 μm	72 μm

The measurement setup for the rat-race hybrid is shown in Figure 5.3. Rohde & Schwarz ZVA40 vector network analyzer (VNA) and channel switch are used for the 4-port measurement. The channel switch connects two ports and terminates the other two ports

with a 50 Ω load during measurement. The S -parameters of every two ports are measured by switching the connection of the ports. The ZVA 40 VNA is calibrated with an automatic calibration unit, which can only calibrate to the ends of coaxial cables. Hence the measured S -parameters are for the circuit formed by the device under test (DUT), RF probes with coaxial-to-CPW configuration, and coaxial connectors linking the coaxial cables and RF probes, as the dotted line circle in Figure 5.3. The S -parameters of the coaxial connectors and the RF probe of each port are also measured for the subsequent de-embedding operation. Then the measured results of the whole circuit, and the coaxial connectors and RF probes for each port are loaded into ADS for de-embedding which will provide the result for the DUT. The 4-port DUT can be measured up to the maximum frequency of 20 GHz, which is limited by our measurement system setup.

Fig. 5. 3. Schematic of the measurement setup for fabricated rat-race hybrid.

The measured S -parameters of the microcoaxial rat-race hybrid in both differential mode and symmetrical mode after de-embedding are shown in Figure 5.4. It can be seen that

the measured S -parameters are perfectly matched with the simulation results. The transmission coefficients are greater than -3.32 dB for the sum ports ($|S_{21}|$, $|S_{31}|$) and -3.25 dB for the difference ports ($|S_{24}|$, $|S_{34}|$) respectively. This is corresponding to a transmission loss of less than 0.32 dB and 0.25 dB respectively. The isolation for the isolation port ($|S_{14}|$) is better than 20 dB. And the return loss of input signal at all ports is more than 12 dB. The simulated and measured phase differences between two output ports in differential mode and symmetrical mode are plotted in Figure 5.5. The measured phase difference between two output ports at f_0 are 180.4° in difference mode ($\angle S_{31} - \angle S_{21}$) and -1.33° in symmetrical mode ($\angle S_{24} - \angle S_{34}$). The operation bandwidths are in the range of 13.5 ~ 18.8 GHz in symmetrical mode and 13.5 ~ 17.9 GHz in differential mode for phase variance less than $\pm 10^\circ$.

Table 5.2 compares our microcoaxial rat-race hybrid with the reported state-of-the-art on-chip rat-race hybrids on different substrates. Only the Ku-band rat-race hybrids are listed in the table for comparison. Our work shows the lowest transmission loss of less than 0.32 dB. The hybrid also shows comparable characteristics for other parameters, such as the isolation for the isolation ports, the reflection for input signal at all ports, and the operation bandwidth of the rat-race hybrid. The air-gapped microcoaxial structure effectively eliminated the substrate loss of the Si chips and is fully compatible with CMOS BEOL technology. Furthermore, the integration of the microcoaxial passives with the planar circuits can be done through the microcoaxial-to-CPW transition presented in Chapter 3.

Fig. 5. 4. Simulated and measured S -parameters of the microcoaxial rat-race hybrid: (a) Symmetrical mode; (b) Differential mode.

Fig. 5. 5. Simulated and measured phase difference between two output ports: upper plot is for symmetrical mode; lower plot is for differential mode.

Table 5. 2. Microcoax cross-sectional dimensions of rat-race hybrid

Parameter	Fabrication Technology	Substrate Material	Approach	f_0 [GHz]	Loss @ f_0 , [dB]		Isolation [dB]	Reflection [dB]	Bandwidth	
					Sum	Diff.			Sum	Diff.
Tseng [188]	BiCMOS	SiGe	phase inverter	14	5	n.a.	> 25	< -10	128 %	n.a.
Tseng [190]	GIPD	Glass	phase inverter	12	0.7	1.1	> 15	< -15	65 %	n.a.
Tamitsuna [191]	MMIC	GaAs	lumped distribution	15	1.2	n.a.	> 30	< -20	18 %	n.a.
Ng [192]	3D MMIC	GaAs	lumped distribution	12	2.2	n.a.	> 25	< -15	25 %	n.a.
This Work	Microcoax	Si	compact layout	16	0.32	0.25	> 25	< -12	34 %	29 %

5.3. Microcoaxial Wilkinson Power Divider

5.3.1. Introduction of Wilkinson power divider

Wilkinson power dividers (WPDs) are important components in microwave circuits and systems. They can be used to split a single input signal into two or more in phase output signals for application in power amplifiers, balanced mixers, frequency multipliers, balanced mixers, and antenna distribution circuits [196-202]. The typical 1-to-2 WPD is formed by two $\lambda/4$ transmission lines and one resistor for a given frequency band.

However, the typical WPDs usually occupy very large footprints which cause tremendous cost increase for fabrication on MICs and MMICs. Various technologies were proposed for their miniaturization. One popular approach employed coupled lines to eliminate the dummy space [201, 203]. Another popular approach used lumped components to compensate the typical WPD designs [197, 204-206]. Yet another approach used slow-waved lines to reduce the length of $\lambda/4$ transmission lines [207].

5.3.2. Design of microcoaxial Wilkinson power divider

Figure 5.6 shows the schematic structures of conventional microstrip WPDs. The two $\lambda/4$ transmission lines are usually arranged as far as possible to avoid crosstalk between each other. Hence the conventional microstrip WPDs are not area-efficient due to the large dummy space occupied. The air-gapped MCTLs proposed in this thesis can eliminate the cross-talk hence can be arranged closely together without affecting the microwave performance. Hence, using the air-gapped MCTLs to form the WPDs will provide area efficiency.

Fig. 5. 6. Schematic structures of the conventional 1-to-2 Wilkinson power dividers formed by microstrips: (a) Square configuration; (b) Circular configuration; (c) Triangle configuration.

Figure 5.7 shows the schematic 3D representation of the circular-configured prototype microcoaxial WPD and the miniature microcoaxial WPD for Ku-band application. The microstrip lines of the conventional WPD are replaced with the air-gapped MCTLs. The

size of the miniaturized WPD design is only about 54.1% of the prototype WPD with ring-shape fabricated using the same air-gapped MCTL technology. The cross-sectional schematic of the air-gapped MCTLs was illustrated in Figure 3.1 (b) and the dimensions of 50 Ω and 70.7 Ω lines used in the hybrid are also listed in Table 5.1.

The SEM photograph of a fabricated Ku-band microcoaxial WPD is shown in Figure 5.8. The gap between the two paths of the WPD is designed as 100 μm , which is corresponding to a 600 μm distance between the signal lines. The ground shield has the opening at the position of $\lambda/4$ from the merging point for connecting the 100 Ω resistor.

The microcoaxial 2-1 WPD is measured up to the maximum frequency of 20 GHz imposed by our measurement system using the Rohde & Schwarz ZVA40 VNA with channel switch. The channel switch connects two ports and terminates the third port with a 50 Ω load during measurement. The S -parameters between every two ports are measured and then combined to the S -parameters of the three-port network.

Fig. 5. 7. 3D representation of the prototype and miniature 1-to-2 microcoaxial WPDs.

Fig. 5. 8. SEM photo of the fabricated 1-to-2 microcoaxial Wilkinson power divider.

As the current air-gapped MCTL process does not include the integrated resistor during the fabrication, this measurement is conducted on the fabricated WPD without the 100 Ω resistor. The opening on the ground shield also provides an opportunity to add an external resistor between the two signal lines by solder reflow bonding and wire bonding. The simulated and measured S -parameters of the microcoaxial WPD are shown in Figure 5.9. The return loss of the output ports (S_{22} & S_{33}) and the isolation between the two output ports (S_{23} & S_{32}) are poor as the lack of the resistor causes impedance mismatch at the output ports. The fluctuation of the measured S -parameters would be most probably due to the absence of the 100 Ω resistor. Small magnitude and phase errors between the two paths of the WPD are also obtained, as shown in Figure 5.10 (a) and (b).

Fig. 5. 9. Simulated and measured *S*-parameters of the 1-to-2 WPD without 100 Ω resistors.

Fig. 5. 10. (a) Measured gain error between two paths, and (b) Phase error between two paths, for our fabricated WPD.

The comparison of the *S*-parameters for the fabricated 1-to-2 WPD and the wideband WPD fabricated using PolyStrata technology [92] is shown in Figure 5.11. The PolyStrata wideband WPD was formed by cascading multiple paired transmission lines and resistors. It has a better impedance matching and wider bandwidth, due to the

availability of the resistors and effect of the cascaded structure. The insertion loss of the PolyStrata wideband WPD is slightly higher than our 1-to-2 WPD, as the cascade structure has longer lines.

(a)

(b)

Fig. 5. 11. S-parameter comparison of the developed 1-to-2 Wilkinson Power Divider with PolyStrata wideband WPD presented in [92].

5.4. Cavity Resonators and Filters

5.4.1. Introduction of cavity resonators

High performance resonators are important blocks for building low-noise oscillators and filters. Si bulk micromachining and packaging have been widely used to develop high Q resonators for the system-in-package (SIP) applications [28-30, 34, 35, 208]. However, these resonators are not compatible with mm-wave IC applications due to the package level process. The developed MCTL 3D passive technologies have also been investigated to develop high Q on-chip resonators which can be used for mm-wave IC applications. Previously reported realizations include high Q cavity fabricated by multilayer polymer/polyimide with metal coating [16], surface micromachined half-coaxial resonator [62], and cavity resonators fabricated with various technologies, such as LTCC [17-19, 209], PolyStrata [210, 211], and a CMOS-compatible BEOL process [212]. Previously reported 3D microwave passive filters include the cavity filters fabricated by waveguide technologies [213-215], bandpass filters fabricated by microcoaxial transmission line technologies [62, 88, 137, 216], cascade cavity filters fabricated using Si micromachining technologies [34, 35, 208, 217].

This section presents the design and fabrication of the on-chip air-gapped cavity resonators and filters using the air-gapped MCTL technology. The on-chip resonators are simply formed by air boxes with metal shielding. The bandpass filters are formed by two identical cavities in cascade with a coupling window between them. The on-chip cavity resonators and filters are fabricated on 4 inch Si substrates using the air-gapped MCTL technology together with other microcoaxial passives without any process modifications. As discussed previously, the air-gapped MCTL technology provides the advantages of ultra-low loss, ultra-low dispersion, ultra-high power handling, and ultra-

high signal isolation. The cavity resonators and filters fabricated by this technology also show advantages of high Q and ultra-low loss, which are suitable for the Si-based mm-wave IC applications and further development of SOC.

5.4.2. Design of cavity resonators

Figure 5.12 shows the 3D schematic structure of a cavity resonator. The cavities are square-shaped air boxes with metal shielding. Small release holes are designed on the cap layer and sidewall of the cavity to strip away the sacrificial resist at the end of fabrication. Two sides of the cavity are connected with microcoax-to-CPW transitions which are used as probe pads for microwave measurement [218]. The feeding of the cavities is through short-circuited vias connected with the probe pads.

Fig. 5. 12. 3D schematic of the air-gapped cavity resonator with weak coupling.

The resonant frequency of a cavity resonator (f_r) is given as:

$$f_r = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{m\pi}{W_c}\right)^2 + \left(\frac{n\pi}{H_c}\right)^2 + \left(\frac{l\pi}{L_c}\right)^2} \quad (5.1),$$

where c is the speed of light in vacuum, μ_r is the permeability of the cavity filling, ϵ_r is the permittivity of the cavity filling, W_c , H_c , L_c are the width, height, length of the

cavity, and the indices m, n, l refer to the number of variations in the standing wave pattern in the x, y, z directions. Using the perturbation method to determine the power dissipated in the cavity walls with finite conductivity, Q of the resonator due only to the lossy walls (Q_{cond}) is given as:

$$Q_{cond} = \frac{(kL_c W_c)^3 H_c \eta}{2\pi^2 R_s} \frac{1}{2l^2 W_c^3 H_c + 2H_c L_c^3 + l^2 W_c^3 L_c + W_c L_c^3} \quad (5.2),$$

where k is the wavenumber and equals $\omega\sqrt{\mu_0\mu_r\epsilon_0\epsilon_r}$, η is the intrinsic impedance and R_s is the conductor surface resistivity. The dielectric Q (Q_{diel}) of a cavity due to the dielectric medium and the total unloaded Q (Q_u) are then given by:

$$Q_{diel} = \frac{1}{\tan \delta} \quad (5.3),$$

$$\frac{1}{Q_u} = \frac{1}{Q_{cond}} + \frac{1}{Q_{diel}} \quad (5.4),$$

where $\tan \delta$ is the loss tangent of the cavity filling. The initial dimensions of the cavity with perfect electric conductor (PEC) walls can be determined using (5.1) for the TE_{101} mode by simply indexing $m = 1, n = 0$, and $l = 1$. Square-faced cavities were used to simplify the design complexity, which gives $W_c = L_c$. The μ_r and ϵ_r of the designed cavity are equal to 1 for an air-filled cavity, while Q_{diel} equals to $+\infty$ as $\tan \delta$ equals to 0 for an air-filled cavity. Then f_r and Q_u are simplified as:

$$f_r = \frac{c}{\sqrt{2}W_c} \quad (5.5),$$

$$Q_u = \frac{k^3 W_c^3 H_c \eta}{4\pi^2 R_s (2H_c + W_c)} \quad (5.6).$$

The release holes are designed on the cavity to remove the photoresist after fabrication and realize the air-gapped structure. The size and density of the release holes should be

small enough to maintain a high Q but big enough to fully remove the photoresist in the cavity. In order to fully confine the electromagnetic waves in the cavity and avoid power leakage, W_H should be designed to a size much smaller than the wavelength of the microwave, which is about 5 mm for f_r of 60 GHz. The dimensions of the designed cavity resonator with f_r at 60 GHz were calculated using equation (5.5) and listed in Table 5.3.

Table 5. 3. Dimensions of the designed cavity resonators

Parameters	W_c / L_c	H_c	W_h
36 GHz Cavity	5.88 mm	0.3 mm	0.15 mm
60 GHz Cavity	3.5 mm	0.3 mm	0.15 mm

5.4.3. Single-resonator measurement

Cavity resonators with f_r of 36 GHz and 60 GHz were fabricated and measured. The fabrication of cavity resonators and filters used the process flows of the 3D air-gapped MCTL technology exactly, without any process modification. The process flow of the MCTLs, CPW-to-microcoax transition, and rectangular cavity is shown in Figure 5.13. The measurement of the cavity resonator with f_r at 36 GHz was conducted using Agilent N5244A PNA. The measurement of the cavity resonator with f_r at 60 GHz was conducted using Agilent N5247A PNA. Cascade ACP probes with 150 μ m pitch were used for the probing of the cavity resonators. An auto calibration unit was used for the N5247A PNA systems calibration, which was placed to the end of the coaxial connectors instead of the tips of RF probes. Hence, the measured S -parameters were for the resonator and RF probes in cascade. The measured S -parameters were then de-embedded through ADS to remove the element of RF probes.

Fig. 5.13. Fabrication of the 3D microcoax, microcoax-to-CPW transition, and cavity.

Fig. 5.14. SEM photographs of the fabricated cavity resonators: (a) for 36 GHz applications; (b) For 60 GHz applications.

Fig. 5. 15. Simulated and measured S_{12} & S_{21} of the cavity resonators with f_r value of: (a) 36 GHz; (b) 60 GHz.

Figure 5.14 shows the fabricated cavity resonators, whose simulated and measured S_{12} and S_{21} parameters are shown in Figure 5.15, where the f_r value of the cavity resonator can also be easily noticed. The measurement results can be used to derive the loaded Q (Q_L) of the cavity resonator, which includes the loading effects caused by external circuitry of the feeding probes and measurement load:

$$Q_L = \frac{f_r}{\Delta f_{3-dB}} \quad (5.7),$$

where Δf_{3-dB} is 3 dB bandwidth. The external (Q_e), which includes the loading effect, can be derived by:

$$S_{21}(dB) = 20 \log_{10} \left(\frac{Q_l}{Q_e} \right) \quad (5.8).$$

The unloaded Q of the cavity can be extracted from:

$$\frac{1}{Q_l} = \frac{1}{Q_u} + \frac{1}{Q_e} \quad (5.9).$$

The parameters of the designed cavity resonator were calculated from the measured S -parameters using equations (5.7) – (5.9) and the results are listed in Table 5.4. The measured results show very good agreement with the simulated results. The lower Q_u for the measured results could due to the imperfection of the materials or the fabrication process, such as surface roughness, sidewall straightness, layer-to-layer alignment accuracy, and residues of thick resist. Additionally, as is obvious from Figure 5.15, the 36 GHz cavity has a much larger footprint than the 60 GHz cavity. This could result in more non-idealities for the cavity that cause a larger difference between measured and simulated Q_u values.

Figure 5.16 compares Q_u of this work with presented on-chip resonators, which are working at different frequencies. Our 60 GHz cavity fabricated by air-gapped MCTL technology shows the highest Q_u value, which is about 2.4 times of Lee *et al*'s LTCC cavity resonator at the same resonant frequency. Miao *et al*'s cavity fabricated by CMOS BEOL technology has the lowest Q_u as the cavity on Si suffers a high substrate loss. The cavities developed by PolyStrata technology presented by Vanhille *et al* have measured Q_u of 442 and 829, with cavity height of 250 μm and 700 μm respectively. The 36 GHz cavity resonator fabricated by air-gapped MCTL technology has much

lower height (300 μm) than the Vanhille’s 36 GHz cavity resonator, which results a slightly lower Q_u . It is not difficult to infer that a higher Q_u will be achieved if our 36 GHz cavity resonator’s height is increased. It is important to notice that the air-gapped MCTL technology is fully compatible with the CMOS BEOL technology. This enables the applications of the fabricated resonators in Si-based microwave ICs.

Figure 5.16: Comparison of Q_u of on-chip resonators fabricated by different technologies.

Table 5. 4. Simulated and measured parameters of two cavities

Parameter	Simulation	Measured	Simulation	Measured
f_r	35.7 GHz	35.7 GHz	60.4 GHz	60.4 GHz
$\Delta f_{3\text{-dB}}$	0.05 GHz	0.067 GHz	0.12 GHz	0.13 GHz
S_{21}	-13.7 dB	-14 dB	-6.6 dB	-7.0 dB
Q_l	714	533	526	476
Q_e	3440	2667	1119	1060
Q_u	901	666	991	864

Fig. 5. 16. Design of the bandpass filters using two cavity resonators in cascade: (a) 3D representation of the filter design; (b) Top view of the filter design.

5.4.4. Design of two-pole bandpass cavity filters

The two-pole bandpass filters were designed using the cavity resonators described above. The schematic structure of such a two-pole bandpass filter is shown in Figure 5.17. The filters are formed by connecting two identical cavities in cascade. An open window is allocated on the shared sidewall for the coupling between two cavities. The feeds on both sides of the filters are designed to provide strong coupling from external ports to the cavity with minimized insertion loss. The cavities in the filters and the cavities resonators described previously have the same W_H and H_c hence can be fabricated together without any process modification. The circuit model of the two-pole bandpass filter is shown in Figure 5.18 (a).

Fig. 5. 17. Circuit model of the two coupled identical square cavities: (a) Fully circuit model; (b) Simplified circuit model; (c) Simplified circuit with coupling reactance divided into two parts.

Here R_a and g represent the series resistance and coupling coefficient of the feeds, respectively. The nominal capacitance, resistance, and inductance of the cavity are represented as C_n , R_n , and L_n , respectively, while the two cavities are coupled by a mutual inductance M . The coupling coefficient k between the two resonators is defined as:

$$k = M/L_n \tag{ 5.10).}$$

An optimized feed will result in a value of g close to 1. The circuit model of the cavity filters can be simplified as Figure 5.18 (b). The coupling reactance can be divided into

two parts with the symmetry plane at $a - a'$, resulting in the equivalent circuit model shown in Figure 5.18 (c). In ultra-low loss filters, R_a and R_n are close to zero. For the symmetric cavity filter structure, the electromagnetic waves in both cavities have identical or opposite phase with respect to the symmetric plane. When the electromagnetic waves have identical phase with respect to the symmetric plane, the symmetry plane is replaced by open circuit, which determines the resonant frequency f_m of each resonant circuit to be:

$$f_m = \frac{1}{2\pi\sqrt{(L_n + M)C_n}} \quad (5.11).$$

Similarly, when the electromagnetic waves have opposite phase with respect to the symmetric plane, the symmetry plane is replaced by short circuit, which changes the resonant frequency of each resonant circuit to a new value, f_e , given by:

$$f_e = \frac{1}{2\pi\sqrt{(L_n - M)C_n}} \quad (5.12).$$

The resonant frequencies f_m and f_e defines the two poles of the designed filter. From (5.11)-(5.12), f_m , f_e and f_r are not all independent, but interdependent, according to the relationship:

$$\frac{1}{f_r^2} = \frac{1}{f_m^2} + \frac{1}{f_e^2} \quad (5.13).$$

From (5.10)-(5.13), f_m and f_e are determined by both resonant frequency of each cavity resonator (f_r) and coupling coefficient (k) between two resonators. As discussed previously, f_r is determined by the dimension of the cavity, while k is determined by the dimensions of the coupling window. As mentioned, H_c and W_H of the cavity filters are the same as those for the cavity resonators listed in Table 5.3. The other optimized

dimensions of the cavity filters for 36 GHz and 60 GHz applications are listed in Table 5.5. The length of the feeds is optimized to maximize the coupling from the external ports, while the window dimensions of W_s and T_s are optimized to provide the coupling for the desired poles.

(a)

(b)

Fig. 5. 18. SEM photographs of the fabricated cavity filters for: (a) 36 GHz applications; (b) 60 GHz applications.

Table 5. 5. Dimensions of the designed cavity filters

Parameters	W_c / L_c	L_f	W_s	T_s
36 GHz Filter	5.8 mm	1.45 mm	1.68 mm	0.1 mm
60 GHz Filter	3.36 mm	0.85 mm	1.2 mm	0.1 mm

Fig. 5. 19. Simulated and measured S -parameters of the cavity filters for: (a) 36 GHz applications; (b) 60 GHz applications.

The fabricated cavity filters for 36 GHz and 60 GHz applications are shown in Figure 5.19, and their measured S -parameters are shown in Figure 5.20. The insertion loss and the return loss in the passband of the two filters are almost zero and better than -20 dB,

respectively. The 36 GHz and 60 GHz filters have the 3-dB bandwidths of 1.4 GHz and 3.8 GHz, respectively.

It can be seen that there is a good match between simulation and measurement results. The oscillation of the measured results could be due to few factors combined together. One factor is the effect of the imperfections of the fabrication process, such as surface roughness, sidewall straightness, layer-to-layer alignment accuracy, and residues of thick resist. Another factor could be the de-embedding process of the RF probes. Finally, the topology difference between the measurement system (cables and RF probes) and the fabricated devices (microcoax-to-CPW transitions and short-circuited via feeds) may also contribute to the observed oscillations.

6. Conclusions and Future Work

6.1. Overview of the Chapter

The previous chapter discussed in detail about the design, fabrication, and measurement of the rat-race hybrids, Wilkinson power dividers, cavity resonators and cavity filters. This chapter will summarize the thesis and discuss about the future works using the air-gapped MCTL technology. The highlights are described in section 6.2 and the suggested future work is discussed in section 6.3.

6.2. Conclusion of the Thesis

A disruptive air-gapped MCTL technology for Si-based microwave applications is presented in this thesis. The development procedure of the air-gapped MCTL technology, including the microwave modeling, device simulation, fabrication process development, and device measurement is described in detail. The air-gapped MCTL technology was realized after going through a hard development procedure, which includes design of the process flow, choosing thick resist and material for signal line support, and Cu plating and CMP.

The fabricated air-gapped MCTLs provide various performance advantages compared with the conventional thin-film transmission lines, such as an ultra-low attenuation <0.02 dB/mm, a constant and low ϵ_{eff} of about 1.02, crosstalk <-60 dB, and power handling up to 139 W in frequency range of 0.1 GHz to 40 GHz. The simulated results show that the air-gapped MCTLs can support pure TEM mode up to 500 GHz. The developed air-gapped MCTLs can be fabricated on the top ground plate of the CMOS

BEOL circuit for direct integration, without changing the existing Cu BEOL process. The air-gapped MCTLs can also be flip-chip bonded with different substrates for hybrid integration. Two microcoax-to-CPW transitions with low loss and perfect impedance matching were developed for integration and measurement of the microcoaxial passives fabricated by the air-gapped MCTL technology.

The right-angled microcoaxial bends were designed using the HFSS simulator and scalable ANN models. The scalable ANN models provide convenience and accuracy for designing complex structures with desired performance. The structural dimensions of the microcoaxial bends with desired impedance can be determined from the ANN models fast and accurately. The designed-and-fabricated microcoaxial straight and curved bends have the advantages of ultra-low attenuation, perfectly matched Z_0 , linear phase constant, high Q -factor, and constant V_p and ϵ_{eff} in the frequency range 0.1 GHz to 40 GHz. With the designed microcoaxial bends, the air-gapped MCTL technology has the unique advantage that the passive components which can be built with it can be folded for size reduction while still providing a low attenuation, matched impedance, constant ϵ_{eff} , and extremely low cross-talk.

Ultra-long miniature microcoaxial TTD lines were realized on Si wafers by using the designed microcoaxial bends to fold the long lines into meander structure to fit the space on a die. The 11.5 cm long TTD line provides a constant τ of 390 ps, a low insertion loss less than 2.5 dB, and return loss better than 15 dB at frequency from 0.1 to 40 GHz. Other passive components can also be realized by the air-gapped MCTL technology and miniaturized using the microcoaxial bends for on-chip application. Examples of such devices that have been implemented and test in this work include the rat-race hybrid, WPD, and cavity resonators and filters. The fabricated microcoaxial rat-race hybrid for

Ku-band application was formed by combining the air-gapped MCTLs and microcoaxial bends, and achieved about 77% size reduction of the typical ring-shaped rat-race hybrid prototype. The rat-race hybrid has achieved transmission loss of less than 0.32 dB and 0.25 dB for the sum and difference ports, respectively. The isolation for the isolation port is better than 20 dB, while the return loss of input signal at all ports is more than 12 dB. The measured phase difference between two output ports is of 180.4° in difference mode and of -1.33° in symmetrical mode at the central frequency. The fabricated WPD has achieved size reduction 45.9% of the ring-shaped prototype WPD. The measured S -parameters of the WPD are well matched with the simulated results using Ansoft HFSS simulator. The cavity resonators fabricated using the air-gapped MCTL technology have achieved measured Q values of 666 and 864 for the 36 GHz and 60 GHz applications, respectively. Two-pole bandpass filters were fabricated by connecting two identical cavity resonators in cascade. The fabricated cavity filters had insertion loss close to zero and return loss better than 20 dB for both 36 GHz and 60 GHz applications.

In conclusion, the passive components fabricated by the air-gapped MCTL technology provide various performance advantages, such as ultra-low attenuation, reduced dispersion, and crosstalk, perfect impedance match, ultra-high power handling capability, and can support pure TEM mode up to hundreds of GHz. The fabrication is based on a multilayer Cu Damascene process which is fully compatible with CMOS BEOL technology. The developed air-gapped MCTL technology provides opportunities for modern Si-based MMIC applications.

Furthermore, the developed air-gapped MCTL technology suggests some disruptive technology reformation, more precisely, some comebacks:

1. Comeback of the coaxial structure for microwave applications

The coaxial topology used by the early transmission lines has gone through an evolution that started with the rectangular coaxial topology, followed by the strip line, and turned into the microstrip and CPW, with the majority of the microwave applications having evolved from system-level to chip-level. The proposed MCTL technology could trigger the comeback of the coaxial topology for future integrated microwave applications, in the frequency range of this study (0.1 GHz to 40 GHz).

2. Comeback of Si substrates for microwave applications

The first experimental MMICs were actually built on Si substrates. As the passive devices suffered high substrate loss and the active devices suffered low power and low cutoff frequency, the III-V compound substrates replaced Si substrates for developing MMIC technologies. In these years, the rapid evolvement of CMOS technology has already improved the power performance of the active devices and pushed the cutoff frequency to hundreds of GHz. The high-quality and area-efficient on-chip passives provided by the MCTL technology could trigger the comeback of the Si-based MMICs.

3. Comeback of thick-film technology

Thick film technology has been widely used for MICs in the early stage. The metals were deposited on the PMCs using various inks pressed through patterned silk screens. Around the time of the introduction of the hybrid miniature MIC, a thick-film variant called LTCC was also introduced. For the later MMIC technologies, thin-film technology using sputtering and etching has replaced the thick-film technology for metal deposition. The air-gapped MCTL technology proposed here provides a new form of thick-film technology for development of future MMICs.

6.3. Future Work

The air-gapped MCTL technology presented in this thesis shows superior performance and provides good opportunities for Si-based microwave applications. There are many aspects of this technology that need to be explored in future work, which are detailed in the following sections.

6.3.1. Patch antenna array with Butler matrix network (BMN)

The 3D switched-beam patch antenna array with BMN can be fabricated using the air-gapped MCTL technology, as shown in Figure 6.1. The proposed antenna array would provide advantages of improved integration of the patch elements, high gain and efficiency, wide bandwidth, and perfect circular polarization performance.

Figure 6.1: Schematic 3D representation of a patch antenna array with BMN realized using air-gapped MCTL technology.

6.3.2. Improvement of WPD

The 1-to-2 WPD discussed in chapter 5 doesn't include an integrated resistor due to the lack of a thin/thick film deposition process for the air-gapped MCTL technology. A TaN layer will be added in the future process development. As the TaN resistor is suspended

between two lines, a polymer layer also needs to be added to support the suspended TaN resistor mechanically.

Besides the fabricated 1-to-2 WPD, which was reported in this thesis, a 1-to-4 WPD could also be designed and fabricated. The schematic of a miniaturized 1-to-4 WPD is shown in Figure 6.2.

Figure 6.2: Schematic 3D representation of a miniaturized microcoaxial 1-to-4 WPD for Ku-band application realized using MCTL technology.

6.3.3. Microwave system-on-chip (SOC) with package

A delicate packaging for a microwave SOC may also be developed. A schematic of the possible package for Si-based MMIC circuits using the air-gapped MCTL technology is shown in Figure 6.3, which the schematic illustrates the transition from MCTLs to SMA connectors and the transition from active circuits to passive ones. Since the signal line is fully-shielded by grounded surrounded sidewalls, fabricated air-gapped MCTLs can

cross over each other without suffering any signal crosstalk. This unique feature provides solutions for circuit miniaturization and more freedom for circuit layout.

Figure 6.3: Possible package of Si-based MMIC fabricated using the air-gapped MCTL technology.

Appendix A

A.1. Detailed process flow of the air-gapped microcoax fabrication

Step 7: CMP to polish the metal surface of the lower side wall layer.

Step 8: Electroplating to form the metal for the lower side wall layer.

Step 9: CMP to polish the metal surface of the lower side wall layer.

Step 10: Spin coat photoresist and pattern it to form plating mold for the lower signal line layer and the middle side wall (mask 3).

Step 11: Electroplating to form the metal for the lower signal line layer and the middle side wall.

Step 12: CMP to polish the metal surface of the lower signal line layer.

Step 13: Deposit and pattern dielectric material to form straps for signal line support (mask 4).

Step 14: Spin coat photoresist and pattern it to form plating mold for the upper signal line layer and the middle side wall (mask 5).

Step 15: Electroplating to form the metal for the upper signal line layer and the middle side wall.

Step 16: CMP to polish the metal surface of the upper signal line layer and the middle side wall.

Step 17: Spin coat photoresist and pattern it to form plating mold for the upper side wall (mask 6).

Step 18: Electroplating to form the metal for the upper side wall.

Step 19: CMP to polish the metal surface of the upper side wall.

Step 18: Spin coat photoresist and pattern it to form plating mold for the top layer (mask 7).

Step 19: Seed layer deposition and electroplating to form the metal for the top layer.

Step 20: CMP to polish the metal surface of the top layer.

Step 21: Strip away the thick resist to release the air-gapped microcoax structure.

Side cross-sectional view of the microcoax structure.

Appendix B

B.1. Thick sacrificial resist experiment

B.1.1. KMPR 1025

B.1.1.1. Resist lithography

KMPR 1000 i-Line photoresist is a high contrast, epoxy based photoresist that can be developed in a conventional aqueous alkaline developer (TMAH) and readily stripped from the wafer. The features of the KMPR resist include high aspect ratio imaging, vertical sidewalls, a deposited-thickness range of 4 ~ 120 μm in a single coat, aqueous developer compatible (TMAH & KOH), wet strips in conventional strippers, and excellent dry etch resistance. Fig. B. 1 shows SEM photos of the patterned KMPR resist.

Fig. B. 1. SEM photos of photolithographically patterned KMPR resist by lithography process.

Fig. B. 2. Patterned KMPR 1025 film: (a) before Au plating; (b) after Au plating.

B.1.1.2. Adhesion problem in Au plating solution

The KMPR 1025 is compatible with Cu plating process. However, it has adhesion problems with the wafer substrate for Au plating process. The KMPR always peels off after Au plating, as shown in Fig. B. 2.

B.1.1.3. Resist Stripping Problem

For the lithography of KMPR 1025, a 2.38% TMAH aqueous alkaline solution is used as the developer. It won't etch the metal line or dissolve most of the polymers or other dielectric materials. The N-methyl Pyrrolidone (NMP)-based solvents such as MICROCHEM Remover PG are used as the stripper for the KMPR. The KMPR stripper can dissolve most of polymers, such as polyamides, polyimides, polyesters, polystyrenes, polyacrylonitriles, polyvinyl chlorides, polyvinyl acetates, polyurethanes, polycarbonates, polysulfones, polymethylmethacrylate, and many copolymers. Therefore, it is not compatible with a polymer-based support material. However, the NMP-based solutions do not react with most of the metals, such as steels, aluminum, nickel, silver, gold, chromium and chromates, copper, tin, and silicon.

The air-gapped microcoaxial devices were fabricated by multilayer resist deposition. After multilayer coating and baking process, KMPR resist had problems with the

stripping. The resist was hard cured after multiple baking processes and became extremely difficult to be stripped away, as shown in Fig. B. 3.

B.1.1.4. Edge bead problem

The edge bead is a common problem for the thick photoresist. The surface profile of one single coated KMPR is measured from center to edge of a wafer by DEKTEK profiler equipment, as shown in Fig. B. 4. The edge bead could be minimized by spraying the edge bead remover (EBR) on the edge of wafer at the end of spin coating step. And the surface polish using Disco DAS 8930 surface planer can remove the edge bead and provide a uniform layer thickness.

Fig. B. 3. Incomplete strip of KMPR 1025 after multilayer depositions.

Fig. B. 4. Resist thickness from center to edge of a 3" wafer.

B.1.2. AZ 4620

AZ 4620 is a general purpose *i*-line/ *h*-line/ *g*-line photoresist engineered for performance in most electroplating and other metal deposition process environments. AZ 4620 exhibits excellent adhesion to metal seed layers and compatibility with gold plating solution. Plating bath lives are maximized with 4620's unique photo active compound (PAC) chemistry which prevents leaching of the diazonaphthoquinone (DNQ). Coated thickness range is approximately 5.0 to 15 μ m (single coat). The development solutions for AZ 4620 include AZ 400K developers, AZ 421K developers, and CD 26.

B.1.2.1. Round corner

Patterned AZ 4620 show angled sidewall and round corner at the openings. The SEM photos of the patterned AZ 4620 with 1, 2, and 3 layers are shown in Fig. B. 5.

Fig. B. 5. SEM photograph of the AZ 4620 for single, double and triple layers deposition.

B.1.2.2. Incompatibility with CMP process and resist developer

As a positive resist, AZ 4620 can be easily dissolved by the CMP slurry. It cannot be used as the stop layer for CMP process. It can also be dissolved by aggressive resist developers such as the SU-8 developer. Hence it is not compatible with the Cu damascene process and most of the resist-based materials for signal line support in microcoaxial structures.

B.1.2.3. Electroplating compatibility

AZ 4620 is suitable to pattern the electroplating mold. It has very good compatibility with Au electroplating solution and maximizes the plating bath lives. However, the positive-tone resist will dissolve in the Cu plating solution. Hence, it is not compatible with the Cu damascene process. Figure B. 6 shows the SEM photograph of the plated Au using AZ 4620 as the mold.

Fig. B. 6. SEM photograph of plated gold using AZ 4620 as the mold.

B.1.2.4. Microcoaxial structure fabrication experiment

Experiments were carried out to develop the microcoaxial structure using AZ 4620 as the sacrificial material and Au plating to form the structure material. Due to the incompatibility of AZ 4620 with the resist developer, the resist-based support is not available for the signal line in the microcoaxial structure. Hence the final device is actually a rectangular micro-waveguide instead of the microcoax. As both AZ 4620 resist and Au structure are not compatible with CMP process, a severe within-wafer thickness variance accumulated with the increase of the structure layers. This limited the total thickness that the process can achieve, as the accumulated thickness variance increased the difficulty in alignment. Figure B. 7 shows the fabrication procedure of the micro-waveguide using AZ 4620 and Au plating. The fabrication steps can be briefly described as:

(i) Bottom Layer

Coat AZ4620, lithography and electroplating Au to form the bottom layer (repeat 2 times).

(ii) Lower Side Wall

Coat AZ4620, lithography and electroplating Au to form the lower side wall (repeat 3 times).

(iii) Upper Side Wall

Coat AZ4620, lithography and electroplating Au to form the upper side wall (repeat 3 times).

(iv) Top Seeding Layer

Evaporate the Ti-Au-Ti for top seeding layer.

(v) Top Layer

Coat AZ4620, lithography and electroplating Au to form the top layer (repeat 2 times).

(vi) Structure Release

Dip the wafer into acetone to remove AZ 4620 sacrificial layers, and then use HF to etch away the Ti Seeding layer

The final micro-waveguide is realized by 10 layers of AZ 4620, and the total thickness of the structure is about 120 μm , as shown in Fig. B. 8. The thickness of each layer is about 12 μm , which is defined by the thickness of the patterned AZ 4620 on each layer. However, the exact thickness of patterned AZ 4620 varies for different layers although the same recipes for spin coat, baking, exposure, and development have been used. It is caused by the difference of the substrate condition on different layers, including the total substrate thickness, surface roughness, and thickness variance of the wafer surface.

Fig. B. 7. Microscope photographs for the fabrication steps of the microcoaxial structure using AZ 4620 lithography and Au plating.

Fig. B. 8. SEM photographs for the fabricated microcoaxial structures using AZ 4620 lithography and Au plating: (a) Overall view of the microcoaxial structures on Si wafer; (b) Zoom-in view of a microcoaxial structure.

Fig. B. 9. Patterned THB 151N film with 50 μm thickness: (a) Microscope photograph, (b) SEM photograph.

B.1.2. JSR THB 151N

THB 151N is an acrylic negative photoresist engineered with excellent plating tolerance for copper and solder plating. The resist provides high aspect ratio patterning with vertical sidewalls. The resist can form a film 25 μm to 80 μm thick with a single spin coat. The microscope and SEM pictures of patterned THB 151N are shown in Fig. B. 9.

Fig. B. 10. SEM photograph of Cu structures after CMP.

B.1.2.1. Damascene Cu Compatibility

The THB 151N resist is compatible with both the Cu plating and CMP processes. It exhibits a perfect adhesion with the substrate when immersed in the Cu plating solution, and also provides high tolerance to corrosive environments and never dissolves in the plating solution, hence maximizes the plating bath lives. This resist also has high tolerance to the CMP slurry, hence it can be used as a stop layer for CMP process. Consequently, the patterned THB 151N film can be used as the mold to define the metal shape in the plating process as well as the stop layer to define the metal height in the CMP process. Additionally, it can be fully stripped away by THB stripper, and many other conventional resist strip solutions. Figure B. 10 shows the SEM photograph of Cu structures formed by electroplating using THB 151N as the mold, CMP using THB 151N as the stop layer, and final strip of THB 151N. The metal bumps have vertical sidewalls defined by the THB 151N opening, dense metal structure deposited by electroplating, smooth and flat surface provided by CMP process, uniform thickness controlled by the THB 151N resist thickness. Note that the white marks on certain portions of the Cu bumps are formed by the residual of the CMP slurry. They can be fully removed by adding a cleaning process after the CMP process. The THB stripper fully removed THB 151N resist and no residue remained.

B.2. Support resist experiment

B.2.1. B-staged BCB

The BCB is an *i*-line-, *g*-line-, and broad band-sensitive photopolymer that was developed for use as dielectric in thin film microelectronics applications.

B.2.1.1. Lithography

The lithography of the negative BCB resist follows the standard procedure including: surface clean, adhesion promoter, spin coat, soft bake, UV exposure, immersion develop, post-develop bake, cure, and plasma descum.

B.2.1.2. Cracking

Before curing, cracking of BCB takes place, as shown in Fig. B. 11. The cracking definitely reduces the mechanical strength and compromises the integrity of the BCB film, which can no longer qualify as the support structure of the signal line if the cracking problem is not solved. The cracking can be reduced if not eliminated by a post-exposure bake (PEB) at 120°C for 2 minutes on a hotplate following by a curing step at 250°C for 1 hour in an oven.

B.2.1.3. Remaining Scum

BCB has more severe scum generation effect after the exposure and development steps of the lithography process, as the BCB has strong crosslinked structures of bisbenzocyclobutene resins. The scum effect occurring after the development of the BCB is shown in Fig. B. 12. The scum on a bare wafer substrate could be effectively removed by an O₂ reactive-ion etch (RIE). However, the BCB deposited on sacrificial photoresist and metal does dissolve together with the resist underneath. This will generate a thick layer of polymer formed by a polymer-resist mixture, which is rather difficult to be removed.

Fig. B. 11. Microscope photograph of BCB.

Fig. B. 12. Microscope photographs of BCB before descum: (a) BCB coated on bare wafer; (b) BCB coated on thick resist and plated metal.

Fig. B. 13. Microscope photographs of single layer BCB patterned with: (a) Negative, and (b) Positive patterns, respectively.

B.2.1.4. Roundcorner

Figure B. 13 shows the microscope photographs of a patterned single layer BCB. The cracking phenomena and resist scum have been eliminated by adding a post-exposure-bake and O₂ RIE process with 80W power for 20 minutes, using a plasma descumming system (PDS). However, the patterned resist has round corners of the openings hence cannot provide vertical sidewalls.

B.2.2. SU-8

SU-8 is a popular epoxy-based thick-film photoresist that has been widely applied in the MEMS industry as it can provide films with high thickness and high aspect-ratio. The standard UV exposure has been applied for SU-8 film patterning. As a negative-tone resist, the exposed and subsequently cross-linked portions of the film are rendered insoluble to liquid developers. The high optical transparency for exposure light above 360 nm makes the SU-8 ideally suited for imaging near-vertical sidewalls in very thick films. The thickness of the SU-8 can be controlled over a wide range, from 1 to more than 700 μm with a single spin coat. SU-8 2025 has modulus of elasticity (E) of 4.4 GPa and elongation before break of 6.5%, which would be able to support the signal lines

Fig. B. 14. Microscope photographs of 25 μm thick single layer SU-8 patterned on: (a) A bare Si wafer, (b) A thick resist layer with electroplated Cu.

when it is deposited with enough thickness and density. The microscope pictures of patterned 25 μm thick SU-8 films are shown in Fig. B. 14.

B.2.2.1. Low Temperature Process

In the UV lithography process for SU-8 resist, two baking steps are requested: the soft bake, and post exposure bake. In our fabrication, two-step process has been used for each of these bake steps, using a contact hotplate. The baking conditions of SU-8 2025 with 25 μm thickness are shown in Fig. B. 13. Hard bakes are normally not required. For applications where the imaged resist is to be left as part of the final device, the resist may be ramp/step-hard baked between 150-200°C on a hot plate or in a convection oven to further crosslink the material. Bake times vary based on the type of bake process and film thickness. In the microcoaxial fabrication process, multiple soft bakes will be conducted for ensuring a hard surface for the subsequent deposition of the remaining layers after the support layer deposition. Hence no curing process is required for the patterned SU-8 as these numerous sequential soft bakes would be sufficient to crosslink the material. The process temperature for SU-8 lithography is less than 100°C, which gives a big thermal margin and ensures compatibility with the CMOS BEOL process.

Table B. 1. Baking Steps of the SU-8 2025

Baking Step	Step 1	Step 2
	65 °C	95 °C
Soft Bake	1 minute	3 minutes
Post Exposure Bake	1 minute	1 minute

B.2.2.2. Permanent Material

After the exposure and post-exposure bake steps, SU-8 2000 becomes a highly cross-linked epoxy, which is extremely difficult to be removed with conventional solvent-based resist strippers. Even the Remover PG from Microchem will not remove fully cured or hard baked SU-8 2000. The Remover PG only swells and lifts off minimally cross-linked SU-8 2000. Hence, the SU-8 2000 resist is very suitable for forming the support polymer film. The resist stripper will only dissolve the sacrificial resist and won't affect the SU-8 at the final release step of the microcoax fabrication. Figure B. 15 shows the SU-8 film deposited on top of metal bumps after the stripping of the sacrificial resist. The stripping was done using the remover PG, followed by acetone, isopropyl alcohol (IPA), and deionized (DI) water cleaning. The SU-8 film is not dissolved or swelled by the stripping solution.

Fig. B. 15. SU-8 2025 on metals after sacrificial resist stripping

Appendix C

C.1. Copper and Gold Electroplating Development

C.1.1. Plating Setup

C.1.1.1. Copper bath setup

The plating bath setup requires a tank, an anode, an agitation tool, plating solution, and some plating addition agents. The suggested plating tank can be made of PVC, PVDC, polypropylene or Teflon, while the suggested anode for Cu plating is Cu plate with a 0.045% to 0.06% certified uniform phosphorous content. In our plating setup, a magnetic stirrer is used for stirring the plating solution, which is the MICROFAB SC from Cookson Electronics, including copper sulfate, sulfuric acid (H_2SO_4), hydrochloric acid (HCl), and DI water.

C.1.1.2. Gold bath setup

The Au plating setup is similar to the one for Cu plating. Proper safety measures are required against leaching of the plating container. The plating solutions are made up of DI water and MICROFAB AU chemicals from Cookson Electronics. Platinized titanium anodes should be used in the plating solution. The ions to form the Au layer on the cathode originate from the makeup chemical instead of the anode. The pH value of the solution needs to be monitored and adjusted in order to be maintained in the range 7.7 to 8.3 at room temperature by adding reagent-grade sulfuric acid or sodium hydroxide, as required to decrease or increase the pH, respectively. The optimal plating temperature is between 45 to 55 °C, with optimized plating rates between 110 to 120 mg/Amin. The agitation of the solution is also realized by a magnetic stirrer.

C.1.1.3. Cu plating solution preparation

The solution container, magnetic stirring stick, anode plate, and peripheral equipment must be leached using 10% by volume sulfuric acid heated to 50 °C for 8 hours prior to setup for the electroplating. A hotplate with magnetic stirrer was used to control the temperature of the plating solution between 21 °C to 27 °C and to provide the agitation of the plating solution. MICROFAB SC chemicals from Cookson Electronics including SC MAKE-UP, SC-28 MAKE-UP, SC MD, and SC LO with the proper ratios were pumped into the container. The Cu anode was first cleaned with acetone, IPA, alkaline soak cleaner, copper etchant, and DI water to remove the oils, organic materials, dirt, oxides, and sulfides, before being installed into the plating container.

C.1.1.4. Au plating solution preparation

The plating solution is made up of the MICROFAB AU make-up unit which provides the necessary additives for the plating solution and the gold recovery carbons (GRC) gold complex. Unlike the Cu ions provided by the anode in Cu plating setup, the Au ions are provided by the GRC gold complex in the Au plating setup. The platinized titanium anode in Au plating setup is inactive and it is necessary to top up the gold complex according to the Au consumption.

C.1.1.5. Plating procedure

A Keithley power source is used to supply the DC and forward-reverse-pulse current. The wafers are connected to the cathode to become metal plated. The plated metal's grain size is mainly controlled by the current density, and the metal thickness is controlled by the plating time. The anodes used should have a surface area between 2 to 3 times that of the cathode plateable area. The current density should be controlled between 10 to 25 ASD (ampere per square decimeter) for Cu plating and between 0.1 to

0.6 ASD for Au plating to achieve a high quality surface. For Cu plating, the agitation of the magnetic stirrer is controlled at a minimum stirring speed to provide the necessary ion supply on the cathode and avoid a poor quality surface of the plated metal due to the over-agitation. In contrast, the agitation speed for Au plating should be controlled at about 14 m/min.

C.1.2. Plating Results

Experiments of Cu and Au plating were conducted on Si wafers with Cu and Ti-Au as seeding layers. The surface quality of the plated metal is determined by multiple factors include the current density, the anode-cathode area ratio and distance, the temperature, the agitation speed, the resist opening condition, the current waveform, and etc.

Figure C. 1 shows Cu plated using different patterns. From Fig (a), it can be seen that the plated Cu with positive mask has more defects at the corners of the opening and in narrower openings. This is probably due to resist residues that remained at these regions after thick resist lithography, as well as to the fact that air bubbles trend to be trapped in these regions. From Fig (b), it can be seen that the Cu plated with a negative mask have more defects at the centers of the big openings. This is probably due to the big openings of photoresist will consume more ions of the plating solution, and cause a low ion density at that region during the plating process. In order to achieve high quality metal, the following measures should be taken: 1) The designers need to understand the phenomena that non-uniform distribution of resist opening would cause non-uniform plating speed and grain size, so should design the metal openings with uniform distribution over wafer; 2) It is necessary to use dummy patterns to fill the spaces between the devices; 3) It is necessary to avoid the too narrow or too wide openings.

The CMP process is a standard process used to improve the surface planarity and thickness uniformity of plated Cu. The CMP process is absolutely necessary in multi-layer fabrication processes, such as the CMOS BEOL process and our microcoax fabrication. Figure C. 2 shows SEM photos of the Cu after CMP process.

The Au plating results are shown in Fig. C. 3. (a) shows the plated Au structure with a relatively low surface quality. The voids could be caused by the contamination of the plating solution caused by photoresist, dust, or other particles. Constant contamination filtration and pH control of the plating solution will help improve the plating results. Brightener chemicals can also be added to improve the surface condition of the plated Au. Figure C. 3. (b) shows the Au structure plated with an optimized solution and an improved process control. As Au is more chemically stable and cannot be etched away by the CMP slurry, hence CMP is not suitable for the planarization of plated Au structures.

Fig. C. 1. SEM photographs of plated Cu using: (a) Positive patterns, (b) Negative patterns.

Fig. C. 2. SEM photographs of the plated Cu after CMP in: (a) Positive patterns, (b) Negative patterns

Fig. C. 3. SEM photographs of the plated Au: (a) With humps and voids, (b) With relative smooth surface.

Publications

Journal Papers

- [1] Yang Tian and Pak Kwon Chan, "Design of high-performance analog circuits using wideband gm-enhanced MOS composite transistors," *IEICE Transactions on Electronics*, vol. E93.C, pp. 1199-1208, 2010.
- [2] Yang Tian, Kokyan Lee, and Hong Wang, "A 390 ps on-wafer true-time-delay line developed by a novel micro-coax technology," *IEEE Microwave and Wireless Components Letters*, vol. 24, pp. 233-235, 2014.
- [3] Yang Tian, Kokyan Lee, and Hong Wang, "Air-gapped microcoaxial transmission line for ultrawide band microwave and millimeter wave ICs," *Microwave and Optical Technology Letters*, vol. 56, pp. 1462-1465, 2014.
- [4] Yang Tian, Qianqian Meng, Kokyan Lee, and Hong Wang, "Low-loss microcoax-to-CPW transition for air-gapped microcoaxial passives," *IEEE Microwave and Wireless Components Letters*, vol. 24, pp. 233-235, 2015.
- [5] Yang Tian, Nan Li, Kokyan Lee, Xiuping Li, and Hong Wang, "Right-angled microcoaxial bends for Si-based RF/microwave integrated circuits," *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 6, pp. 290-297, 2016.
- [6] Yang Tian, Hong Wang, Zhihong Liu, Qianqian Meng, and Kokyan Lee, "Low loss microcoaxial rat-race hybrid for Si-based microwave integrated circuits," *IEEE Microwave and Wireless Components Letters*, vol. 26, pp. 162-164, 2016.
- [7] Yang Tian, Hong Wang, Kokyan Lee, Zhihong Liu, and Weimeng Lim, "On-chip high Q cavity resonator for Si-based mm-wave IC applications," Submitted to *IEEE Microwave and Wireless Components Letters*, 2016.
- [8] Yang Tian, Hong Wang, Kokyan Lee, Qianqian Meng, and Kiansiong Ang, "On-chip air-gapped cavity resonator and filters for mm-wave IC applications," Submitted to *IEEE Transactions on Microwave Theory and Technologies*, 2016.

- [9] Qian Qian Meng, Hong Wang, Chong Yang Liu, Kiansiong Ang, Xin Guo, Bo Gao, Yang Tian, C. M. Manoj Kumar, and Jianjun Gao, "High-photocurrent and wide-bandwidth UTC photodiodes with dipole-doped structure," *IEEE Photonics Technology Letters*, Oct. 2014.
- [10] Bo Gao, Hong Wang, Chongyang Liu, Qianqian Meng, Yang Tian, Kiansiong Ang, and Jinhai Si, "Design and analysis of InP-based waveguide unitraveling carrier photodiode integrated on silicon-on-insulator through Al₂O₃ bonding layer," *IEEE Photonics Journal*, Oct. 2014.

Conference Papers

- [1] Yang Tian, Kokyan Lee, and Hong Wang, "An on-wafer Cu-based microcoaxial technology for wideband and low loss microwave and mm-wave applications," in *Asia-Pacific International Symposium on Electromagnetic Compatibility (APEC)*, May, 2015.
- [2] Hong Wang and Yang Tian, "Analytical studies on temperature dependence of DC characteristics of InP/GaAsSb double heterojunction bipolar transistors," in *2010 International Conference on Indium Phosphide & Related Materials (IPRM)*, 2010, pp. 1-3.
- [3] Hua Zhu, Nan Li, Qi Zhang, Weiwei Feng, Xiuping Li, Yang Tian, and Hong Wang, "Surface micro-machined high efficient and wideband cavity-backed patch antenna array for 60GHz radios," in *2013 Asia-Pacific Microwave Conference Proceedings (APMC)*, 2013, pp. 316-318.
- [4] Xiuping Li, Hua Zhu, Nan Li, Qi Zhang, Bo Ai, Yang Tian, and Hong Wang, "Surface micro-machined wideband circularly polarized antenna array for 60GHz Radios," in *Asia-Pacific Conference on Antennas and Propagation*, 2014.

Biography

- [1] J. H. Bryant, "Coaxial Transmission Lines, Related Two-Conductor Transmission Lines, Connectors, and Components: A U.S. Historical Perspective (Corrections)," *IEEE Transactions on Microwave Theory and Techniques*, vol. 32, pp. 1422-1422, 1984.
- [2] R. M. Barrett, "Microwave Printed Circuits--The Early Years," *IEEE Transactions on Microwave Theory and Techniques*, vol. 32, pp. 983-990, 1984.
- [3] E. C. Niehenke, R. A. Pucel, and I. J. Bahl, "Microwave and Millimeter-Wave Integrated Circuits," *IEEE Transactions on Microwave Theory and Techniques*, vol. 50, pp. 846-857, 2002.
- [4] D. D. Grieg and H. F. Engelmann, "Microstrip-A New Transmission Technique for the Kilomegacycle Range," *Proceedings of the IRE*, vol. 40, pp. 1644-1650, 1952.
- [5] C. P. Wen, "Coplanar Waveguide: A Surface Strip Transmission Line Suitable for Nonreciprocal Gyromagnetic Device Applications," *IEEE Transactions on Microwave Theory and Techniques*, vol. 17, pp. 1087-1090, 1969.
- [6] L. P. Dunleavy, "Monolithic Microwave IC Technology," in *RF and Microwave Semiconductor Device Handbook*, M. Golio, Ed., ed United States of America: Taylor & Francis, 2002, p. 336.
- [7] J. V. Diloranzo and D. D. Khandelwal, *GaAs FET principles and technology*, 3 ed.: Artech House microwave library, 1982.
- [8] D. M. Pozar, *Microwave Engineering*, 3rd ed. United States of America: John Wiley & Sons, 2005.
- [9] M. V. Schneider, "Microstrip dispersion," *Proceedings of the IEEE*, vol. 60, pp. 144-146, 1972.
- [10] M. Y. Frankel, S. Gupta, J. A. Valdmanis, and G. A. Mourou, "Terahertz attenuation and dispersion characteristics of coplanar transmission lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 39, pp. 910-916, 1991.
- [11] D. M. Pozar, "Microwave Engineering," ed: John Wiley & Sons, 2005, pp. 129, 154.
- [12] M. V. Schneider, "Dielectric loss in hybrid integrated circuits," *Proceedings of the IEEE*, vol. 57, pp. 1206-1207, 1969.
- [13] A. Rida, A. Margomenos, M. M. Tentzeris, and Ieee, *Wideband mm-wave Compensated 90 degrees Bends for Grounded Coplanar Waveguide and Microstrip Transmission Lines on Flexible LCP Substrates* vol. 1-4, 2009.
- [14] F. D. Mbairi and H. Hesselbom, "High frequency design and characterization of SU-8 based conductor backed coplanar waveguide transmission lines," in *International Symposium on Advanced Packaging Materials: Processes, Properties and Interfaces*, 2005, pp. 243-248.
- [15] J. R. Thorpe, D. P. Steenson, and R. E. Miles, "High frequency transmission line using micromachined polymer dielectric," *Electronics Letters*, vol. 34, pp. 1237-1238, 1998.
- [16] P. Bo, Y. Li, M. M. Tentzeris, and J. Papapolymerou, "A High-Q Millimeter-Wave Air-Lifted Cavity Resonator on Lossy Substrates," *IEEE Microwave and Wireless Components Letters*, vol. 17, pp. 571-573, 2007.
- [17] L. Rigaudeau, P. Ferrand, D. Baillargeat, S. Bila, S. Verdeyme, M. Lahti, *et al.*, "LTCC 3D resonator applied to the design of a very compact bandpass filter for

- Q band applications," in *European Microwave Conference*, Paris, France, 2005, pp. 3694-3701.
- [18] J. H. Lee, S. Pinel, J. Papapolymerou, J. Laskar, and M. M. Tentzeris, "Comparative Study of Feeding Techniques for Three-Dimensional Cavity Resonators at 60 GHz," *IEEE Transactions on Advanced Packaging*, vol. 30, pp. 115-123, 2007.
 - [19] K. Kwangseon, K. Bonsu, B. Woojin, and M. S. Song, "A 10GHz LTCC Cavity resonator with High Q," in *Asia-Pacific Microwave Conference (APMC) 2007*, pp. 1-3.
 - [20] C. Chen-Yu and G. M. Rebeiz, "Planar millimeter-wave microstrip lumped elements using micro-machining techniques," in *IEEE MTT-S International Microwave Symposium Digest*, San Diego, CA, USA, 1994, pp. 657-660 vol.2.
 - [21] V. Milanovic, M. Gaitan, E. D. Bowen, N. H. Tea, and M. E. Zaghloul, "Design and fabrication of micromachined passive microwave filtering elements in CMOS technology," in *Solid State Sensors and Actuators, 1997. TRANSDUCERS '97 Chicago., 1997 International Conference on*, 1997, pp. 1007-1010 vol.2.
 - [22] B. A. Shenouda and L. Wilson Pearson, "Micromachined waveguide for millimeter wave applications," in *International Conference on Microwave and Millimeter Wave Technology Proceedings (ICMMT)*, 1998, pp. 615-618.
 - [23] W. Y. Ali-Ahmad, C. C. Ling, and G. M. Rebeiz, "Two-dimensional dual-polarized millimeter-wave horn antenna arrays," in *Antennas and Propagation Society International Symposium, 1990. AP-S. Merging Technologies for the 90's. Digest.*, 1990, pp. 1429-1432 vol.4.
 - [24] N. P. Pham, P. M. Sarro, K. T. Ng, and J. N. Burghartz, "IC-compatible two-level bulk micromachining process module for RF silicon technology," *Electron Devices, IEEE Transactions on*, vol. 48, pp. 1756-1764, 2001.
 - [25] J. Tsauro, T. Kobayashi, Y. Murakosi, R. Maeda, and T. Suga, "A ground shielded low loss transmission line using Au-to-Au thermo compressive packaging for RF applications," in *18th IEEE International Conference on Micro Electro Mechanical Systems (MEMS) 2005*, pp. 163-166.
 - [26] S. T. Todd, X. T. Huang, J. E. Bowers, and N. C. MacDonald, "High aspect ratio CPW fabricated using silicon bulk micromachining with substrate removal," in *Asia-Pacific Microwave Conference (APMC) Singapore, 2009*, pp. 2136-2139.
 - [27] S. T. Todd, X. T. Huang, J. E. Bowers, and N. C. MacDonald, "Fabrication, Modeling, and Characterization of High-Aspect-Ratio Coplanar Waveguide," *IEEE Transactions on Microwave Theory and Techniques*, vol. 58, pp. 3790-3800, 2010.
 - [28] M. J. Hill, J. Papapolymerou, and R. W. Ziolkowski, "High-Q micromachined resonant cavities in a K-band diplexer configuration," *IEE Proceedings Microwaves, Antennas and Propagation*, vol. 148, pp. 307-312, 2001.
 - [29] M. Stickel, G. V. Eleftheriades, and P. Kremer, "High-Q bulk micromachined silicon cavity resonator at Ka-band," *Electronics Letters*, vol. 37, pp. 433-435, 2001.
 - [30] M. Stickel, P. C. Kremer, and G. V. Eleftheriades, "High-Q microstrip-fed bulk micromachined silicon cavities," in *Antennas and Propagation Society International Symposium, 2003. IEEE, 2003*, pp. 632-635 vol.2.
 - [31] I. Llamas-Garro and A. Corona-Chavez, "Micromachined Transmission Lines for Millimeter-Wave Applications," in *Electronics, Communications and Computers, 2006. CONIELECOMP 2006. 16th International Conference on*, 2006, pp. 15-15.

- [32] K. Entesari and G. M. Rebeiz, "A Low-Loss Microstrip Surface-Mount K-Band Package," in *36th European Microwave Conference*, 2006, pp. 1763-1766.
- [33] L. Yuan and J. Papapolymerou, "A novel high-Q cavity resonator for reconfigurable WPAN applications at 60 GHz," in *European Microwave Conference (EuMC)*, 2009, pp. 197-200.
- [34] S. Song and S. Kwang-Seok, "A W-Band Air-Cavity Filter Integrated on a Thin-Film Substrate," *IEEE Microwave and Wireless Components Letters*, vol. 19, pp. 200-202, 2009.
- [35] S. Song, C. S. Yoo, and K. S. Seo, "W-Band Bandpass Filter Using Micromachined Air-Cavity Resonator With Current Probes " *IEEE Microwave and Wireless Components Letters*, vol. 20, pp. 205 - 207, 08th, March 2010.
- [36] J. Papapolymerou, C. Jui-Ching, J. East, and L. P. B. Katehi, "A micromachined high-Q X-band resonator," *IEEE Microwave and Guided Wave Letters*, vol. 7, pp. 168-170, 1997.
- [37] N. I. Dib, W. P. Harokopus, Jr., P. B. Katehi, C. C. Ling, and G. M. Rebeiz, "Study of a novel planar transmission line," in *IEEE MTT-S International Microwave Symposium Digest*, 1991, pp. 623-626 vol.2.
- [38] C. Chen-Yu and G. M. Rebeiz, "Planar microwave and millimeter-wave lumped elements and coupled-line filters using micro-machining techniques," *IEEE Transactions on Microwave Theory and Techniques*, vol. 43, pp. 730-738, 1995.
- [39] S. V. Robertson, L. P. B. Katehi, and G. M. Rebeiz, "Micromachined W-band filters," *IEEE Transactions on Microwave Theory and Techniques*, vol. 43, pp. 730-738, 1996.
- [40] P. Blondy, A. R. Brown, D. Crost, and G. M. Rebeiz, "Low loss micromachined filters for millimeter-wave telecommunication systems," in *IEEE MTT-S International Microwave Symposium Digest*, Baltimore, Maryland, USA, 1998, pp. 1181-1184.
- [41] J. E. Harriss, L. W. Pearson, X. Wang, C. H. Barren, Jr., and A. V. Pham, "Membrane-supported Ka band resonator employing organic micromachined packaging," in *IEEE MTT-S International Microwave Symposium Digest*, Boston, MA, USA, 2000, pp. 1225-1228.
- [42] J. R. Reid and R. T. Webster, "A compact integrated coaxial V-band bandpass filter," in *Antennas and Propagation Society International Symposium, 2004. IEEE*, 2004, pp. 990-993.
- [43] I. Llamas-Garro, M. J. Lancaster, and P. S. Hall, "A low-loss wideband suspended coaxial transmission line," *Microwave and Optical Technology Letters*, vol. 43, pp. 93-95, 2004.
- [44] K. Jiang, M. J. Lancaster, I. Llamas-Garro, and P. Jin, "SU-8 Ka-band filter and its microfabrication," *Journal of Micromechanics and Microengineering*, vol. 15, p. 1522, 2005.
- [45] A. Jaimes-Vera, I. Llamas-Garro, M. Ke, Y. Wang, M. J. Lancaster, and L. Pradell, "Polymer-based micromachined rectangular coaxial filters for millimeter-wave applications," *International Journal of Microwave and Wireless Technologies*, vol. 3, pp. 115-120, Apr 2011.
- [46] W. Y. Liu, D. Steenson, and M. B. Steer, "Membrane-supported CPW with mounted active devices," *IEEE Microwave and Wireless Components Letters*, vol. 11, pp. 167-169, 2001.
- [47] D. J. Young, V. Malba, J. J. Ou, A. F. Bernhardt, and B. E. Boser, "Monolithic high-performance three-dimensional coil inductors for wireless communication

- applications," in *Electron Devices Meeting, 1997. IEDM '97. Technical Digest., International*, 1997, pp. 67-70.
- [48] J. Y. Park and M. G. Allen, "High Q spiral-type microinductors on silicon substrates," *Magnetics, IEEE Transactions on*, vol. 35, pp. 3544-3546, 1999.
 - [49] G. W. Dahlmann and E. M. Yeatman, "High Q microwave inductors on silicon by surface tension self-assembly," *Electronics Letters*, vol. 36, pp. 1707-1708, 2000.
 - [50] N. Chomnawang and J.-B. Lee, "On-chip 3D air-core microinductor for high-frequency applications using deformation of sacrificial polymer," Newport Beach, CA, USA, 2001, pp. 54-62.
 - [51] C. Chung-Hui, F. Yean-Kuen, Y. Chih-Wei, and C. S. Tang, "A deep submicron CMOS process compatible suspending high-Q inductor," *Electron Device Letters, IEEE*, vol. 22, pp. 522-523, 2001.
 - [52] L. Guo, M. Yu, Z. Chen, H. He, and Y. Zhang, "High Q multilayer spiral inductor on silicon chip for 5~6 GHz," *Electron Device Letters, IEEE*, vol. 23, pp. 470-472, 2002.
 - [53] Y. Jun-Bo, C. Yun-Seok, K. Byeong-Il, E. Yunseong, and Y. Euisik, "CMOS-compatible surface-micromachined suspended-spiral inductors for multi-GHz silicon RF ICs," *Electron Device Letters, IEEE*, vol. 23, pp. 591-593, 2002.
 - [54] H. Xiao, K. J. Chen, and P. C. H. Chan, "Silicon-based high-Q inductors incorporating electroplated copper and low-K BCB dielectric," *Electron Device Letters, IEEE*, vol. 23, pp. 520-522, 2002.
 - [55] K. J. Chen, H. Wai Cheong, Z. Jinwen, and L. L. W. Leung, "CMOS-compatible micromachined edge-suspended spiral inductors with high Q-factors and self-resonance frequencies," *Electron Device Letters, IEEE*, vol. 25, pp. 363-365, 2004.
 - [56] W. Xi-Ning, Z. Xiao-Lin, Z. Yong, D. Xu-Han, and C. Bing-Chu, "Fabrication and performance of a novel suspended RF spiral inductor," *Electron Devices, IEEE Transactions on*, vol. 51, pp. 814-816, 2004.
 - [57] T. Hau-Yiu and J. Lau, "An on-chip vertical solenoid inductor design for multigigahertz CMOS RFIC," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, pp. 1883-1890, 2005.
 - [58] C. S. Lin, Y. K. Fang, S. F. Chen, C. Y. Lin, M. C. Hsieh, C. M. Lai, *et al.*, "A deep submicrometer CMOS process compatible high-Q air-gap solenoid inductor with laterally laid structure," *Electron Device Letters, IEEE*, vol. 26, pp. 160-162, 2005.
 - [59] L. Chen, L. Huailin, W. Chuan, Y. Jun, H. Ru, and W. Yangyuan, "High-Q Integrated Inductor Using Post-CMOS Selectively Grown Porous Silicon (SGPS) Technique for RFIC Applications," *Electron Device Letters, IEEE*, vol. 28, pp. 763-766, 2007.
 - [60] G. Lei and L. Xinxin, "High-Q Solenoid Inductors With a CMOS-Compatible Concave-Suspending MEMS Process," *Journal of Microelectromechanical Systems*, vol. 16, pp. 1162-1172, 2007.
 - [61] I. Llamas-Gano, K. Yongsung, B. Chang-Wook, and K. Yong-Kweon, "A micromachined monolithic half coaxial transmission line filter," in *Asia-Pacific Microwave Conference Proceedings (APMC)*, Suzhou, China, 2005, p. 4 pp.
 - [62] I. Llamas-Garro, K. Yongsung, B. Chang-Wook, and K. Yong-Kweon, "A Planar High-Q Micromachined Monolithic Half-Coaxial Transmission-Line Filter," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, pp. 4161-4168, 2006.

- [63] V. M. Hietala, Y. P. Kwon, and K. S. Champlin, "Low-loss slow-wave propagation along a microstructure transmission line on a silicon surface," *Electronics Letters*, vol. 22, pp. 755-756, 1986.
- [64] I. Jeong, S. Seong-Ho, G. Ju-Hyun, L. Joong-Soo, N. Choong-Mo, K. Dong-Wook, *et al.*, "High-performance air-gap transmission lines and inductors for millimeter-wave applications," *IEEE Transactions on Microwave Theory and Techniques*, vol. 50, pp. 2850-2855, 2002.
- [65] D. P. Newlin, A. V. H. Pham, and J. E. Harriss, "Development of low loss organic-micromachined interconnects on silicon at microwave frequencies," *IEEE Transactions on Components and Packaging Technologies*, vol. 25, pp. 506-510, 2002.
- [66] G. Lihui, Z. Yibin, and J. S. Y. Jie, "Transmission lines fabricated with CMOS interconnect technology," *Microelectronic Engineering*, vol. 72, pp. 223-229, 2004.
- [67] G. Six, G. Prigent, R. Eric, G. Dambrine, and H. Happy, "Fabrication and characterization of low-loss TFMS on silicon substrate up to 220 GHz," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, pp. 301-305, 2005.
- [68] T. S. D. Cheung and J. R. Long, "Shielded passive devices for silicon-based monolithic microwave and millimeter-wave integrated circuits," *Solid-State Circuits, IEEE Journal of*, vol. 41, pp. 1183-1200, 2006.
- [69] C. Hsiu-Ying, Y. Tzu-Jin, S. Liu, and W. Chung-Yu, "High-Performance Slow-Wave Transmission Lines With Optimized Slot-Type Floating Shields," *Electron Devices, IEEE Transactions on*, vol. 56, pp. 1705-1711, 2009.
- [70] J. A. Bishop, M. M. Hashemi, K. Kiziloglu, L. Larson, N. Dagli, and U. Mishra, "Monolithic coaxial transmission lines for mm-wave ICs," in *High Speed Semiconductor Devices and Circuits, 1991., Proceedings IEEE/Cornell Conference on Advanced Concepts in*, 1991, pp. 252-260.
- [71] R. Osorio, "Micromachined Transmission Lines for Microwave Applications," Ph.D, Albert-Ludwigs, Freiburg, 2003.
- [72] R. Osorio, M. J. K. Klein, H. Massler, and J. Korvink, "Micromachined Strip Line with SU-8 as the Dielectric," presented at the Proc. 11th Gallium Arsenide Application Symposium, Munich, Germany, 2003.
- [73] S. P. Natarajan, A. M. Hoff, and T. M. Weller, "Polyimide core 3D rectangular micro coaxial transmission lines," *Microwave and Optical Technology Letters*, vol. 52, pp. 1291-1293, 2010.
- [74] S. P. Natarajan, T. M. Wellerand, and A. M. Hoff, "Integrated Micro Coaxial Air-Lines with Perforations," in *IEEE MTT-S International Microwave Symposium Digest*, San Francisco, CA, USA, 2006, pp. 424-427.
- [75] Y. Jun-Bo, K. Byeong-Il, C. Yun-Seok, and Y. Euisik, "3-D lithography and metal surface micromachining for RF and microwave MEMS," in *The Fifteenth IEEE International Conference on Micro Electro Mechanical Systems*, 2002, pp. 673-676.
- [76] J.-B. Yoon, B.-I. Kim, Y.-S. Choi, and E. Yoon, "3-D construction of monolithic passive components for RF and microwave ICs using thick-metal surface micromachining technology," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, pp. 279-288, 2003.
- [77] P. Eun-Chul, C. Yun-Suk, K. Byeong-Il, Y. Jun-Bo, and Y. Euisik, "A low loss MEMS transmission line with shielded ground," in *The Sixteenth IEEE Annual International Conference on Micro Electro Mechanical Systems (MEMS)*, Kyoto, Japan, 2003, pp. 136-139.

- [78] E. R. Brown, A. L. Cohen, C. A. Bang, M. S. Lockard, B. W. Byrne, N. M. Vandelli, *et al.*, "Characteristics of microfabricated rectangular coax in the Ka band," *Microwave and Optical Technology Letters*, vol. 40, pp. 365-368, 2004.
- [79] J. R. Reid and R. T. Webster, "A compact integrated coaxial V-band bandpass filter," in *Antennas and Propagation Society International Symposium, 2004. IEEE*, 2004, pp. 990-993 Vol.1.
- [80] R. T. Chen, E. R. Brown, and R. S. Singh, "A compact 30 GHz low loss balanced hybrid coupler fabricated using micromachined integrated coax," in *Radio and Wireless Conference, 2004 IEEE*, 2004, pp. 227-230.
- [81] R. T. Chen, E. R. Brown, and C. A. Bang, "A compact low-loss Ka-band filter using 3-dimensional micromachined integrated coax," in *17th IEEE International Conference on Micro Electro Mechanical Systems (MEMS)*, 2004, pp. 801-804.
- [82] J. R. Reid, E. D. Marsh, and R. T. Webster, "Micromachined rectangular-coaxial transmission lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, pp. 3433-3442, 2006.
- [83] S. E. Alper, I. E. Ocak, and T. Akin, "Ultrathick and High-Aspect-Ratio Nickel Microgyroscope Using EFAB Multilayer Additive Electroforming," *Journal of Microelectromechanical Systems*, vol. 16, pp. 1025-1035, 2007.
- [84] D. W. Sherrer and J. J. Fisher, "Coaxial waveguide microstructures and methods of formation thereof," U.S. Patent, Dec. 30, 2004.
- [85] K. Vanhille, D. Filipovic, C. Nichols, D. Fontaine, W. Wilkins, E. Daniel, *et al.*, "Balanced Low-Loss Ka-Band u-Coaxial Hybrids," presented at the IEEE MTT-S International Microwave Symposium Digest, Honolulu, HI, USA, 2007.
- [86] K. Vanhille, M. Lukic, S. Rondineau, D. Filipovic, and Z. Popovic, "Integrated micro-coaxial passive components for millimeter-wave antenna front ends," presented at the The Fourth IASTED International Conference on Antennas, Radar and Wave Propagation, Montreal, Quebec, Canada, 2007.
- [87] Milan V. Lukic' and Dejan S. Filipovic, "Modeling of 3-D Surface Roughness Effects With Application to u-Coaxial Lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, 2007.
- [88] D. S. Filipovic, M. V. Lukic, L. Yongjae, and D. L. Fontaine, "Monolithic Rectangular Coaxial Lines and Resonators With Embedded Dielectric Support," *IEEE Microwave and Wireless Components Letters*, vol. 18, pp. 740-742, 2008.
- [89] Z. Popovic, K. Vanhille, N. Ehsan, E. Cullens, Y. Saito, J. M. Rollin, *et al.*, "Micro-fabricated micro-coaxial millimeter-wave components," in *33rd International Conference on Infrared, Millimeter and Terahertz Waves (IRMMW-THz)* 2008, pp. 1-3.
- [90] N. Ehsan, E. Cullens, K. Vanhille, D. Frey, S. Rondineau, R. Actis, *et al.*, "Micro-coaxial lines for active hybrid-monolithic circuits," in *IEEE MTT-S International Microwave Symposium Digest*, Boston, MA, USA, 2009, pp. 465-468.
- [91] Y. Saito, M. V. Lukic, D. Fontaine, J. M. Rollin, and D. S. Filipovic, "Monolithically Integrated Corporate-Fed Cavity-Backed Antennas," *IEEE Transactions on Antennas and Propagation*, vol. 57, pp. 2583-2590, 2009.
- [92] N. Ehsan, K. Vanhille, S. Rondineau, E. D. Cullens, and Z. B. Popovic, "Broadband Micro-Coaxial Wilkinson Dividers," *IEEE Transactions on Microwave Theory and Techniques*, vol. 57, pp. 2783-2789, 2009.
- [93] J. Mruk, H. Zhou, U. Manseok, Y. Saito, and D. Filipovic, "Wideband mm-wave log-periodic antennas," in *Antennas and Propagation, 2009. EuCAP 2009. 3rd European Conference on*, 2009, pp. 2584-2587.

- [94] Z. Popovic, "Micro-coaxial micro-fabricated feeds for phased array antennas," in *Phased Array Systems and Technology (ARRAY), 2010 IEEE International Symposium on*, 2010, pp. 1-10.
- [95] N. Ehsan, K. J. Vanhille, S. Rondineau, and Z. Popovic, "Micro-Coaxial Impedance Transformers," *IEEE Transactions on Microwave Theory and Techniques*, vol. 58, pp. 2908-2914, 2010.
- [96] Z. Hongyu, N. A. Sutton, and D. S. Filipovic, "W-band endfire log periodic dipole array," in *IEEE International Symposium on Antennas and Propagation (APSURSI)*, 2011, pp. 1233-1236.
- [97] E. D. Cullens, L. Ranzani, K. J. Vanhille, E. N. Grossman, N. Ehsan, and Z. Popovic, "Micro-Fabricated 130-180 GHz Frequency Scanning Waveguide Arrays," *IEEE Transactions on Antennas and Propagation*, vol. 60, pp. 3647-3653, 2012.
- [98] J. M. Oliver, J. M. Rollin, K. Vanhille, and S. Raman, "A W-Band Micromachined 3-D Cavity-Backed Patch Antenna Array With Integrated Diode Detector," *IEEE Transactions on Microwave Theory and Techniques*, vol. 60, pp. 284-292, 2012.
- [99] P. Ralston, M. Oliver, K. Vummidi, and S. Raman, "Liquid-Metal Vertical Interconnects for Flip Chip Assembly of GaAs C-Band Power Amplifiers Onto Micro-Rectangular Coaxial Transmission Lines," *Solid-State Circuits, IEEE Journal of*, vol. 47, pp. 2327-2334, 2012.
- [100] J. R. Reid, J. M. Oliver, K. Vanhille, and D. Sherrer, "Three dimensional metal micromachining: A disruptive technology for millimeter-wave filters," in *Silicon Monolithic Integrated Circuits in RF Systems (SiRF), 2012 IEEE 12th Topical Meeting on*, 2012, pp. 17-20.
- [101] N. A. Sutton, J. M. Oliver, and D. S. Filipovic, "Wideband 18-40 GHz Surface Micromachined Branchline Quadrature Hybrid," *IEEE Microwave and Wireless Components Letters*, vol. 22, pp. 462-464, 2012.
- [102] N. Chamberlain, M. Sanchez Barbetty, G. Sadowy, E. Long, and K. Vanhille, "A dual-polarized W-band metal patch antenna element for phased array applications," in *Antennas and Propagation Society International Symposium (APSURSI), 2014 IEEE*, 2014, pp. 1640-1641.
- [103] Y. Jun-Bo, K. Byeong-II, C. Yun-Seok, and E. Yoon, "3-D construction of monolithic passive components for RF and microwave ICs using thick-metal surface micromachining technology," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, pp. 279-288, 2003.
- [104] H. A. Wheeler, "Formulas for the Skin Effect," *Proceedings of the IRE*, vol. 30, pp. 412-424, 1942.
- [105] B. Todd, W. W. Flack, and S. White, "AZ4620 Resist Photolithography," presented at the Proceedings of SPIE MEMS: Micromachining and Microfabrication Process Technology V, Santa Clara, CA, 1999.
- [106] B. Magali and et al., "Thick photoresist development for the fabrication of high aspect ratio magnetic coils," *Journal of Micromechanics and Microengineering*, vol. 12, p. 444, 2002.
- [107] R. Chang, "AZ4620 Resist Photolithography," UCI Integrated Nanosystems Research Facility 2008.
- [108] A. E. Materials, "AZ®40XT Series, Chemically Amplified Positive Tone Thick Photoresist For Etch and Plating Applications."
- [109] A. E. Materials, "AZ®50XT Photoresist, Thick Positive Novolak Photoresist For Plating & Etch Applications."

- [110] A. E. Materials, "AZ® Exp 125nXT-10A Developed in AZ 300 MIF and AZ 303N."
- [111] K. MICROCHEM, "KMPR®1000 Chemically Amplified Negative Photoresist."
- [112] C. H. Lee and K. Jiang, "Fabrication of thick electroforming micro mould using a KMPR negative tone photoresist," *Journal of Micromechanics and Microengineering*, vol. 18, 21 April 2008.
- [113] S. Young-Min, D. Gamzina, L. R. Barnett, F. Yaghmaie, A. Baig, and N. C. Luhmann, "UV Lithography and Molding Fabrication of Ultrathick Micrometallic Structures Using a KMPR Photoresist," *Journal of Microelectromechanical Systems*, vol. 19, pp. 683-689, 2010.
- [114] N. Basak, G. L. Harris, J. Griffin, and K. D. Wise, "Characterisation and application of BPR-100 thick photoresist," *IET Micro & Nano Letters*, vol. 2, pp. 115-117, 2007.
- [115] R. HAAS, "BPR 100™ Photoresist, For Advanced Packaging Applications."
- [116] W. W. Flack, W. P. Fan, and S. White, "Optimization and characterization of ultrathick photoresist films," in *Proceedings of SPIE*, 1998, pp. 1288-1303.
- [117] R. Vempati Srinivasa, K. Vaidyanathan, Y. Seung Wook, and A. O. T. Andrew, "A thick photoresist process for advanced wafer level packaging applications using JSR THB-151N negative tone UV photoresist," *Journal of Micromechanics and Microengineering*, vol. 16, p. 1841, 2006.
- [118] P. Niedermann, H. Berthou, S. Zwickl, U. Schönholzer, K. Meier, C. Gantner, *et al.*, "A novel thick photoresist for microsystem technology," *Microelectronic Engineering*, vol. 67-68, pp. 259-265, 2003.
- [119] Y. Gonin, F. Munnik, F. Benninger, F. Dias, Mikha, x, *et al.*, "Comparison of a new photoresist (DiaPlate 133) with SU-8 using both x-ray and ion beam lithographies," *Journal of Vacuum Science & Technology B: Microelectronics and Nanometer Structures*, vol. 22, pp. 1982-1986, 2004.
- [120] M. Hideki, "Low Temperature Deposition of Silicon Nitride by the Catalytic Chemical Vapor Deposition Method," *Japanese Journal of Applied Physics*, vol. 28, p. 2157, 1989.
- [121] O. Shinya and M. Hideki, "Improved Properties of Silicon Nitride Films Prepared by the Catalytic Chemical Vapor Deposition Method," *Japanese Journal of Applied Physics*, vol. 36, p. 7035, 1997.
- [122] T. J. Sanders, E. L. Caraway, C. Hall, J. H. Linn, C. Adkins, H. Deju, *et al.*, "Silicon nitride deposition process for low cost microelectronics applications," in *Proceedings of the Twelfth Biennial University/Government/Industry Microelectronics Symposium (UGIM)*, Rochester, NY, USA, 1997, pp. 173-176.
- [123] C.-H. Chen, Y.-K. Fang, C.-S. Lin, C.-W. Yang, and J.-C. Hsieh, "A novel multi-level interconnect scheme with air as low K inter-metal dielectric for ultradeep submicron application," *Solid-State Electronics*, vol. 45, pp. 199-203, 1/1/ 2001.
- [124] L. Faraone and D. L. Patterson, "Process of producing thick layers of silicon dioxide," United States Patent, 1986.
- [125] F. Sloomman and P. Bouard, "PROCESS FOR PRODUCING A SILICON OXIDE DEPOSIT OF THE SURFACE OF A METTALIC OR METALLIZED POLYMER SUBSTRATE USING CORONA DISCHARGE AT PRESSURES UP TO APPROXIMATELY ATMOSPHERIC," FRANCE Patent, 1996.
- [126] C. Zhang and K. Najafi, "FABRICATION OF THICK SILICON DIOXIDE LAYERS USING DRIE, OXIDATION AND TRENCH REFILL," presented at the Proc. 15th IEEE Int. Conf. MEMS, 2002.

- [127] Z. Chunbo and N. Khalil, "Fabrication of thick silicon dioxide layers for thermal isolation," *Journal of Micromechanics and Microengineering*, vol. 14, p. 769, 2004.
- [128] N. N. TOAN, "Spin-On Glass Materials and Applications in Advanced IC Technologies," Universiteit Twente, Vietnam, 1970.
- [129] A. A. Hamzah, B. Y. Majlis, and I. Ahmad, "Formation of thick spin-on glass (SOG) sacrificial layer for capacitive accelerometer encapsulation," in *Proc. SPIE 6037*, Brisbane, Australia, 29 December 2005.
- [130] A. C. Madayag and Z. Zhiping, "Optimization of spin-on-glass process for multilevel metal interconnects," in *University/Government/Industry Microelectronics Symposium, 2001. Proceedings of the Fourteenth Biennial*, 2001, pp. 136-139.
- [131] P. H. Chen, T. Chen, and H. C. Shih, "Spin-on-glass (SOG) process-induced tungsten residue in a semiconductor manufacturing procedure," *Microelectronics Journal*, vol. 35, pp. 985-987, 12// 2004.
- [132] A. N. Al-Omari and K. L. Lear, "Dielectric characteristics of spin-coated dielectric films using on-wafer parallel-plate capacitors at microwave frequencies," *Dielectrics and Electrical Insulation, IEEE Transactions on*, vol. 12, pp. 1151-1161, 2005.
- [133] T. Shimoto, K. Matsui, and K. Utsumi, "Cu/photosensitive-BCB thin-film multilayer technology for high-performance multichip modules," *Components, Packaging, and Manufacturing Technology, Part B: Advanced Packaging, IEEE Transactions on*, vol. 18, pp. 18-22, 1995.
- [134] P. E. Garrou, R. H. Heistand, M. G. Dibbs, T. A. Manial, C. E. Mohler, T. M. Stokich, *et al.*, "Rapid thermal curing of BCB dielectric," *IEEE Transactions on Components, Hybrids, and Manufacturing Technology*, vol. 16, pp. 46-52, 1993.
- [135] M. Tagami, T. Fukai, M. Hiroi, J. Kawahara, K. Shiba, M. Tada, *et al.*, "Cu single damascene interconnects with plasma-polymerized organic polymers ($k=2.6$) for high-speed, 0.1 μm /m CMOS devices," in *Electron Devices Meeting, 2000. IEDM '00. Technical Digest. International*, 2000, pp. 851-853.
- [136] C. Qianwen, H. Cui, T. Zhimin, and W. Zheyao, "Low Capacitance Through-Silicon-Vias With Uniform Benzocyclobutene Insulation Layers," *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 3, pp. 724-731, 2013.
- [137] M. J. Lancaster, Z. Jiafeng, K. Maolong, W. Yi, and K. Jiang, "Design and High Performance of a Micromachined K -Band Rectangular Coaxial Cable," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, pp. 1548-1553, 2007.
- [138] R. Feng and R. J. Farris, "Influence of processing conditions on the thermal and mechanical properties of SU8 negative photoresist coatings," *Journal of Micromechanics and Microengineering*, vol. 13, pp. 80-88, 2003.
- [139] V. Brusic, R. Zhou, and C. Thompson, "Gold CMP Composition and Method," US Patent US 7368066 B2, 2008.
- [140] G. Karbasian, "Chemical Mechanical Polishing of Gold," Master of Science, Electrical Engineering, University of Notre Dame, Indiana, 2013.
- [141] D. Giancoli, "Electric Currents and Resistance," in *Physics for Scientists and Engineers with Modern Physics*, 4th Edition ed Upper Saddle River, New Jersey: Prentice Hall, 2009.
- [142] R. A. Serway, *Principles of Physics*, 2nd edition ed. Fort Worth, Texas; London: Saunders College Pub, 1998.

- [143] A. M. Helmenstine. (2014). *Table fo Electrical Resistivity and Conductivity*. Available: <http://chemistry.about.com/od/moleculescompounds/a/Table-Of-Electrical-Resistivity-And-Conductivity.htm>
- [144] *Technical data for Tantalum*. Available: <http://periodictable.com/Elements/073/data.html>
- [145] K. H. Lau, "Loss calculations for rectangular coaxial lines," *IEE Proceedings Microwaves, Antennas and Propagation*, vol. 135, pp. 207-209, 1988.
- [146] T. S. Chen, "Determination of the Capacitance, Inductance, and Characteristic Impedance of Rectangular Lines," *IRE Transactions on Microwave Theory and Techniques*, vol. 8, pp. 510-519, 1960.
- [147] A. Issaoun, X. Yong Zhong, S. Jinglin, J. Brinkhoff, and L. Fujiang, "On the Deembedding Issue of CMOS Multigigahertz Measurements," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, pp. 1813-1823, 2007.
- [148] J. Martens, "Common adapter/fixture extraction techniques: sensitivities to calibration anomalies," in *74th ARFTG Microwave Measurement Symposium*, 2009, pp. 1-10.
- [149] R. ma, H. Guorui, C. Xinwei, and Z. Wenmei, "Calibrating an Arbitrary Test Fixture for a Symmetric Device by Three Measurements," *IEEE Transactions on Instrumentation and Measurement*, vol. 59, pp. 145-152, 2010.
- [150] A. M. Mangan, S. P. Voinigescu, Y. Ming-Ta, and M. Tazlauanu, "De-embedding transmission line measurements for accurate modeling of IC designs," *Electron Devices, IEEE Transactions on*, vol. 53, pp. 235-241, 2006.
- [151] Y. Tian, K. Lee, and H. Wang, "Air-gapped microcoaxial transmission line for ultrawide band microwave and millimeter wave ICS," *Microwave and Optical Technology Letters*, vol. 56, pp. 1462-1465, 2014.
- [152] Y. Saito and D. S. Filipovic, "Analysis and Design of Monolithic Rectangular Coaxial Lines for Minimum Coupling," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, pp. 2521-2530, 2007.
- [153] N. P. Pham, P. M. Sarro, K. T. Ng, and J. N. Burghartz, "IC-Compatible Two-level Bulk Micromachining for RF Silicon Technology," in *Solid-State Device Research Conference, 2000. Proceeding of the 30th European*, 2000, pp. 204-207.
- [154] K. Takahashi, U. Sangawa, S. Fujita, M. Matsuo, T. Urabe, H. Ogura, *et al.*, "Packaging using microelectromechanical technologies and planar components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 49, pp. 2099-2104, 2001.
- [155] C. Microtech. Available: http://ekv.epfl.ch/files/content/sites/ekv/files/mos-ak/erfurt/INFINITY_QA_1002c.pdf
- [156] T. Kamei, Y. Utsumi, N. Q. Dinh, and N. Thanh, "Wide-band coaxial-to-coplanar transition," *IEICE Transactions on Electronics*, vol. E90C, pp. 2030-2036, Oct 2007.
- [157] G. Askari, H. Fadakar, H. Mirmohammad-Sadeghi, and A. Electromagnet, "Analysis, Design and Implementation of a Useful Broadband Coaxial-to-microstrip Transition," presented at the Progress in Electromagnetics Research Symposium (PIERS), Moscow, Russia, 2012.
- [158] J. Boone, S. Krishnan, and S. Bhansali, "Silicon based vertical micro-coaxial transition for high frequency packaging technologies," *Progress In Electromagnetics Research B*, vol. 50, pp. 1-17, 2013.

- [159] Y. Tian, K. Lee, and H. Wang, "A 390 ps On-Wafer True-Time-Delay Line Developed by a Novel Micro-Coax Technology," *IEEE Microwave and Wireless Components Letters*, vol. 24, pp. 233-235, 2014.
- [160] K. J. Vanhille, "Design and Characterization of Microfabricated Three Dimensional Millimeter-Wave Components," Doctor of Philosophy, Department of Electrical and Computer Engineering, University of Colorado, University of Colorado, 2007.
- [161] N. Ehsan, "Broadband Microwave Lithographic 3D Components," Doctor of Philosophy, Department of Electrical, Computer, and Energy Engineering, University of Colorado, University of Colorado, 2010.
- [162] G. Rongzhi, H. Lei, and Z. Jue, "Experimental studies on bonding pressure in wire bonding," in *High Density Microsystem Design and Packaging and Component Failure Analysis, 2006. HDP'06. Conference on*, 2006, pp. 125-129.
- [163] A. Group. *Hot Bar Reflow Soldering Fundamentals*. Available: http://www.amadamiyachi.com/servlet/servlet.FileDownload?retURL=%2Fapex%2Feducationalresources_fundamentals&file=01580000001d0TZ
- [164] R. Horton, "The Electrical Characterization of a Right-Angled Bend in Microstrip Line (Short Papers)," *IEEE Transactions on Microwave Theory and Techniques*, vol. 21, pp. 427-429, 1973.
- [165] R. J. P. Douville and D. S. James, "Experimental Study of Symmetric Microstrip Bends and Their Compensation," *IEEE Transactions on Microwave Theory and Techniques*, vol. 26, pp. 175-182, 1978.
- [166] M. D. Wu, S. M. Deng, R. B. Wu, and P. Hsu, "Full-wave characterization of the mode conversion in a coplanar waveguide right-angled bend," *IEEE Transactions on Microwave Theory and Techniques*, vol. 43, pp. 2532-2538, 1995.
- [167] D. Jaisson, "Coplanar waveguide bend with radial compensation," *IEE Proceedings Microwaves, Antennas and Propagation*, vol. 143, pp. 447-450, 1996.
- [168] M. I. Montrose, "Time and frequency domain analysis for right angle corners on printed circuit board traces," in *Electromagnetic Compatibility, 1998. 1998 IEEE International Symposium on*, 1998, pp. 551-556 vol.1.
- [169] D. Brooks. (1998, Jan.) 90° Corners, the Final Turn. *Printed Circuit Design*. 551-556.
- [170] P. M. Watson and K. C. Gupta, "EM-ANN models for microstrip vias and interconnects in dataset circuits," *IEEE Transactions on Microwave Theory and Techniques*, vol. 44, pp. 2495-2503, 1996.
- [171] P. M. Watson and K. C. Gupta, "Design and optimization of CPW circuits using EM-ANN models for CPW components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 45, pp. 2515-2523, 1997.
- [172] X. Li, Y. Li, and J. Zhao, "ANN-Based Pad Modeling Technique for MOSFET Devices," *Progress In Electromagnetics Research B*, vol. 118, pp. 303-319, 2001.
- [173] N. Li, X. Li, and S. G. Quan, "An ANN-based Small-signal Equivalent Circuit Model for MOSFET Device," *Progress In Electromagnetics Research B*, vol. 122, pp. 47-60, 2012.
- [174] Z. Qi-Jun, K. C. Gupta, and V. K. Devabhaktuni, "Artificial neural networks for RF and microwave design - from theory to practice," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, pp. 1339-1350, 2003.
- [175] P. I. Somlo, "Some Aspects of the Measurement of the Q Factor of Transmission Lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 11, pp. 472-478, 1963.

- [176] S. Barker and G. M. Rebeiz, "Distributed MEMS true-time delay phase shifters and wide-band switches," *IEEE Transactions on Microwave Theory and Techniques*, vol. 46, pp. 1881-1890, 1998.
- [177] J. B. Hacker, R. E. Mihailovich, K. Moonil, and J. F. DeNatale, "A Ka-band 3-bit RF MEMS true-time-delay network," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, pp. 305-308, 2003.
- [178] C. D. Nordquist, W. Dyck, M. Kraus, I. C. Reines, L. Goldsmith, D. Cowan, *et al.*, "A DC to 10-GHz 6-b RF MEMS time delay circuit," *IEEE Microwave and Wireless Components Letters*, vol. 16, pp. 305-307, 2006.
- [179] R. Buttiglione, S. Catoni, G. De Angelis, M. Dispenza, A. Fiorello, K. Kautio, *et al.*, "Alumina and LTCC Technology for RF MEMS Switches and True Time Delay Lines," in *European Microwave Integrated Circuit Conference (EuMIC)*, Amsterdam, Netherlands, 2008, pp. 366-369.
- [180] D. Kuylensstierna, A. Vorobiev, P. Linner, and S. Gevorgian, "Ultrawide-band tunable true-time delay lines using ferroelectric varactors," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, pp. 2164-2170, 2005.
- [181] W. Yi, S. Hieng Tiong, H. Frederick, and M. J. Lancaster, "Wide-band superconducting coplanar delay lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, pp. 2348-2354, 2005.
- [182] J. Zhang, S. W. Cheung, and T. I. Yuk, "Compact composite right/left-handed transmission line unit cell for the design of true-time-delay lines," *IET Microwaves, Antennas & Propagation*, vol. 6, pp. 893-898, 2012.
- [183] L. Xing, M. Kintis, C. Hansen, W. Chan, G. Tseng, and M. Tan, "A Simple DC to 110 GHz MMIC True Time Delay Line," *IEEE Microwave and Wireless Components Letters*, vol. 22, pp. 369-371, 2012.
- [184] H. Y. Lim, G. I. Ng, and Y. C. Leong, "Compact true time delay line with partially shielded coplanar waveguide transmission lines," in *IEEE International Symposium on Radio-Frequency Integration Technology (RFIT)*, Singapore, 2012, pp. 62-64.
- [185] Y. Tian, K. Y. Lee, and H. Wang, "An On-Wafer Cu-based Microcoaxial Technology for Wideband and Low Loss Microwave and Mm-wave Applications," presented at the Asia-Pacific International Symposium on Electromagnetic Compatibility (APEMC), Taipei, 2015.
- [186] M. K. Chirala and N. Cam, "Multilayer Design Techniques for Extremely Miniaturized CMOS Microwave and Millimeter-Wave Distributed Passive Circuits," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, pp. 4218-4224, 2006.
- [187] M. K. Chirala and B. A. Floyd, "Millimeter-Wave Lange and Ring-Hybrid Couplers in a Silicon Technology for E-Band Applications," in *IEEE MTT-S International Microwave Symposium Digest*, San Francisco, CA, USA, 2006, pp. 1547-1550.
- [188] S. C. Tseng, C. c. Meng, C. H. Chang, S. H. Chang, and G. W. Huang, "A Silicon Monolithic Phase-Inverter Rat-Race Coupler Using Spiral Coplanar Striplines and Its Application in a Broadband Gilbert Mixer," *IEEE Transactions on Microwave Theory and Techniques*, vol. 56, pp. 1879-1888, 2008.
- [189] H. J. Wei, C. c. Meng, S. W. Yu, and C. H. Chang, "A Chebyshev-Response Step-Impedance Phase-Inverter Rat-Race Coupler Directly on Lossy Silicon Substrate and Its Gilbert Mixer Application," *IEEE Transactions on Microwave Theory and Techniques*, vol. 59, pp. 882-893, 2011.

- [190] Y. C. Tseng and T. G. Ma, "On-chip miniaturised wideband rat-race coupler using integrated passive device technology," *IET Microwaves, Antennas & Propagation*, vol. 8, pp. 959-964, 2014.
- [191] H. Kamitsuna, "A very small, low-loss MMIC rat-race hybrid using elevated coplanar waveguides," *IEEE Microwave and Guided Wave Letters*, vol. 2, pp. 337-339, 1992.
- [192] C. Y. Ng, M. Chongcheawchamnan, and I. D. Robertson, "Lumped-distributed hybrids in 3D-MMIC technology," *IEE Proceedings Microwaves, Antennas and Propagation*, vol. 151, pp. 370-374, 2004.
- [193] I. Haroun, Y. C. Hsu, and D. C. Chang, "60-GHz rat-race coupler using LG-CPW transmission lines in IPD technology," in *International Topical Meeting on & Microwave Photonics Conference (MWP)*, Singapore, 2011, pp. 284-287.
- [194] W. Sen and C. K. C. Tzuang, "Compacted Ka-Band CMOS Rat-Race Hybrid Using Synthesized Transmission Lines," in *IEEE MTT-S International Microwave Symposium Digest*, Honolulu, HI, USA, 2007, pp. 1023-1026.
- [195] H. Y. Ding, K. Lam, G. Wang, and W. H. Woods, "On-chip millimeter wave Rat-race Hybrid and Marchand Balun in IBM 0.13um BiCMOS technology," in *Asia-Pacific Microwave Conference (APMC)*, Hong Kong, 2008, pp. 1-4.
- [196] X. Ren, K. Song, B. Hu, and Q. Chen, "Compact filtering power divider with good frequency selectivity and wide stopband based on composite right-/left-handed transmission lines," *Microwave and Optical Technology Letters*, vol. 56, pp. 2122-2125, 2014.
- [197] M. C. Scardelletti, G. E. Ponchak, and T. M. Weller, "Miniaturized Wilkinson power dividers utilizing capacitive loading," *IEEE Microwave and Wireless Components Letters*, vol. 12, pp. 6-8, 2002.
- [198] X. Wang, I. Sakagami, Z. Ma, A. Mase, M. Yoshikawa, and M. Ichimura, "Miniaturized Dual-Band Wilkinson Power Divider With Self-Compensation Structure," *Components, Packaging and Manufacturing Technology, IEEE Transactions on*, vol. 5, pp. 389-397, 2015.
- [199] U. T. Ahmed and A. M. Abbosh, "Modified Wilkinson power divider using coupled microstrip lines and shunt open-ended stubs," *Electronics Letters*, vol. 51, pp. 838-839, 2015.
- [200] U. T. Ahmed and A. M. Abbosh, "Extremely wideband in-phase power divider using modified Wilkinson design," *Microwave and Optical Technology Letters*, vol. 57, pp. 1799-1802, 2015.
- [201] X. Wang, Z. Ma, I. Sakagami, A. Mase, and M. Yoshikawa, "Miniaturized Wilkinson power divider with extra coupled line section for compensation," *Microwave and Optical Technology Letters*, vol. 57, pp. 1869-1873, 2015.
- [202] K. Kyoungwoon and N. Cam, "An Ultra-Wideband Low-Loss Millimeter-Wave Slow-Wave Wilkinson Power Divider on 0.18 μm SiGe BiCMOS Process," *IEEE Microwave and Wireless Components Letters*, vol. 25, pp. 331-333, 2015.
- [203] T. Xinyi and K. Mouthaan, "Analysis and design of compact two-way Wilkinson power dividers using coupled lines," in *Asia-Pacific Microwave Conference (APMC)* Singapore, 2009, pp. 1319-1322.
- [204] W. Xiaolong, I. Sakagami, A. Mase, and M. Ichimura, "Wilkinson Power Divider With Complex Isolation Component and Its Miniaturization," *IEEE Transactions on Microwave Theory and Techniques*, vol. 62, pp. 422-430, 2014.
- [205] X. Wang, Z. Ma, and M. Yoshikawa, "Coupled microstrip line Wilkinson power divider with compensating capacitor," *Microwave and Optical Technology Letters*, vol. 57, pp. 328-332, 2015.

- [206] T. Hirota and M. Muraguchi, "K-band frequency up-converters using reduced-size couplers and dividers," in *Gallium Arsenide Integrated Circuit (GaAs IC) Symposium, 1991. Technical Digest 1991., 13th Annual, 1991*, pp. 53-56.
- [207] K. Rawat and F. M. Ghannouchi, "Design of reduced size power divider for lower RF band using periodically loaded slow wave structure," in *IEEE MTT-S International Microwave Symposium Digest*, Boston, MA, USA, 2009, pp. 613-616.
- [208] G. Xun, A. Margomenos, L. Bosui, S. Hajela, L. P. B. Katehi, and W. J. Chappell, "Precision fabrication techniques and analysis on high-Q evanescent-mode resonators and filters of different geometries," *IEEE Transactions on Microwave Theory and Techniques*, vol. 52, pp. 2557-2566, 2004.
- [209] L. Rigaudeau, P. Ferrand, D. Baillargeat, S. Bila, S. Verdeyme, M. Lahti, *et al.*, "LTCC 3-D resonators applied to the design of very compact filters for Q-band applications," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, pp. 2620-2627, 2006.
- [210] K. J. Vanhille, D. L. Fontaine, C. Nichols, D. S. Filipovic, and Z. Popovic, "Quasi-planar high-Q millimeter-wave resonators," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, pp. 2439-2446, 2006.
- [211] K. J. Vanhille, D. L. Fontaine, C. Nichols, Z. Popovic, and D. S. Filipovic, "Ka-Band Miniaturized Quasi-Planar High-Q Resonators," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, pp. 1272-1279, 2007.
- [212] M. Miao and C. Nguyen, "A Novel Multilayer Aperture-Coupled Cavity Resonator for Millimeter-Wave CMOS RFICs," *IEEE Transactions on Microwave Theory and Techniques*, vol. 55, pp. 783-787, 2007.
- [213] E. A. Alos, A. U. Zaman, and P. Kildal, "Ka-Band Gap Waveguide Coupled-Resonator Filter for Radio Link Diplexer Application," *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 3, pp. 870-879, 2013.
- [214] Z. Wang, R. Xu, and B. Yan, "A covering Ka-band two-way switch filter module using a three-line and an E-plane waveguide band-pass filters," *International Journal of RF and Microwave Computer-Aided Engineering*, vol. 25, pp. 305-310, 2015.
- [215] R. Levy, "Filters with Single Transmission Zeros at Real or Imaginary Frequencies," *IEEE Transactions on Microwave Theory and Techniques*, vol. 24, pp. 172-181, 1976.
- [216] J. R. Mruk and D. S. Filipovic, "Micro-coaxial V-/W-band filters and contiguous diplexers," *IET Microwaves, Antennas & Propagation*, vol. 6, pp. 1142-1148, 2012.
- [217] K. J. Herrick, Y. Jong-Gwan, and L. P. B. Katehi, "Microtechnology in the development of three-dimensional circuits," *IEEE Transactions on Microwave Theory and Techniques*, vol. 46, pp. 1832-1844, 1998.
- [218] Y. Tian, Q. Meng, K. Lee, and H. Wang, "Low-Loss Microcoax-to-CPW Transition for Air-gapped Microcoaxial Passives," *IEEE Microwave and Wireless Components Letters*, vol. PP, pp. 1-1, 2015.

Page intentionally left blank