In compliance with the
Canadian Privacy Legislation
some supporting forms
may have been removed from
this dissertation.

While these forms may be included
in the document page count,
their removal does not represent
any loss of content from the dissertation.

**Millimetre-Wave Ferrite-Loaded Laminated
Waveguide Phase Shifters Embedded in
Low Temperature Cofired Ceramic**

by

Joey R. Bray B.A.Sc, M.A.Sc.

A thesis submitted to the Faculty of Graduate Studies and Research
in partial fulfilment of the requirements for the degree of

Doctor of Philosophy
in Electrical Engineering

Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
Carleton University
Ottawa, Ontario, Canada.

September 2003

Copyright © 2003, Joey R. Bray

National Library
of Canada

Bibliothèque nationale
du Canada

Acquisitions and
Bibliographic Services

Acquisitiions et
services bibliographiques

395 Wellington Street
Ottawa ON K1A 0N4
Canada

395, rue Wellington
Ottawa ON K1A 0N4
Canada

Your file Votre référence

ISBN: 0-612-88710-3

Our file Notre référence

ISBN: 0-612-88710-3

The author has granted a non-exclusive licence allowing the National Library of Canada to reproduce, loan, distribute or sell copies of this thesis in microform, paper or electronic formats.

L'auteur a accordé une licence non exclusive permettant à la Bibliothèque nationale du Canada de reproduire, prêter, distribuer ou vendre des copies de cette thèse sous la forme de microfiche/film, de reproduction sur papier ou sur format électronique.

The author retains ownership of the copyright in this thesis. Neither the thesis nor substantial extracts from it may be printed or otherwise reproduced without the author's permission.

L'auteur conserve la propriété du droit d'auteur qui protège cette thèse. Ni la thèse ni des extraits substantiels de celle-ci ne doivent être imprimés ou autrement reproduits sans son autorisation.

Canada

The undersigned recommend to the Faculty of Graduate Studies and Research

Acceptance of this thesis

“Millimetre-Wave Ferrite-Loaded Laminated Waveguide Phase Shifters Embedded in Low Temperature Cofired Ceramic”

submitted by Joey Bray (M.A.Sc.) in partial fulfillment of
the requirements for the degree of Doctor of Philosophy in Engineering

Chairman, Department of Electronics
Professor Langis Roy

Thesis Supervisor
Professor Langis Roy

External Examiner
Dr. Ke Wu

Carleton University

August 28, 2003

Abstract

Phased array antennas, capable of forming multiple agile beams of radiation, are being deployed at ever-increasing operating frequencies. Systems operating at frequencies of 30 GHz and above, a range known as millimetre-waves, include military radar, commercial collision-avoidance radar, inter-satellite communication links, and future wideband mobile internet satellite communications services. Millimetre-wave phase shifters, the topic of this thesis, are key components of many phased array antennas.

A phase shifter refers to any device that can dynamically alter the phase, or the timing, of a signal. Although a number of concurrent millimetre-wave phase shifter designs presently exist, none simultaneously satisfies all of the desired requirements, including: light weight, small size, low cost, and good electrical performance. Phase shifters may be broadly categorised based on the material they use, either semiconductor or ferrite. Semiconductor shifters satisfy the former two requirements but suffer from high electrical losses. Conversely, traditional ferrite phase shifters have excellent electrical performance but are bulky and difficult to integrate with other components.

Given the superior electrical performance of ferrite phase shifters, a more compact means of fabricating these is proposed in this thesis. In addition, new types of ferrite phase shifters are devised, and these are subsequently made from a chip-packaging technology called low temperature cofired ceramic. The resulting phase shifters are easily fabricated, are simple to interconnect to other integrated circuits, are more compact and lightweight than their predecessors, and have excellent electrical performance at millimetre-wave frequencies. Prototype phase shifters have been successfully

manufactured using an experimental ferrite ceramic material and have yielded a maximum phase shift of 53° at 36 GHz for bias currents of -500 to $+500$ mA.

Acknowledgements

May I express my sincerest gratitude to those who made this research opportunity possible, Dr. Pierre Berini, Dr. Harri Kopola, and Dr. Langis Roy, through whose discussions a fruitful research partnership was established between Carleton University and Valtion Teknillinen Tutkimuskeskus (VTT), the national research centre of Finland. I further wish to thank Mr. Kari Kautio and the entire Manufacturing Technology group at VTT Electronics, Oulu Finland, for their expertise in fabricating the LTCC components that are described in this thesis.

I gratefully acknowledge the Natural Sciences and Engineering Research Council of Canada (NSERC), La fondation Alma et Baxter Ricard, EMS Technologies, VTT Electronics, the IEEE Microwave Theory and Techniques Society, the Government of Ontario, the University of Ottawa, Carleton University, and Dr. L. Roy for providing the financial support for this endeavour.

I would also like to acknowledge Dr. Malcolm Stubbs and Ms. Carole Glaser at the Communications Research Centre, Canada (CRC) for their assistance in performing the circuit measurements.

Finally, I would like to sincerely thank my wife and family for their unwavering support and patience throughout the years.

Dedicated to my wife, Elaine.

Table of Contents

List of Tables	x
List of Figures	xii
List of Appendices	xxvii
List of Acronyms and Abbreviations	xxviii
Glossary	xxx
 1 Introduction	 1
1.1 Motivation.....	1
1.2 Objectives.....	3
1.3 Thesis organisation.....	5
1.4 Thesis contributions	7
 2 Introduction to millimetre-wave phase shifters and to LTCC packaging technology	 9
2.1 Terminology.....	10
2.1.1 Phase definition and convention.....	10
2.1.2 Constant phase and true time delay phase shifters.....	11
2.1.3 Phase errors and figures of merit.....	12
2.2 A review of semiconductor phase shifter devices.....	15
2.2.1 Varactor diodes for analogue phase shifters.....	15
2.2.2 PIN diode switches for digital phase shifters.....	18
2.2.3 Transistor switches.....	21
2.2.4 Microelectromechanical system (MEMS) switches and varactors.....	25
2.3 Semiconductor phase shifters.....	26
2.3.1 Switched line phase shifters.....	27
2.3.2 Switched filter phase shifters.....	31
2.3.3 Loaded line phase shifters.....	33
2.3.4 Reflection phase shifters.....	38
2.3.4.1 Analogue reflection phase shifters.....	39
2.3.4.2 Digital reflection phase shifters.....	41
2.3.5 Combination phase shifters.....	42
2.3.6 Vector modulators as phase shifters.....	43
2.3.7 Semiconductor phase shifter summary.....	46
2.4 Ferrite material and ferrite millimetre-wave phase shifters.....	47
2.4.1 Ferrimagnetism, gyrotropic interaction, and ferrite preparation.....	48
2.4.2 Dual-mode ferrite phase shifters.....	50
2.4.3 Ferrite slab-loaded rectangular waveguide phase shifters.....	52
2.4.4 Reggia-Spencer and other suppressed rotation phase shifters.....	54
2.4.5 Planar ferrite phase shifters.....	56
2.4.6 Ferrite phase shifter summary.....	57

2.5	An introduction to multi-chip module packaging technologies and to low temperature cofired ceramic.....	58
2.6	State of the art and research goal.....	62
2.7	Chapter Two references.....	63
3	Characterisation of LTCC materials and transmission media at microwave frequencies.....	74
3.1	Microstrip characterisation techniques.....	75
3.1.1	Absolute Magnitude method for the loss characterisation of microstrip.....	75
3.1.2	Line Comparison method for the loss characterisation of microstrip.....	80
3.1.3	Thru-Reflect-Line (TRL) calibration method for characterising microstrip.....	81
3.1.4	Microwave characterisation using microstrip ring resonators.....	85
3.1.4.1	Analytical behaviour of the microstrip ring resonator.....	87
3.1.4.2	Characterisation method using the microstrip ring.....	91
3.1.4.3	Physical circuits and loss characterisation.....	93
3.1.4.4	Lumped element model capacitance evaluation of the gap-coupled ring.....	96
3.1.4.5	Resonant frequency correction.....	101
3.1.4.6	Microstrip ring lumped element model resonant frequency prediction.....	105
3.2	Laminated waveguide characterisation techniques.....	106
3.2.1	LWG propagation characterisation using the grid impedance method.....	107
3.2.2	LWG propagation characterisation using numerical methods.....	111
3.2.3	LWG propagation characterisation using a commercial solver.....	112
3.2.4	Resonant frequencies of TE _{10δ} modes in LWG cavities.....	113
3.2.5	LWG attenuation constant and substrate loss characterisation.....	119
3.3	Summary.....	132
3.4	Chapter Three references.....	134
4	Characterisation of the experimental Ferro ferrite LTCC tape.....	136
4.1	Ferrite tape processing.....	138
4.2	Measurement of the ferrite hysteresis curve.....	140
4.2.1	Theoretical operation of the transformer for hysteresis measurements.....	141
4.2.2	Theoretical transformer design in an LTCC environment.....	143
4.2.3	Practical solenoid design considerations.....	145
4.2.4	Measured ferrite permeability and hysteresis using the LTCC solenoidal transformer.....	151
4.2.5	Magnetostatic characterisation summary.....	157
4.3	Microwave characterisation of the Ferro ferrite.....	159
4.3.1	Characterisation of the ferrite permittivity, permeability, and loss tangent using ring resonators.....	159
4.3.1.1	Input impedance of the line-connected ring resonator.....	160

4.3.1.2	Resonance analysis of the line-connected ring	166
4.3.1.3	Microstrip line-connected ring design	169
4.3.1.4	Microstrip line-connected ring measurements	170
4.3.2	Characterisation of the ferrite permittivity, permeability, and loss tangent using LWG cavities	175
4.4	Summary and conclusions of the Ferro ferrite characterisation	179
4.5	Chapter Four references	181
5	LTCC rectangular waveguide phase shifters using via posts made of ferrite paste	183
5.1	Wet Stack multilayer ferrite paste and its compatibility with Heraeus CT2000 LTCC	185
5.2	Integration of ferrite paste and laminated waveguide for the design of phase shifters	189
5.3	Two-post ferrite-loaded RWG field derivation and phase shift calculation	193
5.4	Four-post ferrite-loaded RWG field derivation and phase shift calculation	197
5.5	The ferrite filling factor	210
5.6	Figure of merit: magnetic field circular polarisation axial ratio (CPAR)	214
5.7	Magnetostatic bias	217
5.8	The fabricated two-post ferrite-loaded laminated waveguide prototype	224
5.9	Summary	230
5.10	Chapter Five references	233
6	The ferrite-filled, antisymmetrically-biased (Ferrite-FAB) rectangular waveguide phase shifter	234
6.1	Definition of phase shift in a nonreciprocal medium	235
6.2	Observations of ferrite dual-slab rectangular waveguide phase shifters	238
6.3	Behaviour of the dual-slab phase shifter in a dielectric-filled waveguide	241
6.4	Ferrite-FAB field derivation	242
6.5	Ferrite-FAB prototype design	251
6.6	Prototype measurements	255
6.7	Validation of the observed phase shift	260
6.8	Summary	265
6.9	Chapter Six references	267
7	Conclusions, contributions, and future work	268
7.1	Conclusions	268
7.1.1	Millimetre-wave phase shifter state of the art and the potential role for LTCC	268
7.1.2	Millimetre-wave characterisation of LTCC materials	270
7.1.3	Ferrite-loaded waveguide phase shifters in LTCC	272
7.2	Thesis contributions	274
7.3	Future Work	278
8	Appendices	281

List of Tables

Table 2.1	A comparison between PIN and FET switches up to 20 GHz.....	24
Table 2.2	A comparison between PIN and FET switches in V-Band.....	24
Table 2.3	Switched line phase shifter performance summary.....	29
Table 2.4	Switched filter phase shifter performance.....	32
Table 2.5	Performance of single-bit loaded line phase shifters.....	36
Table 2.6	Performance of periodically loaded line phase shifters.....	37
Table 2.7	Performance of an analogue reflection phase shifter.....	40
Table 2.8	Performance of digital reflection phase shifters.....	41
Table 2.9	Performance of combination phase shifters.....	43
Table 2.10	Performance of IQ vector modulators.....	46
Table 2.11	Performance of a 60 GHz dual-mode ferrite waveguide phase shifter....	52
Table 2.12	Toroid dimensions for millimetre-wave ferrite slab-loaded waveguide phase shifters.....	53
Table 2.13	Performance of ferrite slab-loaded rectangular waveguide phase shifters.....	54
Table 2.14	Performance of ferrite dielectric waveguide suppressed rotation phase shifters.....	56
Table 2.15	Selected features of various multi-chip module technologies.....	59
Table 3.1	Parameters of the Thru and Line calibration standards for the Heraeus and Dupont LTCC material systems.....	82
Table 3.2	Measured dimensions of the three microstrip ring resonators.....	93
Table 3.3	Frequency correction and permittivity of the Ferro1 ring resonator.....	104

Table 3.4	Frequency correction and permittivity of the Dupont 943 ring resonator.....	105
Table 3.5	Circuit parameters of twelve LWG cavity resonators.....	115
Table 3.6	LWG circuit parameters for radiation loss simulations.....	125
Table 3.7	Resonant cavity quality factor analysis results using the parameters of Table 3.6 and a guide length of $d = 5.7$ mm.....	127
Table 3.8	Dupont 943 LWG cavities and their resonant frequencies.....	129
Table 4.1	Component values and expected gain levels.....	148
Table 4.2	LTCC solenoid transformer statistics.....	149
Table 5.1	Selected material properties of CT2000 dielectric tape, CT2000 pastes, and Wet Stack IP9xxx ferrites.....	186
Table 5.2	Material parameters for a dielectric-filled, two-post ferrite-loaded RWG.....	195
Table 5.3	Results of the DC bias conductivity tests for the Heraeus phase shifters.....	229

List of Figures

Fig. 2.1	Absolute phase of a true time delay phase shifter for nine different states...	12
Fig. 2.2	Model of a reverse-biased PN or Schottky junction.....	16
Fig. 2.3	Gallium arsenide integrated Schottky varactor diodes: (a) mesa type, (b) planar type.....	16
Fig. 2.4	PIN diode schematic and equivalent circuit.....	18
Fig. 2.5	PIN diode construction: (a) vertical mesa, (b) planar.....	19
Fig. 2.6	Unbiased switching FET conduction curves and equivalent circuits: (a) conduction curves, (b) high-impedance state, (c) low impedance state...	22
Fig. 2.7	Microelectromechanical system (MEMS) transducer configurations: (a) cantilever switch, (b) bridge varactor.....	25
Fig. 2.8	Switched line phase shifter.....	27
Fig. 2.9	Switched line phase shifter configurations: (a) series, (b) shunt.....	28
Fig. 2.10	A switched filter (low pass/high pass) phase shifter.....	31
Fig. 2.11	Shunt loaded line phase shifter.....	33
Fig. 2.12	Phasor diagrams of s_{21} showing the three classes of operation of a loaded line phase shifter.....	34
Fig. 2.13	Two-port reflection phase shifters: (a) using a circulator, (b) using a quadrature hybrid coupler.....	38
Fig. 2.14	A tandem attenuator/phase shifter vector modulator.....	44
Fig. 2.15	Block diagram of the IQ vector modulator.....	44
Fig. 2.16	Model of (a) a spinning electron and its magnetic moment, (b) when exposed to an external magnetic bias field (dashed ellipse is the precession motion).....	48

Fig. 2.17	Structure of a dual-mode waveguide ferrite phase shifter: (a) longitudinal cross-section, (b) quadrupole magnetostatic transverse bias in the polarisers. Areas in grey are made of ferrite material.....	51
Fig. 2.18	E-plane cross sections of ferrite slab-loaded rectangular waveguide phase shifters: (a) dual-slab, (b) dielectric loaded, (c) dual toroid.....	53
Fig. 2.19	Suppressed rotation ferrite phase shifters: (a) Reggia-Spencer, (b) ferrite dielectric waveguide. The $\otimes$ symbolises longitudinal magnetostatic bias into the page.....	55
Fig. 2.20	Microscope photograph of an emulsion mask on a stainless steel screen.....	60
Fig. 3.1	Measurement of a transmission line embedded between two transitions.....	76
Fig. 3.2	Coplanar waveguide-to-microstrip transition.....	76
Fig. 3.3	Reactively end-loaded transmission line.....	77
Fig. 3.4	Measured (a) reflection and (b) transmission s-parameter magnitudes of a Heraeus CT2000 microstrip transmission line of width $W = 210 \mu\text{m}$, length $d = 18.3 \text{ mm}$, and substrate height $h = 223 \mu\text{m}$	78
Fig. 3.5	Simulated and averaged measured reflection coefficients for the Heraeus microstrip line.....	79
Fig. 3.6	Attenuation constant of the Heraeus microstrip line using the Absolute Magnitude method (a) for a line length of 18.3 mm, (b) for three different line lengths: 1.80 mm, 2.65 mm, and 18.3 mm.....	80
Fig. 3.7	Attenuation constant of the Heraeus microstrip line using the Line Comparison method with $\Delta = 16.5 \text{ mm}$	81
Fig. 3.8	Transmission magnitudes of the microstrip Thru and Line standards for (a) the Heraeus CT2000 and (b) the Dupont 943 material systems listed in Table 3.1.....	82

Fig. 3.9	Illustration of the dimensional tolerance of LTCC processing: (a) a CPW-to-microstrip transition as designed, and (b) a photograph of the manufactured transition.....	83
Fig. 3.10	Successive transmission measurements of the same Heraeus microstrip Thru standard of Table 3.1.....	84
Fig. 3.11	Layout of a two-port, co-linear fed microstrip ring resonator.....	86
Fig. 3.12	Circuit models of the unloaded ring resonator, (a) ring decomposed into two sections, (b) transmission line model.....	87
Fig. 3.13	Loaded ring resonator models for (a) one-port gap coupling and (b) co-linear two-port gap coupling.....	89
Fig. 3.14	Magnitude of s_{21} for the loaded two-port ring resonator of Fig. 3.13(b) versus the ring circumference at 5.7 GHz. The microstrip width is $W = 280 \mu\text{m}$, the dielectric constant is 6.15, and the substrate height is $h = 187 \mu\text{m}$. The coupling gap length is $S = 50 \mu\text{m}$ and the attenuation of the microstrip is set to 0.1 dB/cm.....	90
Fig. 3.15	Smith Chart input impedance loci of undercoupled resonators, (a) parallel and (b) series.....	90
Fig. 3.16	Series model of the two-port ring resonator without parasitics, as seen by the external circuitry.....	91
Fig. 3.17	Microscope photograph of one of the coupling gaps: the vertical arc is a section of the ring whereas the horizontal line is the feed line, leading to the CPW probe pads.....	94
Fig. 3.18	Measured transmission magnitudes of the (a) Ferro1 and (b) DP943 ring resonators, showing seven and six resonance peaks, respectively.....	94
Fig. 3.19	Measured attenuation of the (a) Ferro and (b) Dupont 943 microstrip lines versus frequency.....	95

Fig. 3.20	Calculated loss tangents of the (a) Ferro A6-S and (b) Dupont 943 substrates.....	96
Fig. 3.21	Microstrip gap discontinuities: (a) a two-port end-to-end, (b) a three-port T, (c) the equivalent lumped element model for the T-gap.....	97
Fig. 3.22	C_g capacitor values for (a) a 100- μm microstrip gap using the Ferro A6-S parameters and (b) a 135- μm gap using the Dupont 943 parameters, both extracted from 2.5D simulations of a microstrip T-gap and expressed with respect to the values obtained by Yu's closed form equations.....	98
Fig. 3.23	C_p capacitor values for (a) a 100- μm microstrip gap using the Ferro A6-S parameters and (b) a 135- μm gap using the Dupont 943 parameters, both extracted from 2.5D simulations of a microstrip T-gap and expressed with respect to the values obtained by Yu's closed form equations.....	99
Fig. 3.24	Measured and calculated (a) transmittance and (b) reflectance of the Ferro A6-S microstrip rings at resonance. The calculated values are obtained using the lumped element ring model with capacitance values extracted from the T-gap simulations.....	100
Fig. 3.25	Measured and calculated (a) transmittance and (b) reflectance of the Dupont 943 microstrip rings at resonance. The calculated values are obtained using the lumped element ring model with capacitance values extracted from the T-gap simulations.....	101
Fig. 3.26	Input reflection loci of ring resonators on the Smith Chart. The dashed line represents the unloaded ring and is the same in both graphs, the arrow marking the unloaded parallel resonance. The solid line represents the two-port loaded ring for the cases of (a) tight coupling and (b) loose coupling.....	102

Fig. 3.27	Relative permittivity obtained from (a) the Ferro ring resonators and (b) the Dupont 943 ring resonator, using frequency correction.....	105
Fig. 3.28	Difference between the observed and calculated resonant frequencies, the latter simulated using the experimentally obtained material parameters and the lumped element model, for (a) the Ferro A6-S system and (b) the Dupont 943 system.....	106
Fig. 3.29	Hypothetical H-plane view of wave-fronts travelling down a laminated waveguide	107
Fig. 3.30	Two interfering plane waves with perpendicular polarisation to the plane...	108
Fig. 3.31	(a) Perpendicularly polarised plane wave incident on a row of wires and (b) its equivalent circuit with respect to propagation in the x-direction.....	109
Fig. 3.32	Equivalent rectangular waveguide having an effective width.....	110
Fig. 3.33	Laminated waveguide cavity with equivalent RWG width and length extensions Δa and Δd , respectively.....	114
Fig. 3.34	E-plane cross sectional view of the waveguide excitation showing the top layer coplanar probe pads.....	115
Fig. 3.35	Percentage error between the measured and calculated resonant frequencies for the three Dupont 943 LWG resonators, $\square$ - calculated using (3.21), $\bigcirc$ - calculated without any correction.....	116
Fig. 3.36	Percentage error between the measured and calculated resonant frequencies for the three Heraeus CT2000 LWG resonators having $p = 7.5r$, $\triangle$ - calculated using the grid impedance method, $\square$ - calculated using (3.21), and $\bigcirc$ - calculated without any correction.....	117

Fig. 3.37	Percentage error between the measured and calculated resonant frequencies for the three Heraeus CT2000 LWG resonators having $p = 10r$, $\triangle$ - calculated using the grid impedance method, $\square$ - calculated using (3.21), and $\circ$ - calculated without any correction.....	118
Fig. 3.38	A rectangular waveguide cavity.....	120
Fig. 3.39	Conductor and dielectric quality factors of a RWG when the loss tangent is (a) constant at $\tan\delta = 0.0025$ and (b) linearly increasing 0.0025–0.0033 over 15–70 GHz.....	123
Fig. 3.40	LWG structures for radiation loss simulations: (a) cavity mode and (b) transmission line mode.....	126
Fig. 3.41	Attenuation constants and quality factors extracted using the transmission line simulation method using the parameters of Table 3.6 and a guide length $d = 2.4$ mm.....	127
Fig. 3.42	Reflection and transmission magnitude of the Dupont 943 res4 resonator...	129
Fig. 3.43	Simulated RWG and measured LWG conductor and dielectric (a) Q ; (b) attenuation constant.....	131
Fig. 3.44	Dupont 943 loss tangent obtained from a ring resonator and by LWG cavities.....	132
Fig. 4.1	Firing profile of the experimental Ferro ferrite LTCC.....	139
Fig. 4.2	Microscope photographs of prototype ferrite LTCC modules.....	139
Fig. 4.3	Simplified transformer circuit for $B(H)$ measurements. Connection to external circuit elements is done through wafer probes.....	141
Fig. 4.4	Single-loop winding having three turns per unit cell length.....	143
Fig. 4.5	Double-loop winding having seven turns per unit cell length.....	144
Fig. 4.6	Five layer solenoidal transformer cross-section: (a) conceptual, (b) design rule adherence.....	145

Fig. 4.7	Cross section of the eight layer LTCC solenoid transformer.....	147
Fig. 4.8	Complete circuit schematic for the measurement of the ferrite B(H) curves..	148
Fig. 4.9	Measured input (V_i) and output (V_v) waveforms to and from the integrator..	149
Fig. 4.10	LTCC solenoid transformer: (a) layout screen capture, (b) photograph of the manufactured module.....	150
Fig. 4.11	Measurement of the LTCC ferrite solenoid transformer with low frequency probes.....	150
Fig. 4.12	Linear μ_{lin} and incremental μ_{incr} permeability are defined by the slopes of the peak B versus peak H graph.....	151
Fig. 4.13	Minor B(H) curves for two low-level H-fields: (a) $H_{peak} = 141$ A/m, (b) $H_{peak} = 382$ A/m.....	152
Fig. 4.14	Major hysteresis curve of the Ferro ferrite material.....	153
Fig. 4.15	(a) Peak flux density B_{peak} versus peak magnetic field H_{peak} , (b) linear μ_{lin} and incremental μ_{incr} relative permeability versus peak magnetic field.....	153
Fig. 4.16	(a) Remanent flux density B_r and (b) coercive magnetic field H_c versus peak magnetic field at 20 °C.....	155
Fig. 4.17	Thermally controlled chuck and Temptronic controller rack.....	156
Fig. 4.18	Saturation flux density B_s versus temperature.....	156
Fig. 4.19	(a) Remanent flux density B_r and (b) coercive magnetic field H_c versus temperature.....	157
Fig. 4.20	Two-port co-linear fed ring resonators: (a) gap-coupled, (b) line-connected..	159
Fig. 4.21	Circuit model of the two-port colinear-fed ring resonator (a) gap-coupled, and (b) line-connected (oversimplified).....	161

Fig. 4.22	Excitation of the ring with (a) even and (b) odd modes, and their equivalent transmission line circuits.....	162
Fig. 4.23	Terminated, parallel-connected, parallel two-port resonator.....	164
Fig. 4.24	Reflection coefficient magnitude of a line-connected two-port ring resonator. The dashed line uses the oversimplified model of Fig. 4.21(b), the solid line uses the even/odd mode model, and the circles are from a 2.5D simulation.....	166
Fig. 4.25	Reflection magnitude of a line-connected two-port ring near resonance: the solid line uses the even/odd mode model whereas the dashed line uses the lumped element equivalent circuit.....	167
Fig. 4.26	Measured reflection coefficient magnitude of the line-connected ring resonator.....	171
Fig. 4.27	Total attenuation of the line-connected ring resonator obtained by measuring the quarter-power unloaded bandwidth.....	172
Fig. 4.28	Effective permittivity versus frequency.....	173
Fig. 4.29	(a) Combined relative permittivity and relative permeability, and (b) loss tangent of the experimental ferrite.....	174
Fig. 4.30	Top view of a laminated waveguide cavity layout in ferrite. The top broad wall has been removed to show the inside structure and the ground-signal-ground pads have been made transparent to show their underlying vias.....	176
Fig. 4.31	Transmission magnitude of (a) the ferrite cavity having $a = 2350 \mu\text{m}$ and $d = 5750 \mu\text{m}$, and (b) the ferrite cavity having $a = 2000 \mu\text{m}$ and $d = 5000 \mu\text{m}$	177
Fig. 4.32	Combined relative permittivity and relative permeability, obtained using laminated waveguide cavities (squares) and ring resonators (circles). 178	

Fig. 4.33	(a) Measured unloaded quality factors of the first (triangles) and second (circles) cavities, compared to simulations using RWG formulae (Xs); (b) loss tangents required to simulate the unloaded Q of the first (triangles) and second (circles) cavities, superimposed on the results from the line-connected ring resonator (squares).....	179
Fig. 5.1	Wet Stack thick film ferrite layers and solenoidal windings for the fabrication of inductors.....	185
Fig. 5.2	Firing profiles of Heraeus CT2000 LTCC, of Wet Stack ferrite pastes IP9xxx, and of ferrite paste CL90-7057.....	188
Fig. 5.3	E-plane cross-section of a laminated waveguide showing ferrite paste used as thick film or as via fill material.....	190
Fig. 5.4	Variants of the slab-loaded nonreciprocal RWG phase shifter, where arrows indicate the direction of magnetic bias: (a) two thin slabs flush with the broad walls, (b) one full-height slab, (c) two antisymmetrically-biased full-height slabs, (d) a twin-toroid with dielectric loading (hashed area).....	192
Fig. 5.5	Nonreciprocal, transverse-biased ferrite phase shifter implemented using laminated waveguide, ferrite vias, and ferrite thick film.....	193
Fig. 5.6	Transverse-biased ferrite RWG phase shifters: (a) dual slab, (b) cylindrical vias, (c) E-plane cross-section through the vias.....	194
Fig. 5.7	Instantaneous phase shift within the via versus the distance from the centre of the via using the parameters of Table 5.2.....	196
Fig. 5.8	Phase shift of the two-post ferrite-loaded RWG having the parameters listed in Table 5.2 with a via pitch of 375 μm , calculated analytically and simulated using HFSS.....	197
Fig. 5.9	E-plane cross-section of half of a four-slab ferrite-loaded RWG, explicitly showing the five field regions I–V.....	198

Fig. 5.10	Magnitude of the left hand side of the transcendental equation for a four-slab ferrite-loaded RWG having the material parameters given in Table 5.2, with vias located 360 and 830 μm from the narrow walls.....	204
Fig. 5.11	Magnitude of the determinant of the leading matrix in (5.22) for (a) $-3000 < \beta < 3000$ rad/m, and (b) $1800 < \beta < 2600$ rad/m.....	205
Fig. 5.12	Percentage difference between H_1 and H_2 versus the propagation constant near the asymmetrical roots at (a) $-\beta_- = -1797.1$ rad/m and (b) $\beta_+ = 1909.5$ rad/m.....	207
Fig. 5.13	Percentage difference between H_1 and H_2 versus the propagation constant near the symmetrical roots at (a) $\beta_A = 2212.8$ rad/m and (b) $\beta_B = 2489.6$ rad/m.....	208
Fig. 5.14	Normalised field magnitudes of the reverse wave β_- along half the width of the guide: (a) h_z component, (b) e_y (and b_x) component.....	209
Fig. 5.15	H-plane cross-section of half of a four-post ferrite-loaded RWG.....	211
Fig. 5.16	Ferrite filling factor F^3 versus via diameter and waveguide width (a) three-dimensional graph, (b) same graph, viewed from above, where shading now indicates the relative value of F^3	213
Fig. 5.17	Precession of an electron's magnetic moment m about the bias axis H_0	214
Fig. 5.18	Weighted CPAR_WXZ figures of merit for the two-post phase shifter of Table 5.2 versus the offset of the via centres with respect to the waveguide narrow wall. The lower solid line is the weighted CPAR of the forward wave whereas the dashed line is that of the reverse wave. The bold dashed line is the sum of the two CPAR curves, normalised to the phase shift per centimetre, which is plotted as the bold solid line.....	216

- Fig. 5.19** E-plane cross sections of a two-post ferrite-loaded rectangular waveguide showing the complete magnetostatic circuit: (a) single toroid, and (b) double toroid. The light grey areas are the waveguide whereas the dark grey lines are the ferrite. The black segments are the DC bias lines, where the direction of the current is indicated by the $\odot$ and $\otimes$ arrowheads..... 217
- Fig. 5.20** A three-dimensional view of a unit cell of one of the ferrite toroids, with two upper and two lower flux return arms. A nine-layer double-loop bias winding excites the left via, which transfers the flux through the return arms to the interaction via that is in the waveguide, on the right..... 219
- Fig. 5.21** Screen captures of the magnetostatic field within the unit cell of a ferrite toroid for a bias current of 200 mA circulating through the nine-layer double loop: (a) within the return arm and (b) in the middle of the interaction via..... 220
- Fig. 5.22** Two-post ferrite loaded rectangular waveguide phase shift versus via placement at 35 GHz for a guide width of $a = 2300 \mu\text{m}$, a via diameter of $\varnothing = 150 \mu\text{m}$, a ferrite via pitch of $p = 375 \mu\text{m}$, and a flux density of $B = 90 \text{ mT}$ 221
- Fig. 5.23** Top view of an LTCC layer containing laminated waveguide narrow wall vias, ferrite vias, and bias windings..... 222
- Fig. 5.24** Top and side view of the transition from coplanar waveguide-to-microstrip, and of the transition from microstrip-to-laminated waveguide using a via probe extending into the laminated waveguide through an aperture in the microstrip ground plane. The via walls are not shown for sake of clarity..... 223
- Fig. 5.25** Cross-section of the two-post ferrite-loaded laminated waveguide phase shifter, showing the conductors in grey and the ferrite in black. All dimensions are in micrometres..... 224

Fig. 5.26	LTCC via filling procedure.....	225
Fig. 5.27	Photographs illustrating the blotting of the IP9220 ferrite paste on the back side of the green tape sheet after via filling.....	226
Fig. 5.28	Transverse view of two two-post ferrite-loaded laminated waveguide phase shifter samples made from Heraeus CT2000 LTCC.....	227
Fig. 5.29	Worst-case delamination sites around a stacked ferrite via.....	227
Fig. 5.30	Top view of the cracked return arms of two ferrite toroids.....	228
Fig. 5.31	Transmission amplitudes of five Heraeus zero-biased two-post ferrite-loaded waveguide phase shifters (A, C–F).....	229
Fig. 6.1	(a) A forward wave travelling in the $+z$ direction, from port 1 to port 2, and (b) are reverse wave travelling in the $-z$ direction, from port 2 to port 1.....	236
Fig. 6.2	Nonreciprocal propagation (a) with a certain magnetic bias H_0 , and (b) with a reverse bias H_0	237
Fig. 6.3	A ferrite dual-slab rectangular waveguide phase shifter.....	239
Fig. 6.4	Dual-slab remanant phase shift versus slab thickness t and slab spacing S at 10 GHz for a ferrite having a relative permittivity of $\epsilon_{rf} = 13$ and a remanant magnetisation of $M_r = 142$ kA/m (1786 G).....	240
Fig. 6.5	Dual-slab remanant phase shift versus slab thickness t for a slab spacing $S = 0$ at 10 GHz for a ferrite having a relative permittivity of $\epsilon_{rf} = 13$ and a remanant magnetisation of $M_r = 142$ kA/m inside a dielectric-filled waveguide.....	241
Fig. 6.6	Ferrite-filled, antisymmetrically-biased (ferrite-FAB) waveguide.....	242

Fig. 6.7	Magnitude of the transcendental equation for a ferrite-FAB phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm. The three lines represent different frequencies: the solid line is taken at 12 GHz, the dotted line at 8 GHz, and the dashed line at 6 GHz.....	245
Fig. 6.8	Ferrite-FAB normalised forward and reverse field magnitudes for a phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm: (a) e_y fields, (b) h_x fields, and (c) h_z fields.....	249
Fig. 6.9	Theoretical and HFSS-simulated phase shift versus frequency for a ferrite-FAB phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm.....	250
Fig. 6.10	HFSS simulated return loss of the ferrite-FAB phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm, with loss parameters $\tan\delta = 0.002$ and $\Delta H = 7.8$ kA/m (100 Oe).....	251
Fig. 6.11	Top and side view of the transition from coplanar waveguide-to-microstrip, and of the transition from microstrip-to-laminated waveguide using a via probe extending into the laminated waveguide through an aperture in the microstrip ground plane. The via walls are not shown for sake of clarity.....	253
Fig. 6.12	Cross-sections of the laminated waveguide in the interaction region. Windings exist on three LTCC layers: two winding columns are external to the waveguide whereas one is buried in the centre of the guide. Figure (a) is the cross-section taken through vias whereas (b) is taken between the vias. The waveguide core is represented by the shaded area, and the dashed lines represent the DC magnetic field.....	254
Fig. 6.13	Cross-sectional dimensions of the ferrite-FAB interaction region, all units in micrometers. The via diameter is $150\ \mu\text{m}$ and the layer thickness is approximately $105\ \mu\text{m}$	255

Fig. 6.14	Photograph of the fabricated LTCC module containing ferrite-FAB phase shifters 'A' and 'B'	256
Fig. 6.15	Measured (a) reflection and (b) transmission of the ferrite-FAB phase shifter 'A' for 21 bias levels ranging from -500 to +500 mA. The black lines are s_{n1} whereas the grey lines are s_{n2} , where $n \in \{1, 2\}$	256
Fig. 6.16	Measured phase shift of the FAB A shifter: (a) at 36 GHz versus bias, where the bias is varied in the direction of the arrows, and (b) at 500 mA bias versus frequency	257
Fig. 6.17	Measured phase shift of modules FAB C (no internal windings) and FAB D (no manifolds), (a) at 36 GHz versus bias current, (b) at 500 mA bias versus frequency	259
Fig. 6.18(a)	Magnetostatic flux density contour lines over one-quarter of the ferrite-FAB cross-section, <u>through vias</u> , at a bias level of 500 mA. Numbers indicate the flux density of the shades, expressed in teslas	261
Fig. 6.18(b)	Magnetostatic flux density contour lines over one-quarter of the ferrite-FAB cross-section, <u>between vias</u> , at a bias level of 500 mA. Numbers indicate the flux density of the shades, expressed in teslas	262
Fig. 6.19	Phase shift obtained using the mean flux density from 2D simulations combined with a theoretical analysis, compared to that of the measured FAB A circuit	263
Fig. B.1	Microstrip T-gap junction and its equivalent lumped element model	287
Fig. C.1	Unloaded (a) series and (b) parallel lumped element resonators	291
Fig. C.2	Two-port resonant circuits; Transmission: (a) series-connected series resonator, (b) parallel-connected parallel resonator; Absorption: (c) series-connected parallel resonator, (d) parallel-connected series resonator	294

Fig. C.3 Equivalent circuits as viewed by the LC tanks and their
corresponding coupling coefficient κ 295

List of Appendices

8.1 Appendix A: Specification data sheets of selected LTCC materials.....	282
8.1.1 Heraeus CT2000 LTCC specification data sheets.....	282
8.1.2 Ferro A6-S LTCC specification data sheet.....	284
8.1.3 Heraeus CL90-7057 ferrite paste specification data sheet.....	285
8.1.4 Heraeus IP9150 ferrite paste specification data sheet.....	286
8.2 Appendix B: Extraction of the gap capacitor values from a 2.5D simulation of a three-port microstrip T-gap.....	287
8.3 Appendix C: The measurement of quality factors of one- and two-port resonators using scattering parameter magnitudes at arbitrary power levels.....	290
8.3.1 Bandwidth levels of a two-port resonator.....	293
8.3.2 Bandwidth levels of a one-port resonator.....	301
8.3.3 Summary.....	303
8.3.4 Appendix C references.....	303

List of Acronyms and Abbreviations

2.5D	two-and-one-half dimensional
AC	alternating current
AJ	abrupt junction
AM-PM	amplitude modulation to phase modulation
B(H)	magnetic flux density versus magnetic field, (hysteresis)
BPSK	bipolar phase shift keying
CMOS	complimentary metal oxide semiconductor
CPAR	circular polarisation axial ratio
CPW	coplanar waveguide
CW	continuous wave
D/A	digital-to-analogue
DC	direct current
EM	electromagnetic
E-plane	the plane containing the electric field vector
F ³	ferrite filling factor
FAB	filled, antisymmetrically-biased
FEM	finite element method
FET	field-effect transistor
G	gauss, a deprecated unit of magnetic flux density
GaAs	gallium arsenide
GHz	gigahertz
GSG	ground-signal-ground
HAJ	hyperabrupt junction
HBT	heterojunction bipolar transistor
HEMT	high-electron-mobility transistor
HFSS	high frequency structure simulator
HMIC	hybrid microwave integrated circuit
H-plane	the plane containing the magnetic field vector
HPF	high pass filter
HTCC	high temperature cofired ceramic
IC	integrated circuit
IL	insertion loss
IQ	inphase-quadrature
InGaAs	indium gallium arsenide
Ka-band	the frequency range of 26.5–40 gigahertz
LPF	low pass filter
LTCC	low temperature cofired ceramic
LWG	laminated waveguide
MBE	molecular beam epitaxy
MCM	multi-chip module
MCM-C	multi-chip module, ceramic
MCM-D	multi-chip module, deposited
MCM-L	multi-chip module, laminated
MEMS	microelectromechanical systems

MESFET	metal-semiconductor field-effect transistor
MgMn	magnesium-manganese
MHz	megahertz
MIC	microwave integrated circuit
MMIC	monolithic microwave integrated circuit
mT	millitesla, a unit of magnetic flux density
NiZn	nickel-zinc
O.C.	open circuit
Oe	oersted, a deprecated unit of magnetic field strength
PCB	printed circuit board
PEC	perfect electrically conducting
pHEMT	pseudomorphic high-electron-mobility transistor
PIN	p-type, intrinsic, n-type
PTFE	polytetrafluoroethylene
Q	quality factor
QAM	quadrature amplitude modulation
RC	resistor-capacitor
RF	radio frequency
RMS	root-mean-squared
RWG	rectangular waveguide
S.C.	short circuit
Si	silicon
SMT	surface mount technology
TE	transverse electric
TEM	transverse electromagnetic
THz	terahertz
TM	transverse magnetic
TRL	thru-reflect-line
TTD	true time delay
TTL	transistor-transistor logic
V-band	the frequency range of 50–75 gigahertz
VNA	vector network analyser
VSWR	voltage standing wave ratio
X-band	the frequency range of 8–12 gigahertz

Glossary

a	waveguide width	r	radius, normalised resistance
A	area, auxiliary variable	R	resistance
b	waveguide height	$\mathfrak{R}$	reluctance
B	susceptance, flux density, auxiliary variable	s_{mn}	scattering parameter
B_r	remanent flux density	S	spacing
B_s	saturation flux density	t	time, thickness
c	speed of light, auxiliary variable	$\tan\delta$	loss tangent
C	capacitance, auxiliary variable	T	temperature, transmission coefficient
d	distance, auxiliary variable	T_c	Curie temperature
D	auxiliary variable	u_p	phase velocity
e	auxiliary variable	W	width, energy
E	electric field, auxiliary variable	x	Cartesian coordinate, normalised
f	frequency, auxiliary variable		reactance, auxiliary variable
f_c	cutoff frequency	X	reactance
F	auxiliary variable	y	Cartesian coordinate
G	conductance, voltage gain, auxiliary variable	Y	admittance
h	height	z	Cartesian coordinate, normalised impedance
H	magnetic field, auxiliary variable	Z	impedance
H_c	coercive magnetic field (coercivity)	α	attenuation constant
I	current	α_c	conduction attenuation constant
j	imaginary number	α_d	dielectric attenuation constant
k	wavenumber	α_r	radiation attenuation constant
ℓ	length	β	imaginary propagation constant
L	length, inductance	β_+	forward propagation constant
m	magnetic moment, integer index, auxiliary variable	β_-	reverse propagation constant
M	magnetisation	γ	sensitivity, complex propagation constant
M_r	remanent magnetisation	γ_r	gyromagnetic ratio
M_s	saturation magnetisation	Γ	reflection coefficient
n	integer index, turn density	δ	double-sided bandwidth, mode order
N	integer number	Δ	length difference
p	pitch, power fraction	ΔIL	insertion loss variation
P	power	Δt	time delay
Q_0	unloaded quality factor	$\Delta\phi$	phase shift
Q_c	conductor quality factor	ϵ	permeability
Q_d	dielectric quality factor	ϵ_0	permeability of vacuum
Q_E	external quality factor	ϵ_{eff}	effective permeability
Q_L	loaded quality factor	ϵ_r	relative permeability
Q_r	radiation quality factor	ζ	auxiliary variable

η	wave impedance
η_{rad}	radiation efficiency
θ	electrical length, angle
κ	coupling coefficient, permeability tensor variable
λ	wavelength
λ_g	guided wavelength
μ	permeability, permeability tensor variable
μ_0	permeability of vacuum
μ_r	relative permeability
ν	auxiliary variable
ξ	auxiliary variable
σ	conductivity
τ	auxiliary variable
ϕ	phase, angle
Φ	magnetic flux
$\varnothing$	diameter
ψ	phase offset, auxiliary variable
ω	radian frequency
π	auxiliary variable
$\mathcal{K}$	auxiliary variable

Chapter One — Introduction

1.1 Motivation

A tremendous number of technological advances in the field of electronic packaging have been made over the past few years. One packaging technology in particular, known as low temperature cofired ceramic (LTCC), has been in the forefront of innovation. Although LTCC was originally conceived as simply a means to house and to interconnect bare-die integrated circuits, commercial suppliers of LTCC materials have continued to develop more exotic products for use in the fabrication of LTCC. For example, new resistive, dielectric, and ferrite pastes may be printed on LTCC, allowing resistors, capacitors, and inductors to be included, respectively, as integral parts of the package instead of being bought and mounted as separate components.

LTCC is a multilayer packaging system, meaning that an LTCC module is made of several layers of ceramic, each containing conductor traces and vias for connecting to different levels of the module. Novel experimental ceramic tape materials, of which the layers are made, have recently been offered, including layers of mixed dielectric constants, very high dielectric constants, and even tape made of LTCC ferrite. Owing to its multilayer construction, LTCC is a three-dimensional packaging technology, a property that contrasts with more conventional single-layer planar circuitry. The availability of the third dimension allows three-dimensional microwave circuitry to once again be conceived and developed, as was the case before the onset of the planar paradigm in the early 1980s. For example, stripline and even rectangular waveguide transmission lines may be buried deep within the LTCC package, freeing the top layer of the module to receive more

integrated circuits. The low losses of the LTCC tape material and of the high conductivity silver and gold conductors yield excellent performance at RF and microwave frequencies. Given this, researchers have recently begun to characterise the materials at millimetre-wave frequencies.

Concurrent to the advances being made in LTCC, the communications age fostered an unprecedented demand for telecommunications services in the late 1990s. Such was the demand that telecommunications providers envisioned wideband wireless services provided via direct end-user communication with constellations of orbiting satellites. As opposed to satellite telephony services, high bandwidth services such as internet and video streaming were proposed, requiring unprecedented operating frequencies in the millimetre-wave (> 30 GHz) range. For efficient spectrum use, the satellite, using an onboard intelligent phased array of antennas, would dynamically alter the number, size, and direction of the radiated beams to maintain communications with pockets of ground users according to demand. The control of such an array of antennas necessitates hundreds of onboard phase shifters, which leads to a dilemma. At millimetre-wave frequencies, the phase shifters offering the best performance are made from three-dimensional rectangular waveguide that is loaded with ferrite material. Although a mature technology, ferrite phase shifters are heavy and large (each one having proportions of a few centimetres), are expensive to fabricate, and are difficult to integrate into a compact satellite. For integration, the only other alternative is to use planar, semiconductor phase shifters, which are inherently small and amenable to high volume production. Unfortunately, the losses of semiconductor millimetre-wave phase shifters are high compared to their ferrite waveguide counterparts, implying that more amplification of the millimetre-wave signals

would be required to compensate for their loss, exacerbating the requirements on the amplifiers and needlessly consuming more satellite power. What is ultimately required for this application is a better millimetre-wave phase shifter, one that can simultaneously offer excellent performance, compatibility with mass production, and a size that is small enough for integration within a satellite's compact payload.

Given this need, and coupled with the recent advances made in packaging technology, the concept of a millimetre-wave phase shifter made entirely from ferrite LTCC material becomes a plausible idea, the implication of which is remarkable: the phase shifter would no longer be regarded as a separate, purchasable component, but would instead become an integral part of the package. The potential for cost savings alone are significant. Furthermore, being made from a package, the phase shifter would be in intimate proximity and easily connected to other integrated circuits housed within the same module. LTCC modules are inherently small and flat, potentially solving the integration problem. In addition, LTCC packages are specifically intended for low cost mass production. The idea of an LTCC-integrated millimetre-wave phase shifter is pursued in the remainder of this thesis, representing one of the first attempts to make such a control device solely from packaging materials.

1.2 Objectives

Several barriers must be overcome before the ferrite LTCC phase shifter concept can be successfully proven. Given the plethora of alternative phase shifter designs that are currently available, the state of the art in millimetre-wave phase shifters must be assessed and can only be achieved by a thorough review of published material on the subject. Only

after the validation of this first thesis objective can the potential to improve upon alternative technologies be determined.

The selection and procurement of appropriate ferrite materials are not simple tasks. Various suppliers offer different LTCC product lines that are incompatible with each other. Furthermore, ferrite that is compatible with LTCC is still regarded as an experimental material and is only offered by a limited number of companies. As such, obtaining appropriate ferrite samples is presently a challenge and there is no guarantee that the ferrite will be entirely compatible with standard LTCC processing methods, requiring its compatibility to be determined empirically. The exact material properties of the ferrite being unavailable, characterisation methods must also be devised for evaluating its magnetostatic properties, such as its saturation magnetisation, remanant flux density, coercive force, and others, which represents the second objective of this thesis.

The frequency range involved also poses many challenges. LTCC materials have yet to be adequately characterised at millimetre-wave frequencies and variations in the fabrication of the materials can also alter their properties. Hence the third objective is to develop appropriate methods for characterising both the loss and the dielectric constant of LTCC materials in the millimetre-wave range. The same methodology must also be applied eventually to the ferrite materials.

Finally, it is unlikely that conventional ferrite-loaded waveguide phase shifter configurations will be realisable in LTCC because of design-rule limitations that are imposed in the LTCC environment. As such, the fourth and final objective is to devise an appropriate alternative ferrite phase shifter configuration, beginning with the development

of a new theoretical framework, the application of an appropriate design methodology, the fabrication of prototypes, and concluding with their test and validation.

1.3 Thesis organisation

The thesis consists of seven chapters. Chapter Two establishes the state of the art in phase shifter technology at millimetre-wave frequencies, in which the concept of phase shift and the definition of phase shifter performance metrics are introduced. An exhaustive overview of semiconductor phase shifter technology follows, along with various comparisons between published semiconductor millimetre-wave phase shifters. The same is then done for ferrite phase shifters. Chapter Two concludes with an introduction to multi-chip packaging technology, with emphasis on LTCC materials and fabrication methods.

Chapter Three is dedicated to the characterisation of LTCC transmission media and materials at millimetre-wave frequencies, with emphasis placed on techniques that require only common microwave measurement equipment. Characterisation methods are divided into non-resonant and resonant categories, and the latter type is shown to be more reliable than the former. A new theoretical framework for the determination of material loss and permittivity using a two-port microstrip ring resonator is presented, along with measured results. A new theory for the estimation of laminated waveguide cavity resonant frequencies is presented, and the cavities are subsequently used to characterise the loss of the LTCC material.

Chapter Four describes a particular experimental ferrite LTCC tape system that has been secured for this work. Other than its firing profile, no other data have been disclosed by its supplier, implying that all of the material properties must be independently

measured. Chapter Four opens with a description of the tape's processing steps that have been followed to yield high quality fabricated test samples. The magnetostatic properties of the ferrite are then obtained from measurements of a novel solenoidal transformer, the design of which is also presented therein. A new microstrip ring resonator, slightly different from the one used in Chapter Three, is then presented and is used to characterise the permittivity and loss of the ferrite at microwave and millimetre-wave frequencies.

An experimental ferrite paste has also been obtained that can potentially be used to print lines and fill vias. Chapter Five presents a detailed theoretical analysis of laminated waveguide phase shifters that are loaded with ferrite-filled via holes. The derivation is presented for the case of two via holes in the guide's cross section, and then for the case of four via holes. Unfortunately, these configurations are shown to yield little theoretical phase shift, and worse still, the ferrite paste has proven to be incompatible with LTCC.

A new laminated waveguide phase shifter that is made entirely from ferrite tape is presented in Chapter Six. The theory governing the performance of this new device, being ferrite-filled and antisymmetrically-biased, is rigorously derived from first principles and is shown to produce significant amounts of phase shift at millimetre-wave frequencies. Prototypes have been manufactured and their measured results are provided therein.

Chapter Seven contains the conclusions, lists the contributions of this thesis, and suggests future work to enhance the performance of the new phase shifter. A total of three appendices follow: Appendix A provides specification data sheets for LTCC materials, Appendix B contains the mathematical derivations of capacitance values used in Chapter Three, and Appendix C derives new expressions that are used in Chapters Three and Four to calculate the quality factors of resonators.

1.4 Thesis Contributions

This research has provided many significant contributions to the areas of material characterisation and microwave/millimetre-wave phase shifters, and these are summarised in point form below.

- The thesis contains the most exhaustive and comprehensive review to date of microwave phase shifter technology and performance up to the year 2000.
- Non-resonant characterisation structures have been shown to be unreliable in LTCC due to non-repeatability of the probed measurement.
- The one-port microstrip ring resonator analytical model has been improved and extended to the case of a two-port ring, and it has been used to characterise the permittivity and loss tangent of the Dupont 943 and Ferro A6-S material systems from 5 to 40 GHz.
- The two-port *line connected* microstrip ring has also been rigorously analysed and serves as a practical alternative when small coupling gaps cannot be manufactured.
- Laminated waveguide cavities have also been analysed and used to characterise LTCC materials, and a new analytical model, the *grid impedance method* has been evaluated for predicting the phase constant of laminated waveguides.
- The scalar equations for measuring the three quality factors of one- and two-port resonators at arbitrary bandwidths have been rigorously derived and practically applied in this thesis.
- An *in situ* solenoidal transformer has been analysed, designed, and implemented to provide the magnetostatic properties (major hysteresis curve and Curie temperature) of an unknown ferrite LTCC material.

- The analysis of ferrite post-loaded laminated waveguide phase shifters has been presented for both the two-post and four-post cases, including the derivation of the transcendental equations.
- A totally new phase shifter that is ferrite-filled and antisymmetrically biased has been derived from first principles, manufactured in LTCC, and demonstrated at 36 GHz.

Chapter Two — Introduction to millimetre-wave phase shifters and to LTCC packaging technology

Phase shifters are ubiquitous microwave control components that are used extensively in scanning antenna arrays. Although their history spans over half a century, microwave phase shifters are continually evolving, as their performance is directly related to technological advances in the areas of materials science and semiconductor fabrication. The maturity of the technology does not imply that phase shifters always perform well. As will be demonstrated in this chapter, there are severe tradeoffs to be made between the performance and the integrability of phase shifters at millimetre-wave frequencies. Ferrite phase shifters are known to have the best performance but are difficult to integrate with planar and/or monolithic circuitry. Conversely, integrated monolithic semiconductor phase shifters suffer from poor performance at millimetre-wave frequencies. A new approach to the construction of millimetre-wave phase shifters is warranted, and one such alternative, using packaging material, is presented at the end of this chapter.

Chapter Two is subdivided into seven sections. The definitions of phase shift and of criteria related to the evaluation of phase shifters are presented in Section 2.1. The various types of switches and varactors that are used in semiconductor phase shifters are introduced in Section 2.2, whereas Section 2.3 consists of an extensive compilation of semiconductor phase shifters operating at millimetre-wave frequencies. The concepts related to ferrite phase shifters are introduced in Section 2.4, along with a compilation of documented ferrite millimetre-wave phase shifters. Section 2.5 serves as an introduction to multi-chip packaging and to low temperature cofired ceramic (LTCC), wherein LTCC packaging is shown to be a potential candidate for solving the apparent contradiction

between ferrite and integration. A summary of the findings of this chapter is given in Section 2.6, and Section 2.7 lists the references.

2.1 Terminology

Before elaborating on current phase shifter technology, the concept of phase shift must be grasped, and criteria related to the performance of phase shifters must be defined. This section introduces the concept of phase shift as it applies to a time-harmonic signal. In addition, the various metrics that are used to evaluate and to compare phase shifter performance are also presented herein. These figures of merit are used subsequently in this chapter to assess the performance of documented phase shifters.

2.1.1 Phase definition and convention

In the work that follows, a propagating space-time harmonic signal, denoted by $\mathcal{V}(z, t)$, obeys the following equation:

$$\mathcal{V}(z, t) = \Re\{V_0 e^{j(\omega t - \beta z - \psi)}\} = V_0 \cos(\omega t - \beta z - \psi) \quad (2.1)$$

where V_0 is the maximum signal amplitude, $\omega = 2\pi f$ is the radian frequency of the signal, t is time, β is the propagation constant in the $+z$ -direction, and ψ is some constant phase offset. For succinctness, it is common practice to suppress the cosinusoidal time dependence, leaving only $V(z) = V_0 e^{-j(\beta z + \psi)}$ to represent the wave, a form known as *phasor notation*. The *absolute phase* of the signal is

$$\phi = \beta z + \psi. \quad (2.2)$$

A *phase shifter* refers to any device that is capable of altering the absolute phase of a signal, providing a *phase shift* of:

$$\Delta\phi = (\beta_2 z_2 + \psi_2) - (\beta_1 z_1 + \psi_1) \quad (2.3)$$

which is the difference in phase (or relative phase) between two *states* of the phase shifter, which are explicitly indicated by the subscripts. Usually, a phase shifter is only capable of controlling one of the three variables in (2.2), the others remaining constant. For example, as will be shown in subsequent parts of this chapter, a *switched-line* phase shifter changes the path length of the signal z whereas a *ferrite* phase shifter modifies the propagation constant β .

2.1.2 Constant phase and true time delay phase shifters

A device that provides an invariant phase shift with respect to frequency is known as a *constant phase* phase shifter. An example of an application requiring such a response is serrodyne frequency translation [1]. A *true time delay* phase shifter is one that imparts a frequency-invariant delay to a signal, a property that is desirable in wideband antenna arrays. Using (2.1) and (2.2), the apparent relative *time delay* introduced by a frequency-dependent absolute phase $\phi(f)$ is:

$$\Delta t = \frac{-\phi(f)}{2\pi f} \quad (2.4)$$

which, when taken at some reference frequency f_{ref} , is:

$$\Delta t = \frac{-\phi(f_{ref})}{2\pi f_{ref}}. \quad (2.5)$$

For a constant time delay, substitution of (2.5) into (2.4) yields the required absolute phase response as a function of frequency:

$$\phi(f) = \phi(f_{ref}) \frac{f}{f_{ref}} \quad (2.6)$$

which is plotted in Fig. 2.1 for nine different phase states $\phi(f_{ref})$.

Fig. 2.1 Absolute phase of a true time delay phase shifter for nine different states.

The slope of each line is proportional to the phase shift at the reference frequency. With respect to (2.1) and noting that $\beta = \omega/u_p$, where u_p is the phase velocity, the argument of the cosine can be rewritten as $\omega(t - z/u_p) - \psi$. Hence phase shifters that provide constant values of z/u_p and ψ as a function of frequency provide a true time delay.

2.1.3 Phase errors and figures of merit

An ideal phase shifter provides either a constant phase shift or a constant time delay and introduces neither insertion loss nor return loss to the circuit. Since such an ideal device is non-existent, appropriate performance measures are required to assess the phase shifter's deviation from ideal behaviour. Of prime importance is the difference between the desired and the observed phase shift, or *phase error*. This error is especially pertinent to digital phase shifters because of their lack of fine-tuning. Phase error is not often quoted for analogue phase shifters because these can theoretically yield any exact phase

shift value by fine-tuning the driver circuitry or by modifying the values of a lookup table. Some analogue phase shifters are designed to yield a linear phase shift versus the drive voltage so that a lookup table is not required. In this case, a phase shifter's deviation from linear response is another appropriate figure of merit.

Digital phase shifters are comprised of N units that discretise the 360° phase space. Their cascade yields 2^N phase states, admitting a worst-case quantisation error of half the smallest bit's phase value, or $360^\circ/2^{N+1}$. For example, a four-bit phase shifter has 180, 90, 45, and 22.5° elements, so the worst case quantisation error is 11.25° . This quantisation error is inherent and can only be reduced by increasing the number of bits in the cascade. When discussing phase error, it is not the quantisation error but rather the shifter's deviation from nominal that is implied. The extent of the phase error limits the number of bits implemented in the cascade; for example, a four-bit shifter that suffers from 15° errors would not be improved by the addition of a fifth (11.25°) bit since the precision of this bit would not be apparent. Although the performance of individual phase bits is sometimes quoted, the individual errors are compounded when the bits are connected in series, especially when impedance mismatch is present, which produces an additional phase ambiguity [2].

The phase error of a single state at a given frequency is trivial: if the shifter is set to 135° and produces an actual phase shift of 120° , the phase error ϕ_{err} is -15° . Phase error is a function of frequency, and when each state of a multi-bit shifter is considered, a daunting amount of error data may result. To alleviate this problem, phase shifter errors are often compiled into statistical form, which can unfortunately mask the performance of a poor device. Single-valued extrema figures also exist, such as the *maximum phase*

error, which is expressible in either absolute degrees or as a percentage. An error expressed in absolute degrees is misleading for small phase shift values, especially when the shifter's performance is compiled over all of the states from 0 to 360°. Small phase bits can have small absolute errors that are in fact large percentages of the desired phase shift, hence percentage error is the more relevant metric. At a given frequency, the errors of all phase states can be combined into an *RMS phase error*:

$$\phi_{err}^{RMS}(f_0) = \sqrt{\frac{1}{2^N} \sum_{n=1}^N \phi_{err}^2(n, f_0)} \quad (2.7)$$

which has been accepted as a common error statistic, even though it can mask large errors encountered on certain states, e.g. [3]. The RMS error can also be averaged over frequency, yielding a metric known as the *average RMS phase error*.

As opposed to constant phase shifters that usually provide a maximum shift of 360°, the maximum delay of a true time delay (TTD) shifter is application dependent, hence their minimum and maximum available delays are appropriate figures of merit. The time delay error is expressible as a function of frequency and can also be quoted in terms of extrema and RMS values.

Insertion and return loss are other important properties. As with phase errors, losses vary both as a function of frequency and phase state. Return loss is relatively simple in that it should be kept below a certain value, with 10 dB being a common figure. Insertion loss should also be kept low and, in addition, should also be kept invariant. As was the case with phase error, single-valued figures such as the average, extrema, and RMS are equally relevant to insertion loss. In cascaded shifters such as digital ones, the insertion loss increases as more stages are added. Hence there is a trade-off between insertion loss

and quantisation error. When comparing digital phase shifters, the amount of loss per bit is a relevant figure of merit. This value remains constant whenever additional bits add the same average insertion loss, even though the total loss through the phase shifter increases. Comparisons with, or between, analogue phase shifters should use the figure of merit defined by the available phase shift divided by the total loss, expressed in degrees per decibel ($^{\circ}/\text{dB}$). Finally, switching time and power handling (1 dB compression points) are other key performance metrics, although these are rarely quoted in the open literature.

2.2 A review of semiconductor phase shifter devices

The performance of a phase shifter is directly related to the type of semiconductor device that is used in its construction. As such, it is worthwhile to review the devices that have been commonly used in semiconductor phase shifter circuitry, namely varactor diodes, PIN diodes, field-effect transistor (FET) switches, and microelectromechanical system (MEMS) switches.

2.2.1 Varactor diodes for analogue phase shifters

Although the term *varactor* may be used to describe any variable reactance, in the context of microwave semiconductor control, the term usually refers to the reactance of a reverse-biased diode junction. Reverse-biased PN and Schottky barrier diodes provide the voltage controllable capacitance that is exploited in many analogue shifter designs [4]. A reverse-biased junction may be modelled as the connection of two varistors and one variable capacitor, as illustrated in Fig. 2.2.

Fig. 2.2 *Model of a reverse-biased PN or Schottky junction.*

The junction resistance R_J is that of the depletion region and is usually dominated under reverse bias by the lower reactance of the junction capacitance C_J . The series resistance R_S is that of the contact, bulk, and undepleted semiconductor regions, the latter varying inversely with the width of the depletion region. Monolithic integrated Schottky barrier junctions, shown in Fig. 2.3, are usually constructed of N type semiconductor only, which has a higher carrier mobility (and hence less resistance) than P type. For this reason, Schottky barrier junctions are preferred at microwave frequencies over diffused (PN) junctions.

Fig. 2.3 *Gallium arsenide integrated Schottky varactor diodes: (a) mesa type, (b) planar type.*

Although MMIC phase shifters may be constructed in either the mesa [5], [6] or the planar diode configuration [7]–[12], only the planar one lends itself well to economical foundry production [13]. However, planar Schottky diodes usually have lower breakdown voltages than their mesa counterparts due to the higher intensity of the electric field near

the edges of the gate. To maintain the same breakdown voltage, the doping of the N epitaxial layer must be reduced, which increases R_S .

The junction capacitance of a mesa Schottky diode can be approximated by a parallel plate model, wherein the width of the depletion region represents the dielectric gap. More rigorous equivalent models can be derived for planar Schottky diodes, although these can usually be reduced to first order RC circuits [8], [9], [13]. The generalised expression for the junction capacitance is a function of the forward applied bias V_A and is given by:

$$C_J(V_A) = C_{J0} \left(1 - \frac{V_A}{V_{bi}} \right)^{-\gamma} \quad (2.8)$$

where C_{J0} is the zero bias capacitance and V_{bi} is the built-in junction potential. The exponent factor γ , called the sensitivity factor [10], is a parameter that depends on the doping profile of the epitaxial N region. Canonical abrupt junctions (AJ) have $\gamma = 0.5$. Nonlinear tapered doping profiles, called hyperabrupt junctions (HAJ), have been developed to yield higher sensitivity, although the latter becomes a function of the applied bias. The custom doping profiles of HAJ junctions are often doped using Molecular Beam Epitaxy (MBE) or selective ion implantation, however these processes are expensive compared with the standard MESFET channel implantation that is used to construct AJ planar diodes [13], [14].

A common figure of merit for the Schottky varactor is the cutoff frequency resulting from the parallel connection of R_S and $C_J(V_A)$, which is normally quoted at a bias level of $V_A = -4$ V:

$$f_c(V_A) = \frac{1}{2\pi R_S C_J(V_A)} \quad (2.9)$$

which, although expressed as a frequency, is indicative of the insertion loss and isolation that can be expected from the device.

2.2.2 PIN diode switches for digital phase shifters

The PIN diode consists of an intrinsic (undoped) semiconductor layer located between degenerate (very heavily doped) P and N type regions. PIN diodes have long been used as switches for microwave signals [15]. When operated as such, the PIN diode is either forward or reverse biased, providing either a low or a high series impedance, respectively. The model of the PIN diode is illustrated in Fig. 2.4.

Fig. 2.4 *PIN diode schematic and equivalent circuit.*

The junction resistance R_J can vary over several orders of magnitude depending on the forward bias current level. When forward biased, the junction resistance is usually on the order of one ohm and dominates any junction capacitance. Conversely, when reverse biased, it is the junction capacitance that dominates the very high junction resistance. The resistance R_S is due to inevitable bulk resistivity, whereas C_P and L_P represent parasitic capacitance and inductance, respectively. The junction capacitance is smaller and less voltage dependent than that of a varactor, thereby allowing the PIN diode to better approximate an open circuit when reverse biased.

Increasing the intrinsic region width has the simultaneous desirable effects of decreasing the junction capacitance and increasing the power handling of the device. The only detriment of a wide intrinsic region is that the switching speed of the diode also increases [7:pp.409–412]. When the diode bias is switched from forward to reverse, the charge stored in the intrinsic layer must be removed, a process that is accomplished by carrier drift and by recombination within the layer. Larger reverse bias voltage levels can decrease this *reverse recovery* time only up until carrier velocity saturation occurs. Further, high forward bias injection levels, which are required to maintain a low impedance state, also increase the stored charge, hence there exists a compromise between the diode's attainable low impedance level and its switching time. The vertical mesa structures shown in Fig. 2.5(a) are rendered compatible with planar processing techniques by the use of an air-bridge [16].

Fig. 2.5 PIN diode construction: (a) vertical mesa, (b) planar.

Mesa PIN diodes require multiple epitaxial growth steps [17]–[19], which, along with the subsequent mesa etching steps, are incompatible with MESFET processing and thereby incur a cost penalty. However, high performance GaAs mesa diodes, fabricated using

MBE, are fully compatible and available in *heterojunction bipolar transistor* (HBT) technology [20], [21], although GaAs HBTs are not yet as widely available as MESFETs and again incur a cost penalty. Planar millimetre-wave PIN diodes, such as the one shown in Fig. 2.5(b), have been fabricated using selective ion implantation [22]–[25]. Apart from the P+ implant requirement, these diodes are compatible with MESFET techniques.

As is the case for the varactor diode, figures of merit have also been proposed for PIN diodes. Under reverse bias conditions, the junction resistance is always dominated by C_J , and the former is usually omitted from the definition of the cutoff frequency:

$$f_c = \frac{1}{2\pi C_J R_F}. \quad (2.10)$$

where R_F is the total equivalent forward bias resistance. Planar devices have demonstrated f_c values that are comparable to mesa PINs, and have been shown to handle power levels as high as 1.6 W [23]. It must be noted, however, that the width of the intrinsic region in planar devices cannot be made as thin as that of the mesa devices, hence the mesa devices have a switching speed advantage over planar ones. Monolithic mesa PIN diodes have been fabricated in both GaAs [17], [18], [26] and more recently in InGaAs [19]. Typical f_c values are in the range of a few THz for the former and between 10 and 20 THz for the latter. Although InGaAs PIN diodes have demonstrated the best performance, the material system is more costly, and its lower bandgap limits its power handling capability as compared to GaAs. Planar PIN diodes have been demonstrated in both GaAs and silicon with both types achieving f_c values of a few THz.

2.2.3 Transistor switches

Owing to their very high cutoff frequency values, PIN diodes continue to offer the best insertion loss and isolation values when used in switching applications. In addition, it has also been found that PIN diodes are more linear devices than are FETs [27], [28]. The major detriment to the widespread use of PIN diodes is that they have remained incompatible with low cost FET fabrication technology. Because of this, much research has focused on the refinement of FET switches.

Switching is achieved in a FET by modulating the drain-source channel using an applied gate voltage. No amplification is required, therefore no DC bias is applied between the drain and source of the FET [29]. Such unbiased devices are often referred to as *passive* FETs or *cold* FETs. With the bi-directional RF path passing between the source and drain, the FET's off-state corresponds to the channel being pinched off, whereas its on-state consists of a conducting channel. For depletion mode FETs, the on-state requires a null gate voltage V_g whereas the off-state requires a gate voltage that is more negative than the pinchoff voltage V_p . The familiar linear operation curves of the FET, along with its equivalent circuits in both states, are given in Fig. 2.6 [30].

Fig. 2.6 Unbiased switching FET conduction curves and equivalent circuits: (a) conduction curves, (b) high-impedance state, (c) low impedance state.

When the FET is on, the total resistance R_L is simply the sum of the source, drain, and channel resistances. When the FET is off, the high resistivity of the pinched off channel is shunted by the drain-to-source capacitance C_d as well as by the gate-to-drain and gate-to-source capacitances C_g , the latter two being of equal value when the FET is unbiased. When in the off-state, a signal path exists between the channel and the gate. To prevent signal leakage into the gate, a choke consisting of either a large integrated resistor and/or an inductance is used at that terminal. Since negligible current flows into the gate, the addition of the resistor has no impact on the switching performance. This contrasts with a PIN diode, which consumes DC power when the diode is conducting. For this reason, PIN diodes require bias chokes and blocking capacitors—elements that can quickly consume die space. Furthermore, although FET drivers must usually convert the CMOS or TTL control signal to appropriate gate voltages, minimal current is drawn by the FETs, thereby simplifying the driver design [20]. PIN drivers, conversely, in addition to shifting voltage levels, must also be designed to supply a regulated current when the diode is on, and to sink significant quantities of charge during reverse recovery [31], [32].

When in the off-state, it can be shown that the equivalent FET circuit can also be reduced to that of a series RC circuit [33]. The value of the equivalent off-state capacitance and resistance R_L can then be used to express the switching cutoff frequency f_c of the FET for comparisons with PIN diodes. As previously mentioned, while PIN diode f_c values typically range in the terahertz region, FET switching cutoff values are typically hundreds of gigahertz, with HEMT/pHEMT technologies and smaller gate lengths yielding the highest values [16], [34]–[36]. The difference in the f_c values indicates that FET switches will have higher insertion loss and lower isolation values than identical designs using PIN diodes.

FET switches have a further advantage over PIN diodes with respect to switching speed. As previously mentioned, PIN diode transitions involve injecting surplus carriers into the intrinsic region and the removal of this charge via drift and/or the relatively slow process of recombination. The controllable resistance in the FET switch is determined by modulating the population of majority carriers and is not limited by minority carrier lifetime. FET switching speed is therefore primarily dependent on the rise time of the external circuit that controls the gate voltage [29]. Whereas PIN diode switching times range from nanoseconds to tens of nanoseconds and more [37], [23], [26], [38], FETs can be switched in hundreds of picoseconds [33], [39]–[41].

Marked improvements in the power handling abilities of FET switches have been made, which are achieved by using multiple gate FETs, by series-connecting FETs (“stacking”), or by controlling the voltage levels by modifying line impedances [32], [42]. While FET switches have not reached kilowatt levels, functional switches have been

demonstrated recently that can handle levels ranging from a few watts to beyond 100 watts CW [32], [42], [43].

Direct comparisons between FET switches, let alone between PIN diode and FET switches, are difficult because of the many different system parameters, i.e. frequency, switch configuration, power level, desired switching time, etc. A PIN and a FET switch, each operating up to 20 GHz and using the same switch configuration, are compared in Table 2.1.

Table 2.1 A comparison between PIN and FET switches up to 20 GHz.

	PIN diode switch [37] 1991	FET switch [33] 1987
insertion loss	< 1 dB	< 2 dB
isolation	> 42 dB	> 25 dB
switching time	15 ns	< 400 ps
1 dB compression	21 dBm	18 dBm

In Table 2.2, a PIN and a FET switch are compared at V-band.

Table 2.2 A comparison between PIN and FET switches in V-Band.

	PIN diode switch [23] 1987	FET switch [40] 1988
frequency	62–66 GHz	59–61 GHz
insertion loss	< 1.5 dB	< 1.5 dB
isolation	> 30 dB	> 25 dB
switching time	< 2.4 ns	< 1 ns
1 dB compression	> 1.6 W	450 mW
die size	1.9 x 3.4 mm	0.8 x 2.5 mm

The insertion loss and especially the isolation of PIN switches is superior to that of FET switches. Further, the comparisons illustrate the faster switching ability of FETs over PIN diodes. Finally, due of their more complicated bias requirements, PIN diode switches consume more die space than FET switches.

2.2.4 Microelectromechanical system (MEMS) switches and varactors

MEMS transducers, although still considered as semiconductor devices, are quite different from the others presented thus far, in that the signal does not flow through any bulk semiconductor material. Instead, MEMS transducers consist of microscopic metallic membranes that are mechanically moved and/or distorted by electrostatic attraction. Two common configurations of a MEMS transducer are illustrated in Fig. 2.7 [44].

Fig. 2.7 *Micromechanical system (MEMS) transducer configurations: (a) cantilever switch, (b) bridge varactor.*

The configuration of Fig. 2.7(a) is a true metal-to-metal switch that has either an on-state (lever down) or an off-state (lever up). The second configuration, shown in Fig. 2.7(b) is used as a varactor, where the electrostatic attraction draws the metal bridge down to modulate the capacitance. Because current never flows through a semiconductor material, the equivalent series resistance of a MEMS switch/varactor can be very low, and this is especially true of the varactor. For example, the off-state capacitance and series resistance of one such device are quoted as 24 fF and 150 m Ω , respectively, resulting in a cutoff frequency of 44 THz [45], [46]. Other reported devices have series resistances in the range of 100–400 m Ω and off-state capacitances in the range of 20–60 fF, yielding a maximum cutoff of 80 THz [47]. These values are 2–4 times higher than the highest

value of 20 THz reported for a mesa InGaAs PIN diode [19], and one to two orders of magnitude higher than the cutoff frequencies of FET switches. Cantilever MEMS switches have lower cutoff frequencies due to their unavoidable ohmic loss and contact resistance, the latter being a strong function of the actuation voltage [48].

In general, the low series resistance leads to a direct improvement in the insertion loss of MEMS switches, although their isolation remains comparable to those using PIN diodes and FETs [44]. Although the actuation voltage is generally above 10 V for most MEMS devices (especially to close the metal-to-metal switch), techniques continue to be devised for lowering this level, often by adding extra mechanical complexity [49]. Other than the actuation voltage, no steady-state power is consumed during the operation of a MEMS device, meaning that the driver circuitry remains relatively simple, similar to that of FET switches. Another advantage to MEMS transducers is that they are inherently linear. Due to their mechanical motion, however, the switching time of MEMS is on the order of several microseconds [44], which is many orders of magnitude slower than their semiconductor counterparts. MEMS are typically constructed on silicon or quartz substrates, meaning have seldom been integrated with millimetre-wave monolithic GaAs circuitry.

2.3 Semiconductor phase shifters

Having introduced their various switching elements, semiconductor phase shifters are now presented. Many different circuit topologies exist, and these have been classified into six broad categories, including: switched line, switched filter, loaded line, reflection, combination, and vector modulator. In practice, sources that divulge all aspects of a phase shifter's performance are rarely encountered; instead, the shifter's advantages are usually

stressed and its shortcomings are omitted. As a result, the open literature contains a hodgepodge of figures, graphs, and metrics. In the literature search that follows, an attempt has been made to unify the performance metrics of a large number of published sources. As such, the compilation is, to the author's knowledge, the only comparison of phase shifters using standard performance criteria. Although a brief summary of phase shifter technology has been published in 1989, the reader is cautioned that several errors have been detected therein [50]. Given the extensive amount of published work on the topic, the scope will be limited to works that have been published between 1980–2000, and only for frequencies above 26.5 GHz. Technological improvements are inevitable with time, therefore the date of the publication and the technology used have been listed. Difficulties often arise when shifters are designed at higher frequencies or over large bandwidths, hence these values have also been included in what follows.

2.3.1 Switched line phase shifters

The switched line shifter uses a double-pole double-throw switch to select between two paths of differing lengths, as illustrated in Fig. 2.8.

Fig. 2.8 *Switched line phase shifter.*

In a dispersionless transmission medium, a difference in the time delay exists between the two branches, given by:

$$\Delta t = u_p (d_2 - d_1). \quad (2.11)$$

The resulting phase shift is a linear function of the frequency, as demonstrated by:

$$\Delta\phi = \beta(d_2 - d_1) = f \frac{2\pi}{u_p} (d_2 - d_1). \quad (2.12)$$

resulting in a true time delay (TTD) response. Most of the documented switched line shifters use FETs as their switching elements, which can either be arranged in series with the lines, as shown in Fig. 2.9(a), or in shunt with them, as shown in Fig. 2.9(b).

Fig. 2.9 Switched line phase shifter configurations: (a) series, (b) shunt.

The operation of the series layout is straightforward since the FETs in the desired branch are conducting whereas those of the other branch are not. In the shunt case, the switching FETs are placed $\lambda/4$ away from the T-junction to present a high impedance at the junction when the switch is shorting the line to ground, effectively turning off that branch. Although the shunt configuration consumes more area than the series configuration, it is quoted as being less sensitive to device parameter fabrication variations [51] and is capable of larger bandwidths since the off-state is achieved using the inherently large bandwidth of the low impedance FET state (quarter-wave transformed) [29].

Although the switched line shifter provides a true time delay (TTD), it has the disadvantage of being physically large, especially at lower frequencies and for high phase shift values. A variation to counter this problem is the use of an all pass filter or of a low pass filter in the delay arm [30], [52]. Switched line shifters operating at, and above, Ka-band are listed in Table 2.3.

Table 2.3 *Switched line phase shifter performance summary.*

reference, year, bits	frequency [GHz], bandwidth	phase error at f_0 [GHz]	phase error across the band	max. IL at midband and worst deviation across the band	technology, die size [mm]
[51] 1983 1-b shunt 180°	27.5–30 8.7%	0° at 30	< 20° of 16.7 ps	3 dB 3 dB/b Δ IL = 1 dB	1 μ m FET 2.7 $\times$ 3
[53] 1998 4-b series/shunt	33–36 8.7%	3.3° rms at 34.5	< 7° rms	17.2 dB 4.3 dB/b Δ IL = 4 dB	0.15 μ m pHEMT 2.2 $\times$ 2.5
[54] 1985 3-b series	30	N/A	N/A	6–8.7 dB 2–2.9 dB/b Δ IL N/A	N/A
[55] 1-b series 180°	43–45 6.8%	+10° at 44	+3° to +10° of 11.4 ps	3.8 dB 3.8 dB/b Δ IL = 0.5 dB	0.5 μ m FET size: N/A
[56], [57] 1992/1993 3-b series	43–45 6.8%	$\pm$ 5° at 44	7° of 19.9 ps	8.1 dB 2.7 dB/b Δ IL = 1.3 dB	0.2 μ m pHEMT 2 $\times$ 2.8
[58] 1997 4-b shunt	43–45 6.8%	< 20° at 44.5	N/A	12 dB 3.0 dB/b Δ IL = 4 dB	0.25 μ m pHEMT 1.5 $\times$ 3.5
[59] 1999 4-b shunt	33–36 8.7%	< 13.0° at 35	N/A	3 dB 0.75 dB/b Δ IL = 1.9 dB	45 V Si MEMS size N/A
<i>ibid.</i> 3-b shunt	<i>ibid.</i>	13.1° at 35	N/A	2.8 dB 0.7 dB/b Δ IL = 1.3 dB	<i>ibid.</i>
[60] 1989 4-b series	43–45 6.8%	< 3.3° at 44	N/A	11.2 dB 2.8 dB/b Δ IL = 2.2 dB	0.6 μ m FET 1.3 $\times$ 2.5

The results indicate that MEMS have a clear insertion loss advantage over FET switches, the former yielding an insertion loss of approximately 0.6 dB/b at 35 GHz, whereas the latter yield insertion losses ranging from 2.7 to 4 dB/b from 28–45 GHz. Note however that a 45 VDC bias was required for the MEMS as opposed to only 2 VDC for the pHEMT switches [56]. Although the phase accuracy of the MEMS shifter was very poor, it is expected to improve in subsequent design cycles. The figures of merit for the semiconductor implementations range from 28–60 °/dB whereas those of the two MEMS implementations are 113 °/dB for both the four- and three-bit shifters. As expected, the figure of merit decreases significantly as the number of bits increases.

Unfortunately, although the switched line shifter is one of the only configuration that provides a TTD, only three of the references provide any time delay performance data. Furthermore, no switching speeds are provided in the quoted works, except for the MEMS reference ($< 6 \mu\text{s}$). Noting the discrepancies of the insertion loss performance at 44 GHz, the progressive improvement of insertion loss appears to coincide with the device technology improvement: from 3.6 dB/b in 1985, to 2.6 dB/b using $.25 \mu\text{m}$ pHEMTs, and down to 0.2 dB/b using $0.2 \mu\text{m}$ pHEMTs. The highest insertion loss per bit of 3.8 dB corresponds to unresonated FET devices, which are known to have higher losses because of uncompensated parasitic capacitance [55]. As a final observation, as previously mentioned, the switched line shifters with the highest bandwidths do indeed assume the shunt configuration.

2.3.2 Switched filter phase shifters

The switched filter is a constant-phase type shifter that relies on the phase differential between a low pass filter (LPF) and a high pass filter (HPF) [61]. A LPF provides a phase lag, and is constructed of series inductors and shunt capacitors in either a tee or pi configuration. The phase advance of a HPF is produced by series capacitors and shunt inductors. The phase shift is achieved by switching between the two filters, as shown in Fig. 2.10.

Fig. 2.10 A switched filter (low pass/high pass) phase shifter.

When using simple pi or tee networks, it can be shown that the phase error will be maintained to within $\pm 2^\circ$ over an octave bandwidth for phase shifts of less than 90° . The reason for this phase stability is that the phase deviations of both filters track each other, although beyond 90° , a tradeoff exists between the achievable bandwidth and phase error [61]. These bandwidth limitations may be remedied by using more complex filter topologies or by cascading both high pass and low pass sections [61]–[66].

Oftentimes, when FET switches are used, their parasitic drain-source capacitance are incorporated into the filter design, thereby reducing the size of the circuit as well as the number of transistors required [62], [63], [65], [67], [68]. Since the parasitic capacitance of the FET switches are not resonated, the switched filter approach benefits from an

extended bandwidth over FET switched line shifters. Table 2.4 lists two switched filter millimetre-wave shifters.

Table 2.4 Switched filter phase shifter performance.

reference, year, bits	frequency [GHz] and bandwidth	max. phase error at midband, across the band, worst state variation	max. IL at midband and worst deviation across the band	technology, die size [mm]
[64] 1988 3-b	18–40 (76%)	$\pm 14^\circ$ $\pm 14^\circ$ 21° (rms: 2–9.8° ave. rms: 7.2°)	9.6 dB 3.2 dB/b $\Delta IL = 6.0$ dB $\Delta IL_{rms} = 0.4$ – 1.2 dB ave. $\Delta IL_{rms} = 0.97$ dB	0.5 μ m FET 1.3 $\times$ 2.1
[69] 1994 3 single bit chips (not cascaded)	30–36 (18%)	5° $-6/+10^\circ$ 15°	2.5 dB/b (typ) $\Delta IL = 1.8$ dB	12 $\times$ 12 μ m GaAs HBT PIN die size N/A

The phase and loss responses are reasonably stable over relatively large bandwidths when compared to the switched line configuration. The loss is comparable, being near 3 dB/b. The figure of merit of the first phase shifter in Table 2.4 is only 32.8 °/dB. The primary disadvantage of the switched filter shifter is that the design cycle can be more complex than other types because of the large number of filter configurations that are available. Further, some configurations yield component values that are difficult to realise, and performance tradeoffs also exist between the filter configurations for different phase shift values [64], [68]. The switched filter has not been demonstrated above 40 GHz, which is likely due to the fact that high quality inductors are difficult to synthesise at these frequencies.

2.3.3 Loaded line phase shifters

The loaded line phase shifter, illustrated in Fig. 2.11, has been studied extensively over the past four decades, beginning with a seminal paper in 1965 [70] and culminating in a comprehensive design overview in 1985 [71].

Fig. 2.11 *Shunt loaded line phase shifter.*

The shifter consists of two equal shunt admittance elements Y_i that are separated by an electrical length θ of transmission line of characteristic impedance Z_c . The overall circuit is matched to the system's characteristic impedance of Z_0 . The admittance elements Y_i are switched simultaneously between two states ($i = 1, 2$) by suitable solid state devices to provide the insertion phase shift. By duality, series reactance elements could also be used [68], however the shunt configuration has dominated the literature since only a fraction of the power circulates through the shunt elements.

The characteristic impedance of the line section Z_c , its electrical length θ , and the values of the admittance elements Y_i are chosen in such a way as to yield the desired insertion phase shift $\Delta\phi$ while maintaining a low return loss. Under the assumptions of lossless elements ($Y_i = jB_i$) and a perfectly matched impedance, the required form of the s_{21} parameter is [71]:

$$s_{21} = \frac{1}{(\cos \theta - B_i Z_c \sin \theta) + j Z_c Y_0 \sin \theta} \quad (2.13)$$

where Y_0 denotes the characteristic system admittance ($1/Z_0$). The form of s_{21} is $\frac{1}{a + jb}$

and thus its phase angle is given by $\arctan \frac{-b}{a}$. The magnitude of s_{21} must be unity under lossless, perfectly matched conditions. A useful visualisation aid depicting the phasor's operation is a vector diagram of the s_{21} , which is easily constructed using the resulting s_{21} phase equations:

$$\left. \begin{aligned} \cos \phi &= \cos \theta - B_i Z_c \sin \theta \\ \sin \phi &= -Z_c Y_0 \sin \theta \end{aligned} \right\} \quad (2.14)$$

During state switching, the quadrature component of s_{21} remains invariant while the in-phase component varies as a function of B_i . The resulting vector diagrams are depicted in Fig. 2.12 and demonstrate three distinct classes of operation [71], [72].

Fig. 2.12 Phasor diagrams of s_{21} showing the three classes of operation of a loaded line phase shifter.

In the absence of shunt elements, the transmission line section merely provides a constant phase delay of θ , as illustrated by the dashed vector in Fig. 2.12. The addition of the *capacitive* shunt elements further delays the phase when $\theta < 90^\circ$. Fig. 2.12(a), Class I, is defined when the susceptance values of the two states are unequal. Fig. 2.12(b) illustrates Class II operation, where one of the susceptance states B_i is zero. Such a configuration is

also called load/unload. Fig. 2.12(c) illustrates Class III, or complex conjugate loading, where $B_1 = -B_2$ and θ is forcibly equal to 90° to maintain the magnitude of s_{21} at unity.

A myriad of circuit topologies and components have been proposed to synthesise the required shunt elements and there does not appear to be any marked performance advantage to using one type over another. Among the shunt components that can be chosen are: lumped elements and PIN diodes [61], [70], [73], varactor diodes [74], line-transformed switch impedance elements [71], [73], [75]–[77], switched length stubs [70], [72], [73], and switched stub topologies providing conjugate susceptance values [71], [72]. These shifters generally provide a phase response that has a single ripple at the centre frequency. As such, depending on the relative frequency of operation, the loaded line shifter can provide either a constant phase shift or a quasi-constant time delay as a function of frequency [61], [70], [74]–[76], [78].

Another important figure of merit is the insertion loss variation for different phase states. It has been shown analytically that when lossy switching elements are used, and if the admittance susceptance-to-conductance ratio B/G_i is constant, the insertion loss variation is suppressed only when $\theta = 90^\circ$ [71]. When this ratio condition is not maintained for both states, small theoretical insertion loss variations on the order of 0.1 dB can be expected for stub admittance values when $\theta = 90^\circ$ [73].

In general, the magnitude of the required susceptance B_i is commensurate with the required phase shift $\Delta\phi$. Large phase shifts require large susceptance values: should the susceptance be composed of say, a capacitor, then a larger capacitance would obviously be required. However, as the capacitance increases, so does the susceptance variation as a function of frequency via the equation $B = \omega C$. As such, the bandwidth of loaded line

shifters decreases rapidly as a function of $\Delta\phi$. For example, the bandwidth, defined as the range of frequencies over which $\text{VSWR} < 1.2$ and the phase error is within $\pm 2^\circ$, is typically around 20% for $\Delta\phi = 22.5^\circ$ but drops to only 10% for $\Delta\phi = 45^\circ$ when lossy switches are used [73]. For this reason, the unit loaded line phase shifter is typically designed to provide a phase shift of no more than 45° . Although unit shifter cells can be cascaded to provide larger phase shifts, other topologies are often chosen to provide the larger phase bits [54], [76], [79]–[85]. The performance of phase shifters that use combinations of topologies will be given in Section 2.3.5. Table 2.5 lists single-bit loaded line phase shifters operating above 30 GHz.

Table 2.5 *Performance of single-bit loaded line phase shifters.*

reference, year, value	frequency [GHz] and bandwidth	max. phase error at midband, across the band	max. IL at midband and variation across the band	technology
[86] 1989 11.25°	79–92 15%	0° $\pm 1.3^\circ$	(mean ave. IL = 0.9 dB) $\Delta\text{IL} = 1.0$ dB	PIN HMIC
[84] 1987 22.5°	34–36 GHz 5.7%	+4.5° +2 to +5°	0.4 dB $\Delta\text{IL} = 0.25$ dB	Si PIN HMIC
<i>ibid.</i> 45°	<i>ibid.</i>	+4° +4 to +7°	0.7 dB $\Delta\text{IL} = 0.3$ dB	<i>ibid.</i>

Loaded line shifters have been implemented for phase bits no larger than 45° , and although this configuration is theoretically capable of greater bandwidths, the cited ones are no greater than 15%. The loss per bit is quite low, ranging from 0.3 to 0.9 dB up to 100 GHz, even though discrete *silicon* PIN diodes are used. The low loss is due directly to the fact that only light line loading is required, meaning that little power is actually flowing through the lossy switching devices. The loaded line may therefore handle relatively high power levels, as evidenced by the 29.5 W of CW power used in [84]. When the loss is expressed in $^\circ/\text{dB}$, the best value is only $45/0.7 = 64$ $^\circ/\text{dB}$.

Another configuration that falls under the loaded line umbrella is the distributed transmission line. By controlling the periodic shunt capacitance, the phase velocity of a transmission line can be modulated, thereby forming a TTD phase shifter. This technique has recently been demonstrated up to 60 GHz using MEMS varactors to periodically load a CPW line [45], [46], [87], [88]. As opposed to their switched-line TTD counterparts, the periodically loaded lines have very wide bandwidths, theoretically ranging from DC up to the nonlinear point of their response. Unfortunately, appropriate TTD metrics have not been disclosed in those references, making their phase error evaluation very difficult. The available data has been listed in Table 2.6.

Table 2.6 Performance of periodically loaded line phase shifters.

reference, year	frequency [GHz]	available phase shift	max. IL at midband	technology, line length
[45] 1998	40	82°	1.8 dB	22V MEMS bridges on quartz 10 mm
	60	118°	2.0 dB	
<i>ibid.</i>	40	74°	1.4 dB	5.5 V MEMS bridges on quartz 6.7 mm
[88] 1999	40	80°	1.1 dB	13 V MEMS bridges on quartz 7.6 mm
	60	146°	1.6 dB	

Although the results are very promising, it is difficult to surmise how quartz substrates can be integrated with other millimetre-wave circuitry. The insertion loss variation with respect to frequency and state is very small for these phase shifters, to the point where it is difficult to assess it by visual inspection of the presented graphs [88]. The figures of merit are also good, ranging from 60–90 %/dB.

2.3.4 Reflection phase shifters

Of the phase shifters that have been reviewed up to this point, all have consisted of two-port networks. One-port phase shifting networks, exploiting the phase of their reflection coefficient, can also be used as phase shifters and are referred to as *reflection phase shifters*. For specific applications such as reflectarrays [89], a single reflection network suffices [26]; however, two-port applications require the reflection network to be connected to a circulator or to a quadrature hybrid coupler, as shown in Fig. 2.13.

Fig. 2.13 Two-port reflection phase shifters: (a) using a circulator, (b) using a quadrature hybrid coupler.

In the reflection phase shifter of Fig. 2.13(a), the incident power is directed toward the reflection network via the circulator, and the reflection is then redirected toward the phase shifter's output. Hence the phase of the reflection coefficient Γ determines the output phase. A phase shift is achieved when the reflection network's impedance changes states. Ideally, the magnitude of the reflection coefficient is unity, for any absorption by the reflection network results in an insertion loss of $1 - |\Gamma|^2$. The reflection loss of the shifter is controlled solely by that of the circulator, allowing the reflection network to be independently optimised in terms of bandwidth, phase stability, and loss, an advantage that is not found in other phase shifter configurations.

Factors such as cost, size, and loss have discouraged the use of the circulator. Instead, hybrid couplers, amenable to planar integration, have superseded their use, as illustrated in Fig. 2.13(b). Although initially appearing more complex, the operation of this configuration is the same as that using a circulator. However, two *identical* reflection networks are required instead of only one. Although this can be viewed as a disadvantage of the hybrid configuration, it also increases the power handling of the circuit. Various types of quadrature hybrids have been used in reflection phase shifters. Narrowband couplers such as the branch line [6], [38], [90] have been demonstrated, as well as wideband couplers such as the lumped element [91], [92], Lange [11], [12], [80], [93]–[95], and other interdigitated couplers [96], [97]. Ratrace 180° hybrids can also be used [15:p.439], [84]. The hybrid coupler must be carefully designed when used in a reflection shifter because any intervening mismatch between its ports can introduce significant phase errors [15:pp.443–446], [84].

The type of reflection network depends largely on the shifter's desired characteristics. Both analogue and digital networks offering either true time delay or constant phase have been successfully developed, and are presented in the following two subsections.

2.3.4.1 Analogue reflection phase shifters

Unlike digital shifters, the available phase resolution of an analogue shifter depends primarily on the number of bits in the driver's D/A converter. Therefore, higher levels of resolution do not result in significant increases in circuit complexity or insertion loss. The analogue shifter may also be made smaller than the digital one, resulting in higher yield and lower costs. The variable reactance elements required by analogue shifters are provided by reverse-biased Schottky varactor diodes. In addition to a stable insertion loss and a controlled (constant or linear) phase shift as a function of frequency, desirable features of

analogue shifters are a 360° phase coverage, and a linear phase variation with respect to bias voltage, the latter property simplifying the shifter's driver. Although attempts have been made to either stabilise the phase [6], [12], [97], [98] with respect to frequency, or to linearise the phase with respect to bias [99], these single goals are usually achieved to the detriment of the other desirable characteristics. A comprehensive design technique has also been proposed to simultaneously provide a stable insertion loss, full 360° of phase coverage, and a linear phase with respect to bias [95], however, the resulting phase versus frequency is nonlinear [100], as it usually is when it does not figure in the design cycle [93], [96].

For most analogue reflection shifters not designed for flat phase response, the maximum phase variation occurs at the maximum reverse bias voltage level. The variation can be reduced by reducing the bias, but at the expense of the phase coverage. It is apparent that analogue reflection shifters are subject to tradeoffs between their desired characteristics. When these shifters are designed to be wideband, their insertion loss characteristics are compromised. For the rest, there are clear compromises to be made between phase coverage and phase stability, and, phase stability and bandwidth. No analogue reflection shifter has simultaneously demonstrated low insertion loss, low insertion loss variation, low phase variation, complete phase coverage and linear bias response. Table 2.7 lists the only analogue reflection phase shifter operating above 26 GHz.

Table 2.7 *Performance of an analogue reflection phase shifter.*

reference, year, coupler	frequency [GHz] and bandwidth	phase shift at midband, max. phase variation across the band	max. IL at midband and IL variation across the band	technology, die size [mm]
[101] 2000 Lange	46–54 16%	100° 8°	4.5 dB $\Delta\text{IL} = 1.6$ dB	GaAs pHEMT varactors 1.3×1.3

Although the phase shifter does not offer a 360° coverage, its phase stability across the band is very good. The figure of merit for this phase shifter is poor, only $22.2^\circ/\text{dB}$.

2.3.4.2 Digital reflection phase shifters

Many types of digital reflection networks have also been proposed, with the most popular networks consisting of lumped elements [85], switched lines [11], [90], and impedance transformers [7], [80], [84], [102], although other elements, including loaded-lines, have also appeared in the literature [26]. Although a true time delay is expected from the switched line length shifter, this has only been approximately demonstrated in [11]. In general, the phase response of digital reflection networks is also nonlinear as a function of frequency and the insertion loss varies as a function of the state. However, compensation techniques have again been proposed for lumped element networks to simultaneously stabilise the phase and to provide constant insertion loss [15:pp.446–453], [102]. Table 2.8 lists the digital reflection phase shifters operating at, or above Ka-band.

Table 2.8 *Performance of digital reflection phase shifters.*

reference, year, bits, coupler	frequency [GHz] and bandwidth	max. phase error at midband, across the band, max. state variation	max. IL at midband and worst variation across the band	technology, die size [mm]
[84] 1987 1-b 90° 180° ratrace	34–36 5.7%	$+11^\circ$ $+10$ to $+16^\circ$ 6°	1.25 dB 1.25 dB/b $\Delta\text{IL} = 0.8$ dB	Si PIN MIC
[90] 1987 3-b branch line	60–65 8%	4° $+2/-39^\circ$ ($2.5 - 15^\circ$ rms) 39°	9.9 dB 3.3 dB/b $\Delta\text{IL} = 5.5$ dB	switched lines GaAs Schottky MMIC 1.9×3.2
[26] 1999 3-b no coupler (reflectarray)	~ 32 – 34 6%	$\sim 20^\circ$ (9.5° rms at midband) N/A	3.9 dB 1.3 dB/b $\Delta\text{IL} = 2.3$ dB	40 – 70 μm diameter GaAs mesa PIN MMIC 2.5×3.5

Comparing [26] to [84], it is apparent that modern monolithic PIN diodes are capable of producing the same performance as discrete ones, with insertion losses on the order of 1 dB/b being achievable. The reflectarray phase shifter of [26] has a switching time of 50 ns and sustains 1 W of continuous power, whereas the discrete implementation [84] sustains 29 W of continuous power. The figures of merit of the PIN diode implementations, ranging between 72–81 °/dB, are better than the switched line version, which is only 32 °/dB.

2.3.5 Combination phase shifters

A number of digital phase shifter designs use a cascade of different phase shifter topologies in an attempt to combine the advantages of each type of phase shifter [54], [66], [79], [82], [83], [94], [103], [104]. For instance, loaded line shifters yield lower losses for small phase bits, hence this type may be incorporated into a multi-bit design for the least significant bits. Such a strategy may also be employed when several competing design goals must be met, such as a minimal circuit size, low losses, and large bandwidth. Table 2.9 lists the pertinent phase shifters.

Table 2.9 *Performance of combination phase shifters.*

reference, year, bits	frequency [GHz] and bandwidth	max. phase error at midband, across the band, max. state variation	max. IL at midband and worst variation across the band	technology, die size [mm]
[83] 1989 4-b	32 0%	"7% mean error"	16 dB 4 dB/b Δ IL = 8.5 dB	0.25 μ m GaAs FET MMIC N/A
[54] 1985 4-b	27.5–30 8.7%	(illegible)	10.8 dB 2.7 dB/b Δ IL = 5 dB	GaAs FET MMIC N/A
[103] 1998 5-b	43.5–45.5 4.5%	N/A +4/–4.5° 4°	9 dB 1.8 dB/b Δ IL = 2 dB	0.25 μ m pHEMT 2.7 $\times$ 7.6
[104] 1993 4-b	35–37 5.6%	+3/–7° +3/–9° N/A	14 dB 3.5 dB/b Δ IL = 3.6 dB mean av. IL = 12.8 dB	GaAs FET MMIC 3 $\times$ 7.8

The total loss through these phase shifters is extremely high, yielding figures of merit of only 21–39 %/dB. The three four-bit FET shifters all have similar losses, ranging from 3 to 4 dB/b. The newest technology in this table has the lowest loss, 1.8 dB/b at 44 GHz [103].

2.3.6 Vector modulators as phase shifters

A vector modulator dynamically alters both the phase and amplitude of a signal, thereby imparting any modulation data to a given carrier. In a broader sense, the phase shifters that have been described may also be categorised as modulators in that they modulate the phase of a unit-amplitude vector. By cascading a variable attenuator to a bank of shifters, a complete digital vector modulator may be formed, as illustrated in Fig. 2.14.

Fig. 2.14 *A tandem attenuator/phase shifter vector modulator.*

The drawbacks of this tandem vector modulator, also known as an “alpha-phi” configuration, are the accumulation of cascaded phase errors and an increase in physical size that is commensurate with the required number bits—the same disadvantages of any multi-bit phase shifter. Although this may be alleviated by using an analogue shifter, full 360° phase coverage is often difficult to achieve, requiring cascaded cells or large voltage drive swings. Another analogue vector modulator configuration that alleviates these problems is the IQ vector modulator, shown in Fig. 2.15.

Fig. 2.15 *Block diagram of the IQ vector modulator.*

The quadrature hybrid at the input of the circuit serves to split the incoming signal into two orthogonal channels, imparting the name to this vector modulator. By controlling the attenuation of each of the channels, their sum results in a vector of any amplitude and of

any phase within the range of 0–90°. The addition of the biphase 0°/180° (BPSK) shifters allows the vector to be moved into any of the four quadrants.

The IQ vector modulator may also serve as a pure phase shifter if the output amplitude is maintained constant. Care must be taken in designing the elements of the modulator as any phase or amplitude imbalance between the I and Q arms will result in AM-PM distortions [105]. Only over the past decade has technology allowed the direct modulation of carriers much above X-band to be feasible. Prior to this, only heterodyne modulation schemes were available. Given their compact size, performance, and simple multi-bit operation, it is highly likely that these modulators will gain popularity in many phase shifting applications. More recent implementations of the IQ modulator use reflection circuits in both of the arms to provide attenuation and quadrant control simultaneously, thereby eliminating the biphase phase shifters. These modulators require analogue control signals and hence a D/A converter must precede them; as such, they technically qualify as analogue shifters. Unfortunately, measurement procedures for vector modulators depend heavily on the application. In the references that are quoted, the vector modulators are intended for various applications including digital QAM modulation, serrodyne, and signal cancellation. They are principally evaluated in the context of their application, and the resulting lack of information thereby makes phase shifter characterisation difficult. Only IQ modulators have been implemented above Ka-band, and pertinent results are listed in Table 2.10.

Table 2.10 *Performance of IQ vector modulators.*

reference, year, bits	frequency [GHz] and bandwidth	max. phase error across the band	IL at midband and IL variation across the band	technology, die size
[106] 1998 256 QAM 8-b	38 0%	$\pm 2\%$ for the constellation	N/A	0.25 μm pHEMT 1.6 $\times$ 2.0 mm
<i>ibid.</i>	55–65 16.7%	$\pm 2^\circ$ for the constellation	ave.: 5 dB (0.8 dB/rad) (0.6 dB/b) $\Delta\text{IL} = 1$ dB (one state only)	0.25 μm pHEMT size: N/A
[107] 2000 64 QAM 6-b	75.5–77.5 1.3%	N/A	min.: 12 dB (1.9 dB/rad) (2.0 dB/b) $\Delta\text{IL} = 0.2$ dB (one state only)	0.25 μm pHEMT 1.6 $\times$ 2 mm
[108] 2000 16 QAM 4-b	38 0%	“negligible phase errors”	max.: 14 dB 2.2 dB/rad 3.5 dB/b $\Delta\text{IL} = 6.5$ dB (four states only)	HBT 2.9 mm ²

256 QAM has 196 distinct phase states whereas 64 QAM has 52 phase states. For sake of comparison, assume that they are equivalent to 8-bit and 6-bit shifters, respectively.

The lookup table that is used with analogue IQ vector modulators gives them an excellent phase error performance. Losses are often inadequately cited in the references, although the little information that is disclosed seems to indicate that the total loss through this type of phase shifter is in fact quite high, as is the insertion loss variation as a function of state. When expressed in terms of dB/b, the losses appear reasonable, however the figures of merit for the last two phase shifters in Table 2.10 are a dismal 26 and 30 °/dB, respectively.

2.3.7 Semiconductor phase shifter summary

The primary conclusion that may be drawn from the preceding tables is that the insertion loss of semiconductor monolithic phase shifters is very high in the millimetre-

wave frequency range. Of the monolithic two-port shifters offering at least 300° of phase shift, the range of the average maximum insertion loss values at midband is 7.4–16 dB, yielding an average value of 10.9 dB. Even if one only considers state of the art pHEMT and HBT devices, the average maximum insertion loss at midband is still approximately 9 dB, which means that only one eighth of the input power emerges from the output. Hence there is a severe insertion loss penalty incurred when opting for a monolithic phase shifter implementation. Overall, semiconductor phase shifter figures of merit are in the 20–60 °/dB range, although this improves to 70–80 °/dB when PIN diodes are used. The performance of phase shifters using MEMS architectures is superior, providing losses of 1–3 dB range and figures of merit of 60–110 °/dB. However, until recently, MEMS have seldom been realised on high performance substrates such as GaAs [109], [110], meaning that only HMIC phase shifter configurations have been possible. Furthermore, the performance of the MEMS phase shifters and of the overwhelming majority of the others have been quoted in their unpackaged state. The packaging of the circuits will inevitably further degrade the insertion loss. As will be seen in the next section, the figures of merit of ferrite waveguide phase shifters are far superior to those using semiconductors.

2.4 Ferrite material and ferrite millimetre-wave phase shifters

In this section, the nature of ferrimagnetism is introduced, as is the interaction between a ferrimagnetic material and a microwave field. Next, phase shifter configurations exploiting the gyrotropic behaviour of the ferrite are presented and their performance is compared at millimetre-wave frequencies.

2.4.1 Ferrimagnetism, gyrotropic interaction, and ferrite preparation

Although magnetism originates from several distinct sources in materials, *uncompensated electron magnetic spin moments* produce the strongest macroscopic effects [111]. Although quantum mechanics views particles, including electrons, as zero-volume points, electrons nonetheless possess an inherent magnetic field due to their spin motion. Another theory of atomic structure views the electron as a spinning ring of charge, from which its magnetic field explicitly originates [112], [113]. Regardless of the model chosen, all electrons have a net magnetic field, or moment m , and an angular momentum about them, as illustrated in Fig. 2.16(a).

Fig. 2.16 Model of (a) a spinning electron and its magnetic moment, (b) when exposed to an external magnetic bias field (dashed ellipse is the precession motion).

In most materials, the magnetic moments tend to cancel each other such that no net magnetic effect is observed on the macroscopic scale. In ferro- and ferrimagnetic materials, a large number of uncompensated spins exist, with more in the former than in the latter. These uncompensated moments may be aligned under the influence of an external magnetic field H_0 . When the bias field and moment are not in alignment, the Lorentz force law dictates that a torque will be exerted on the electron such that, if it were not spinning, it would fall into alignment with the bias field. The angular momentum of

the electron prevents this, instead causing the electron to precess about the bias axis, causing its moment to rotate along the dashed line in Fig. 2.16(b) [114]–[116]. Losses associated with this movement eventually cause the moment to spiral into exact alignment with the bias. The term *gyrotropic* refers specifically to this type of rotational motion.

If the electrons are exposed to an additional microwave (propagating) field, the time-varying magnetic field will perturb the aligned electron moments such that they move off-axis again, repeating their precession. This field-matter interaction is responsible for modulating the phase velocity of the microwave signal, resulting in a controllable phase shift that is commensurate with the magnitude of the bias field.

Certain elements have more uncompensated electron spins than others, and this is particularly true of the *transitional metals*, having atomic numbers 21–29: scandium, titanium, vanadium, chromium, manganese, iron, cobalt, nickel, and copper. These magnetic metals cannot be used directly because their low resistivity would incur severe losses at microwave frequencies. However, when these metals are used in a ceramic crystal, their conduction electrons are bound by covalent bonds with oxygen. The resulting ceramic has a very high resistivity and is therefore suitable for use in microwave applications. When inside the crystal, only some of the unbalanced electron magnetic moments cancel, leaving a net ferrimagnetic effect, hence the name *ferrite*. Other non-magnetic elements such as lithium, magnesium, and zinc can be added to the crystal to enhance the ferrimagnetic effect. Although ferrites may assume *garnet* and *hexagonal* lattices, the *spinel* cubic lattice is the most common ferrite crystal.

Single crystal ferrites may be grown in a laboratory, although bulk ferrite parts are usually made from polycrystalline material [117]–[120]. Similar to any other ceramic, ferrite is prepared by blending appropriate proportions of metallic oxides that are then pre-sintered (calcined) at approximately 800 °C, during which thermal agitation intimately diffuses the various metallic ions. After calcination, the powder is formed under high pressure into the desired final shape (toroid, bar, etc.), called a *green body*. If liquid additives are combined with the powder, the resulting *slurry* can also be cast into sheets (green ferrite tape). After the green body is formed, it is sintered (or *fired*) at approximately 1000 °C, promoting the packing and re-ordering of the elements into their final polycrystalline state.

2.4.2 Dual-mode ferrite phase shifters

The dual-mode waveguide phase shifter is a reciprocal implementation of a Faraday rotation phase shifter [121]. As depicted in Fig. 2.17, the phase-shifting core of the device consists of a length of circular or square ferrite-filled waveguide that is exposed to a longitudinal magnetostatic bias provided by a solenoidal winding [122]–[127].

Fig. 2.17 Structure of a dual-mode waveguide ferrite phase shifter: (a) longitudinal cross-section, (b) quadrupole magnetostatic transverse bias in the polarisers. Areas in grey are made of ferrite material.

Although the core is nonreciprocal, the complete phase shifter is rendered reciprocal by the two non-reciprocal polarisers that convert the input/output linear-polarised fields to the circular-polarised field required in the core. The polarisers are implemented as circular or square ferrite-filled guides having a quadrupole transverse bias that can be supplied by additional bias windings or by an external magnet. Mode suppressors, consisting of resistive cards (or *vanes*), are used at the output to extinguish any cross-polarisation left by the polarisers.

Since its inception in 1970, the dual-mode phase shifter has remained as one of the few devices capable of providing truly excellent performance over the millimetre-wave band. Unlike semiconductor phase shifters, very few competing designs are found in the literature; instead, the same originating authors periodically update the state of the art. Dual-mode phase shifters operating at 35 GHz are approximately 3.5 cm long and can provide phase shifts of up to 430° with only 0.8–1.0 dB of loss, yielding figures of merit in the range of 430–540 %dB [127], [127], which are far superior to even those of MEMS

phase shifters, which are near 100 °/dB. The performance of a 60 GHz dual-mode phase shifter is listed in Table 2.11 [127].

Table 2.11 *Performance of a 60 GHz dual-mode ferrite waveguide phase shifter.*

reference, year, length, core diameter	frequency [GHz] and bandwidth	phase coverage (phase error N/A)	max. IL and worst IL variation over the band	ferrite material, saturation magnetisation, switching time
[127] 1982 3 cm 1.57 mm	60–63 4.9%	450° sustained 363° remanent	IL = 2.2 dB Δ IL = 0.7 dB	MgMn $B_s = 300$ mT 12 μ s

Unfortunately, no phase variation information is provided for the shifter, although it is typically very stable for ferrite devices, as is the insertion loss variation. Using the worst insertion loss over the quoted bandwidth, the figure of merit is 205 °/dB. The saturation magnetisation for the magnesium-manganese ferrite is low, $B_s = 300$ mT, although the ferrite was chosen for its good squareness and low dielectric loss instead.

A millimetre-wave nonreciprocal implementation of the core, consisting of a longitudinally-biased ferrite-filled circular waveguide, has also been reported [128]. Using nickel-zinc ferrite with an impressive saturation magnetisation of 510 mT, this device provides 150 °/cm of phase shift at 70 GHz with a corresponding figure of merit of 257 °/dB and can be used over the 60–90 GHz range.

2.4.3 Ferrite slab-loaded rectangular waveguide phase shifters

A ferrite slab-loaded phase shifters consists of a rectangular waveguide that has at least one slab of ferrite running down its longitudinal length [121]. Large phase shift values are obtained when two full-height, oppositely transverse-biased slabs are used, as illustrated in Fig. 2.18(a).

Fig. 2.18 *E-plane cross sections of ferrite slab-loaded rectangular waveguide phase shifters: (a) dual-slab, (b) dielectric loaded, (c) dual toroid.*

To maximise the phase shift, the coupling between the ferrite and the guided mode should also be maximised, which is accomplished by controlling the slab thickness and by placing it in locations where the propagating magnetic field is circularly polarised [129]. Higher phase shift values are obtained when a dielectric is placed between the slabs, as illustrated in Fig. 2.18(b), which concentrates the energy in the centre of the guide. As shown in Fig. 2.18(c), toroids are often used to implement the slabs, which reduce the reluctance of the magnetostatic circuit by providing closed magnetic flux paths.

Although slab-loaded phase shifters perform excellently at millimetre-wave frequencies, their assembly, as well as the production of their toroids, are complicated by the very small dimensions and low tolerances involved. For instance, Table 2.12 lists the required aperture width and wall thickness values for toroids at various millimetre-wave frequencies [130].

Table 2.12 *Toroid dimensions for millimetre-wave ferrite slab-loaded waveguide phase shifters.*

frequency [GHz]	aperture width [μm]	wall thickness [μm]
35	254	508
65	152	279
94	127	165

The required cross sectional dimensions are similar to those of the eye of a needle, requiring costly fabrication techniques [130], [131]. However, due to their unsurpassed performance, this has not deterred their deployment in military applications. Given the sensitivity of the data, few recent articles have appeared in the open literature. Table 2.13 lists the performance of a few reported ferrite slab-loaded phase shifters.

Table 2.13 *Performance of ferrite slab-loaded rectangular waveguide phase shifters.*

reference, year, switching time	frequency [GHz] and bandwidth	phase coverage	insertion loss	length [cm]
[127] 1989 1 μ s	35	360°	max. 0.9 dB	1.78
	44	360°	max. 1.0 dB	2.29
	60	360°	max. 1.6 dB	2.29
[132] 1978 < 1 μ s	35 (1.7%)	422°	max. 0.7 dB	3.43
	65 (1.5%)	87°	max. 0.9 dB	0.79
	94 (1.1%)	38°	max. 1.5 dB	0.64
[133] 1996 N/A	34.5–35.5	N/A	ave. 0.6 dB,	N/A
	2.9%	max. error: 5° max. variation across the band: 2°	max. variation of average: 0.1 dB	

The most recent reference in Table 2.13 is a four-bit latching phase shifter that uses lithium-zinc ferrite having a saturation flux density of 460 mT and a squareness ratio of 0.83. The quoted figure of merit for this phase shifter is very high, 750 °/dB [133]. The other figures of merit at 35 GHz are 400 and 600 °/dB for the first two references of Table 2.13, respectively.

2.4.4 Reggia-Spencer and other suppressed rotation phase shifters

A Reggia-Spencer phase shifter consists of a longitudinal ferrite bar that is suspended in the centre of a rectangular waveguide [121]. Contrarily to the slab-loaded shifter, the bias is applied in the longitudinal direction, as illustrated in Fig. 2.19(a).

Fig. 2.19 *Suppressed rotation ferrite phase shifters: (a) Reggia-Spencer, (b) ferrite dielectric waveguide. The $\otimes$ symbolises longitudinal magnetostatic bias into the page.*

The device acts as a reciprocal phase shifter when, in addition to the dominant quasi- TE_{10} mode, an additional cutoff hybrid mode is also excited in the guide. This additional mode is susceptible to Faraday rotation, but being just below cutoff, couples to the quasi- TE_{10} instead, imparting significant phase shift to the dominant mode [134]. Millimetre-wave Reggia-Spencer phase shifters are available for frequencies covering 26.5–110 GHz [135]. Offering full 360° phase coverage over 2% bandwidths, these devices produce 2.5–3.0 dB of insertion loss, resulting in figures of merit of 120–144 $^\circ/\text{dB}$.

Noting that the energy tends to concentrate in the middle of the guide for slab-loaded and Reggia-Spencer phase shifters, *ferrite dielectric waveguides* have also been devised for use in the millimetre-wave range, as illustrated in Fig. 2.19(b). Similar to dielectric waveguides, the guide consists of a single rectangular slab of ferrite with no metallic walls, exciting a propagating field distribution in its core and an evanescent one external to it. When the bias is applied longitudinally, a suppressed-rotation effect occurs, similar to the one found in Reggia-Spencer phase shifters [136]. Table 2.14 lists the performance of such components operating in the millimetre-wave band.

Table 2.14 *Performance of ferrite dielectric waveguide suppressed rotation phase shifters.*

reference, year, cross section, length	frequency [GHz] and bandwidth	phase shift	insertion loss	material, saturation magnetisation
[136] 1982 1.4 × 1.9 mm 1.27 cm	35	160°	0.5 dB	NiZn 200 mT
[137] 1998 1.3 × 2.6 mm 4.6 cm	27–36 29%	< 420°	0.4–0.5 dB	NiZn 500 mT

The figure of merit of the first shifter is 320 °/dB whereas that of the second is very quite high, 840 °/dB. The ratio of the figures of merit is approximately that of the saturation magnetisation, so the very high figure of merit is in fact plausible. Unfortunately, both references give little information regarding phase error, switching time, and insertion loss variation. The ferrite dielectric waveguide phase shifter is attractive due to the simplicity of its construction as compared to other metallic waveguide implementations.

2.4.5 Planar ferrite phase shifters

Planar phase shifters consisting of transmission media with ferrite substrates have also been demonstrated, but not at millimetre-wave frequencies. Although they are promoted as being more compatible with other planar circuitry, this is only partially true since ferrite substrates are not very common; further, any integration is restricted to HMIC implementations since ferrite films are rarely available for MMICs. Some implementations require bias fields that are perpendicular to the substrate surface, which is an inherently hard direction of magnetisation. To counter this, hexagonal ferrites, being anisotropic, may be used because they offer an easy axis of magnetisation [138]. However, they require the substrate to be made from single crystal ferrite, implying that

the substrate must be specially grown, cleaved, and polished prior to use, an inherently expensive procedure [139], [140]. For frequencies below Ka-band, the following figures of merit have been demonstrated: 250 °/dB for slot line and *microstrip slotline*, 180 °/dB for coplanar waveguide, 150 °/dB for meander line, and 156 °/dB for microstrip [139], [141]. Given that these figures will only be worsened by higher frequencies, it seems likely that all quasi-TEM phase shifters will yield approximately the same performance as MEMS phase shifters currently do in the millimetre-wave range, with values on the order of 100 °/dB.

It is worth noting that a 6-cm long planar phase shifter has been realised using LTCC ferrite tape up to 26.5 GHz [142], [143]. Unfortunately, the references are primarily concerned with the modelling of the device, and very little measured data are disclosed.

2.4.6 Ferrite phase shifter summary

The figures of merit of good ferrite waveguide phase shifters, being generally higher than 500 °/dB in the millimetre-wave range, are several times greater than those of their semiconductor counterparts. Whereas semiconductor shifters suffer from losses near 10 dB, ferrite ones rarely introduce more than a few decibels of loss. Amplification at millimetre-wave frequencies being an expensive commodity, ferrite phase shifters offer a clear performance advantage over all semiconductor designs.

Due to the necessary inductance of the bias windings, ferrite phase shifters will always have much slower switching times than those of semiconductor shifters. Ferrite and MEMS switching times range from less than 1 μ s to tens of microseconds, values that are three to four orders of magnitude larger than those of FETs, which have

switching times on the order of 0.1 ns. PIN diode switching times are in the 1–20 ns range.

The primary disadvantages of ferrite phase shifters are related to cost, size, and integration. The fabrication of small electromagnets and tiny ferrite toroids, and the assembly of these, are time consuming and expensive. When compared to semiconductors, ferrite devices are physically large and heavy. Furthermore, their three-dimensional nature makes them impractical for integration unless planar designs are used. Even so, the ferrite materials themselves are often not readily available. Hence millimetre-wave phase shifters, in general, appear to suffer from an inherent incompatibility between high performance and integration.

2.5 An introduction to multi-chip module packaging technologies and low temperature cofired ceramic

Low temperature cofired ceramic (LTCC) is but one of the many chip-packaging technologies that can be used for the fabrication of multi-chip modules (MCMs). An MCM refers to any single package, usually consisting of a multilayer stack, that is able to house multiple bare-die integrated circuits (ICs), packaged ICs, and passives. This contrasts with standard printed circuit board (PCB) techniques, such as surface mount technology (SMT), in which only individual *packaged* ICs and passives are mounted and interconnected. An MCM, acting as a single package for multiple unpackaged ICs, thereby has a higher wiring density, a smaller size, and a higher reliability. These benefits come at a price, since MCM technologies are more expensive alternatives to traditional PCB implementations.

MCMs are categorised according to the materials that they use: MCM-L uses laminated PCBs, MCM-C uses multilayer thick film ceramic, and MCM-D uses thin film deposition. Table 2.15 lists some of the key properties of the three systems [144].

Table 2.15 *Selected features of various multi-chip module technologies.*

	cost factor	feature size	passives	fabrication	hermetic
MCM-L	$\times 1$	$\sim 100 \mu\text{m}$	mounted, etched	parallel	no
MCM-C	$\times 1.5\text{--}2$	$\sim 100 \mu\text{m}$	printed	parallel	yes
MCM-D	$\times 3\text{--}7$	$\sim 10 \mu\text{m}$	deposited/etched	sequential	depends on carrier

Microwave MCM-L packages are constructed by stacking and laminating low-cost copper-clad fluorocarbon laminates, such as polytetrafluoroethylene (PTFE/Teflon), which are widely available from suppliers such as *Rogers*, *Arlon*, and *Taconic*. A successful commercial deployment of MCM-L is *Merrimac Industries' Multi-Mix* product line, consisting of compact passive and active microwave components built using laminated PTFE layers [145]. The primary disadvantages of using organic laminates include their relatively poor thermal conductivity and their high coefficient of thermal expansion (CTE) mismatch with semiconductor materials. Further, MCM-L packages are not hermetic, requiring a separate hermetic capsule when this feature is required.

MCM-D uses standard IC fabrication equipment and techniques. The package consists of thin film dielectric and metalisation layers that are deposited on a rigid substrate (or *carrier*) and patterned by sequential photolithography. The feature size, on the order of $10 \mu\text{m}$, exceeds that of the other MCM technologies. Large pin-out chips, typical of computer processors, require wiring densities that can only be accommodated by MCM-D, allowing it to command a higher price. Typical substrates upon which the layers are patterned include silicon and high purity alumina.

MCM-C packages are processed using thick film glass-ceramic technology, including high temperature and low temperature cofired ceramic variants (HTCC and LTCC). Fabrication begins with green (unfired) ceramic tape, a malleable and supple material, into which via holes are punched or laser ablated. Apertures may also be punched at this stage, which will later form cavities in the module. A stainless steel stencil is then placed on the tape, having apertures that line up with the via holes in the tape. The vias are then filled by forcing conductor paste through the stencil. Next, a stainless steel screen is placed on the tape. Prior to being placed on the tape, the screen is impregnated with an emulsion mask that is patterned for the conductor traces using photolithography, as illustrated in Fig. 2.20.

Fig. 2.20 *Microscope photograph of an emulsion mask on a stainless steel screen.*

Printing of the conductors is accomplished by forcing conductor paste through the stainless steel screen. Once all of the layers have been patterned, they are stacked and subjected to high pressure, a step called *lamination*, during which the layers are bonded to each other. The modules are then fired (sintered) to crystallise the ceramic. The pure alumina substrates used in HTCC fire at over 1500 °C and can only use low conductivity

refractory (high melting point) metals such as tungsten or molybdenum—higher conductivity metals would melt, turning the conductor traces into beads. LTCC solves this problem by increasing the glass content of the alumina to 50% or more. With its lower firing temperature of 900 °C, high conductivity metals such as gold and silver, having melting points of 1060 and 960 °C, respectively, can be used to make low loss microwave circuitry.

The thick film printing technique limits the feature size to the order of 100 μm , although new, higher density lithographic methods can also be used on the top layer [146], [147]. Metals can be *brazed* to the ceramic surface. Brazing may be regarded as a high temperature (> 450 °C) solder process that produces strong metal-to-ceramic bonds. This allows a hermetic metal lid to be directly attached to the ceramic package. The thermal conductivity of ceramic is excellent, up to several hundred times that of MCM-L or MCM-D polymers, although trade-offs usually exist between CTE mismatch, thermal conductivity, and dielectric constant when using ceramic.

One of the primary advantages of LTCC is that thick film passives such as resistors, capacitors, and inductors may be buried in the stack by using an appropriate resistive, dielectric, or ferrite paste. The tolerances of passives are however quite high, on the order of $\pm 20\%$. Novel ceramic tapes are also beginning to appear, including systems with mixed dielectric constants and high dielectric constants ($\epsilon_r = 60$). Since ferrite is a ceramic, ferrite LTCC tape has also recently emerged [148]–[154].

LTCC has been successfully deployed on a commercial basis. Suppliers of bulk LTCC materials include *Dupont*, *Ferro*, and *Heraeus*. A method for the design of ferrite tape LTCC transformers and inductors has been recently patented by *Midcom* using

materials provided by *Electro-Science Laboratories* [148]. *Mini-Circuits* has developed an extensive line of LTCC-based microwave passives under the *Blue Cell* trademark [151]. Compact antenna modules and integrated antenna arrays have also been designed at microwave and millimetre-wave frequencies [152]–[154], and high performance millimetre-wave filters and waveguides have also been reported [155], [156].

2.6 State of the art and research goal

This chapter has shown that semiconductor phase shifters are smaller, offer faster switching speeds, and are easier to bias than their ferrite counterparts. However, even though a tremendous amount of research has been done on semiconductor phase shifters, their insertion loss and figures of merit remain significantly poorer than those using ferrite. Semiconductor shifters using PIN diodes or FET switches operating in the millimetre-wave range suffer from insertion losses of over 5 dB and yield figures of merit in the tens of degrees per decibel. These figures may be improved slightly by using MEMS transducers, which reduce the insertion loss to a few decibels and increase the figure of merit to approximately 100 °/dB, although MEMS have only recently been realized on GaAs. Although ferrite waveguide phase shifters offer unsurpassed figures of merit of more than 500 °/dB, their construction has relied on the manual assembly of expensive, miniature components. Being three-dimensional, with proportions on the order of a few centimetres, no attempts have been made to integrate waveguide ferrite phase shifters with planar and/or monolithic circuitry.

Although seemingly unrelated, significant advances have been made in multi-chip packaging technology, and especially in the area of low temperature cofired ceramic materials. One of the major advantages of LTCC is that buried passives may be used,

which has fuelled research into resistive, capacitive, and ferrite pastes. This research has culminated with the recent availability of high dielectric and ferrite LTCC tape. LTCC has a number of key properties, including:

- an excellent low-loss performance at millimetre-wave frequencies;
- a competitive price for millimetre-wave packaging material, especially at high volume;
- relatively large dimensions, on the order of a few centimetres;
- the availability of millimetre-wave guiding structures, including waveguide;
- an inherent integrability with semiconductors and other passives;
- the availability of new, ceramic ferrite tapes and ferrite pastes.

The convergence of these points leads to the hypothesis that ferrite LTCC packaging is an appropriate medium for the integration of high-performance millimetre-wave ferrite waveguide phase shifters. One compact package could not only contain a phase shifter, but could possibly also house all of its required driving circuitry as well as other system components. No such phase shifter has yet appeared in the open literature; however, with the recent documentation of ferrite tape transformers and planar ferrite phase shifters, a window of opportunity currently exists for the proof of concept of a ferrite waveguide millimetre-wave phase shifter embedded entirely in LTCC packaging material.

2.7 Chapter Two references

- [1] R. K. Mains, G. I. Haddad, and D. F. Peterson, "Investigations of broad-band, linear phase shifters using optimum varactor diode doping profiles", *IEEE Trans. Microwave Theory Tech.*, vol. 29, pp. 1158–1164, Nov. 1981.
- [2] E. N. Phillips, "The uncertainties of phase measurement", *Microwaves*, pp. 14–21, Feb. 1965.

- [3] M. J. Schindler *et al.*, "Monolithic 6–18 GHz 3 bit phase shifter", in *IEEE GaAs IC Symp. Dig.*, 1985, pp. 129–132.
- [4] S. M. Sze, *Physics of Semiconductor Devices*, 2nd ed., New York: Wiley, 1981.
- [5] J. W. Archer, R. A. Batchelor, and C. J. Smith, "Low-parasitic, planar Schottky diodes for millimeter-wave integrated circuits", *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 15–22, Jan. 1990.
- [6] C.-L. Chen, W. E. Courtney, L. J. Mahoney, M. J. Manfra, A. Chu, and H. A. Atwater, "A low-loss Ku-band monolithic analog phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 315–320, March 1987.
- [7] I. Bahl & P. Bhartia, *Microwave Solid State Circuit Design*, New York: Wiley, 1988, chap. 8 & 12.
- [8] B. N. Scott & G. E. Brehm, "Monolithic voltage controlled oscillator for X- and Ku-bands", *IEEE Trans. Microwave Theory Tech.*, vol. 30, pp. 2172–2177, Dec. 1982.
- [9] P. Philippe, W. El-Kamali, and V. Pauker, "Physical equivalent circuit model for planar Schottky varactor diode", *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 250–255, Feb. 1988.
- [10] K. Chang, *Microwave Solid-State Circuits and Applications*, New York: Wiley, 1994, chap. 5 & 8.
- [11] K. Wilson, J. M. C. Nicholas, G. McDermott, and J. W. Burns, "A novel MMIC X-band phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 33, pp. 1572–1578, Dec. 1985.
- [12] D. M. Krafcsik, S. A. Imhoff, D. E. Dawson, and A. L. Conti, "A dual-varactor analog phase shifter operating at 6 to 18 GHz", *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1938–1941, Dec. 1988.
- [13] S. Lucyszyn, G. Green, and I. D. Robertson, "Accurate millimeter-wave large signal modeling of planar Schottky varactor diodes", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1992, pp. 259–262.
- [14] S. Lucyszyn & I. D. Robertson, "Synthesis techniques for high performance octave bandwidth 180° analog phase shifters", *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 731–740, April 1992.
- [15] J. F. White, *Microwave Semiconductor Engineering*, New York: Van Nostrand Reinhold, 1982.
- [16] J.-L. Lee, D. Zych, E. Reese, and D. M. Drury, "Monolithic 2-18 GHz low loss, on-chip biased PIN diode switches", *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 250–256, Feb. 1995.
- [17] J. V. Bellantoni, D. C. Bartle, D. Payne, G. McDermott, S. Bandla, R. Tayrani, and L. Raffaelli, "Monolithic GaAs p-i-n diode switch circuits for high-power millimeter-wave applications", *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 2162–2165, Dec. 1989.
- [18] J. Putnam, M. Barter, K. Wood, and J. LeBlanc, "A monolithic GaAs PIN switch network for a 77 GHz automotive collision warning radar", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1997, vol. 2, pp. 753–756.

- [19] E. Alekseev & D. Pavlidis, "High switching rate capability of mm-wave InGaAs PIN diodes", in *Proc. 11th Int. Conf. on Indium Phosphide and Related Materials*, IPRM 1999, pp. 443–446.
- [20] D. Tonks, "Controlled rise and fall time switches using GaAs MMICs", *Microwave J.*, vol. 34, pp. 124–130, Aug. 1991.
- [21] K. W. Kobayashi, L. Tran, A. K. Oki, and D. C. Streit, "A 50 MHz–30 GHz broadband co-planar waveguide SPDT PIN diode switch with 45-dB isolation", *IEEE Microwave Guided Wave Lett.*, vol. 5, pp. 56–58, Feb. 1995.
- [22] G. H. Nesbit, D. W. Wong, D. Li, and J. C. Chen, "A W-Band monolithic GaAs PIN diode switch", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1986, vol. 1, pp. 51–55.
- [23] G. H. Nesbit *et al.*, "Monolithic transmit/receive switch for millimeter-wave application", in *IEEE Gallium Arsenide Integrated Circuit Symp. Dig.*, 1987, pp. 147–150.
- [24] D. J. Seymour, D. D. Heston, and R. E. Lehmann, "X-band and Ka-band monolithic GaAs PIN diode variable attenuation limiters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1988, vol. 1, pp. 255–258.
- [25] A. Klaasen and J.-M. Dieudonné, "77 GHz monolithic MMIC Schottky- and PIN-diode switches based on GaAs MESFET and silicon SIMMWIC technology", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1995, vol. 3, pp. 1631–1634.
- [26] H. Takasu, F. Sadaki, M. Kawano, and S. Kamihashi, "Ka-band low loss and high power handling GaAs PIN diode MMIC phase shifter for reflected-type phased array systems", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1999, vol. 2, pp. 467–470.
- [27] R. H. Caverly and G. Hiller, "Distortion properties of MESFET and PIN diode microwave switches", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1992, vol. 2, pp. 533–536.
- [28] R. H. Caverly, "Distortion in GaAs MESFET switch circuits", *Microwave J.*, vol. 37, pp. 106–114, Sept. 1994.
- [29] W. V. McLevige & V. Sokolov, "Resonated GaAs FET devices for microwave switching", *IEEE Trans. Electron Devices*, vol. 28, pp. 198–204, 1981.
- [30] R. Carandang, J. Yonaki, W. L. Jones, R. E. Kasody, W. Lam, and L. C. T. Liu, "A K-band HEMT low noise receive MMIC for phased array applications", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1991, vol. 2, pp. 521–524.
- [31] S. S. Bharj and R. Goyal, "MESFET switch design", *Microwave System News & Communications Technology*, pp. 76–86, Nov. 1987.
- [32] P. Katsin, B. E. Bedard, M. B. Shifrin, and Y. Ayasli, "High speed, 100+ W RF switches using GaAs MMICs", *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1989–1996, Nov. 1992.
- [33] M. J. Schindler & A. Morris, "DC–40 GHz and 20–40 GHz MMIC SPDT switches", *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 1486–1493, Dec. 1987.
- [34] H. Takasu, F. Sasaki, H. Kawasaki, H. Tokuda, and S. Kamihashi, "W-band SPST transistor switches", *IEEE Microwave and Guided Wave Lett.*, vol. 6, pp. 315–316, Sept. 1996.

- [35] K. W. Kobayashi, A. K. Oki, L. B. Sjogren, D. K. Umemoto, T. R. Block, and D. C. Streit, "A monolithic HEMT passive switch with integrated HBT standard logic compatible driver for phased-array applications", *IEEE Microwave and Guided Wave Lett.*, vol. 6, pp. 375–377, Oct. 1996.
- [36] H. Mizutani and Y. Takayama, "A DC–60 GHz GaAs MMIC switch using novel distributed FET", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1997, vol. 2, pp. 439–442.
- [37] D. D. Heston, D. J. Seymour, and D. Zych, "100 MHz to 20 GHz monolithic single-pole, two-, three-, and four-throw GaAs PIN diode switches", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1991, vol. 2, pp. 429–432.
- [38] B. Glance & N. Amitay, "A fast switching, low loss, low drive, 12 GHz PIN phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1979, vol. 1, pp. 232–234.
- [39] P. Bernkopf, M. Schindler, and A. Bertrand, "A high power K/Ka-band monolithic T/R switch", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1991, vol. 1, pp. 15–18.
- [40] G. L. Lan, D. L. Dunn, J. C. Chen, C. K. Pao, and D. C. Wang, "A high performance V-band monolithic FET transmit-receive switch", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1988, pp. 99–101.
- [41] M. Madihiam, L. Desclos, K. Maruhashi, K. Onda, and M. Kuzuhara, "A sub-nanosecond resonant-type monolithic T/R switch for millimeter-wave systems applications", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1016–1019, July 1998.
- [42] M. J. Schindler & T. E. Kazior, "A high power 2–18 GHz T/R switch", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1990, vol. 1, pp. 119–122.
- [43] Y. Ayasli, R. Mozzi, L. Hanes, and L. D. Reynolds, "An X-Band 10 W monolithic transmit-receive GaAs FET switch", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1982, pp. 42–46.
- [44] E. R. Brown, "RF-MEMS switches for reconfigurable integrated circuits", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1868–1880, Nov. 1998.
- [45] N. S. Barker & G. M. Rebeiz, "Distributed MEMS true-time delay phase shifters and wide-band switches", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1881–1889, Nov. 1998.
- [46] J. S. Hayden & G. M. Rebeiz, "One and two-bit low-loss cascable MEMS distributed X-band phase shifters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 1, pp. 161–164.
- [47] J. B. Muldavin & G. M. Rebeiz, "High-isolation inductively-tuned X-band MEMS shunt switches", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 1, pp. 169–172.
- [48] D. Peroulis, K. Sarabandi, and L. P. B. Katehi, "Low contact resistance series MEMS switches", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2002, vol. 1, pp. 223–226.

- [49] D. Hah, E. Yoon, and S. Hong, "A low voltage actuated micromachined microwave switch using torsion springs and leverage", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 1, pp. 157–160.
- [50] A. K. Sharma, "Solid-state control devices: state of the art", *Microwave J.*, vol. 32, 1989 State of the Art Reference, pp. 95–112, Sept. 1989.
- [51] V. Sokolov, J. J. Geddes, A. Contolatis, P. E. Bauhahn, and C. Chente, "A Ka-band GaAs monolithic phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 31, pp. 1077–1083, Dec. 1983.
- [52] J. R. Potukuchi, R. C. Mott, A. I. Zaghloul, R. K. Gupta, F. T. Assal, and R. M. Sorbello, "MMICs insertion in a Ku-band active phased array for communications satellites", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1990, vol. 2, pp. 881–882.
- [53] K. Maruhashi, H. Mizutani and K. Ohata, "A Ka-band 4-bit monolithic phase shifter using unresonated FET switches", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1998, vol. 1, pp. 51–54.
- [54] P. Bauhahn, C. Butter, V. Sokolov, and A. Contolatis, "30 GHz multi-bit monolithic phase shifters", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1985, pp. 4–7.
- [55] V. E. Dunn, N. E. Hodges, O. A. Sy, W. Alyassini, M. Feng, and Y. C. Chang, "MMIC phase shifters and amplifiers for millimeter-wavelength active arrays", in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 1, pp. 127–130.
- [56] M. Aust, H. Wang, R. Carandang, K. Tan, C. H. Chen, T. Trinh, R. Esfandiari, and H. C. Yen, "GaAs monolithic components development for Q-band phased array applications", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1992, vol. 2, pp. 703–706.
- [57] H. Wang *et al.*, "Monolithic Q band active array module and antenna", *Applied Microwave*, pp. 88–102, Winter 1993.
- [58] H. Q. Tserng, P. Saunier, A. Ketterson, L. C. Witkowski, and T. Jones, "Embedded transmission-line (ETL) MMIC for low-cost high-density wireless communication applications", *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 2540–2548, Dec. 1997.
- [59] B. Pillans, S. Eshelman, A. Malczewski, J. Ehmke, and C. Goldsmith, "Ka-band RF MEMS phase shifters", *IEEE Microwave and Guided Wave Lett.*, vol. 9, pp. 520–522, Dec. 1999.
- [60] V. E. Dunn *et al.*, "MMIC components for mm-wavelength active arrays", *Microwave J.*, vol. 32, pp. 109–116, Dec. 1989.
- [61] R. V. Garver, "Broad-band diode phase shifters", *IEEE Trans. Microwave Theory Tech.*, vol. 20, pp. 314–323, May 1972.
- [62] Y. Ayasli, S. W. Miller, R. L. Mozzi, and L. K. Hanes, "Wideband S-C band monolithic phase shifter", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1984, pp. 11–13.
- [63] Y. Ayasli, S. W. Miller, R. Mozzi, and L. K. Hanes, "Wide-band monolithic phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 32, pp. 1710–1714, Dec. 1984.

- [64] M. J. Schindler & M. E. Miller, "A 3 bit K/Ka band MMIC phase shifter", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1988, pp. 95–98.
- [65] C. Moye, G. Sakamoto, and M. Brand, "A compact broadband, six-bit MMIC phasor with integrated digital drivers", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1990, vol. 1, pp. 457–460.
- [66] R. Coats, J. Klein, S. D. Pritchett, and D. Zimmermann, "A low loss monolithic five-bit PIN diode phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1990, vol. 2, pp. 915–918.
- [67] M. E. Goldfarb, "A novel design for a MMIC 180 degree phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1992, vol. 1, pp. 175–177.
- [68] R. J. Naster, A. W. Jacomb-Hood and M. R. Lang, "Affordable MMIC designs for phased arrays", *Microwave J.*, vol. 30, pp. 141–150, March 1987.
- [69] D. Teeter *et al.*, "Ka-band GaAs HBT PIN diode switches and phase shifters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1991, vol. 1, pp. 451–454.
- [70] J. F. White, "High power, p-i-n diode controlled, microwave transmission phase shifters", *IEEE Microwave Theory Tech.*, vol. 13, pp. 233–242, March 1965.
- [71] H. A. Atwater, "Circuit design of the loaded-line phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 33, pp. 626–634, July 1985.
- [72] F. L. Opp & W. F. Hoffman, "Design of digital loaded-line phase-shift networks for microwave thin-film applications", *IEEE Trans. Microwave Theory Tech.*, vol. 16, pp. 462–468, July 1968.
- [73] I. J. Bahl & K. C. Gupta, "Design of loaded-line p-i-n diode phase shifter circuits", *IEEE Trans. Microwave Theory Tech.*, vol. 28, pp. 219–224, March 1980.
- [74] P. R. Shepherd & M. J. Cryan, "Schottky diodes for analogue phase shifters in GaAs MMIC's", *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 2112–2116, Nov. 1996.
- [75] T. Yahara, "A note on designing digital diode-loaded-line phase shifters", *IEEE Trans. Microwave Theory Tech.*, vol. 20, pp. 703–704, Oct. 1972.
- [76] M. E. Davis, "Integrated diode phase-shifter elements for an X-band phased-array antenna", *IEEE Trans. Microwave Theory Tech.*, vol. 23, pp. 1080–1084, Dec. 1975.
- [77] M. Shokrani & V. J. Kapoor, "InGaAs microwave switch transistors for phase shifter circuits", *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 772–778, May 1994.
- [78] W. A. Davis, "Design equations and bandwidth of loaded-line phase shifters", *IEEE Trans. Microwave Theory Tech.*, vol. 22, pp. 561–563, May 1974.
- [79] Y. Ayasli, A. Platzker, J. Vorhaus, Jr., and L. D. Reynolds, "A monolithic single-chip X-band four-bit phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 30, pp. 2201–2206, Dec. 1982.
- [80] C. Andricos & I. Bahl, "C-band 6-bit GaAs monolithic phase shifter", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1985, pp. 8–9.
- [81] R. Coats, J. Klein, S. D. Pritchett, and D. Zimmermann, "A low loss monolithic five-bit PIN diode phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1990, vol. 2, pp. 915–918.

- [82] W. R. Wisseman *et al.*, "X-band GaAs single-chip T/R radar module", *Microwave J.*, vol. 30, pp. 167–172, Sept. 1987.
- [83] A. L. Riley, D. Rascoe, V. Lubecke, J. Huang, and L. Duffy, "Ka-band MMIC beam steered transmitter array", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1989, pp. 65–68.
- [84] R. J. Lang & B. J. Edward, "A 35 GHz electronically steered line array", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1987, vol. 2, pp. 937–940.
- [85] K. Nakahara, S. Chaki, N. Andoh, H. Matsuoka, N. Tanino, Y. Mitsui, M. Otsubo, and S. Mitsui, "A novel three phase-states phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1993, vol. 1, pp. 369–372.
- [86] C. Nguyen, "Design and performance of a W-band broadband microstrip phase shifter", *Microwave J.*, vol. 30, pp. 91–102, July 1989.
- [87] A. S. Nagra & R. A. York, "Distributed analogue phase shifters with low insertion loss", *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1705–1711, Sept. 1999.
- [88] N. S. Barker & G. M. Rebeiz, "Optimization of distributed MEMS phase shifters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 1, pp. 299–302.
- [89] J. Huang & R. J. Pogorzelski, "A Ka-band microstrip reflectarray with elements having variable rotation angles", *IEEE Trans. Antennas Propagat.*, vol. 46, pp. 650–656, May 1998.
- [90] A. W. Jacomb-Hood, D. Seielstad, and J. D. Merrill, "A three-bit monolithic phase shifter at V-band", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1987, pp. 81–84.
- [91] F. Ali & A. Podell, "A wide-band GaAs monolithic spiral quadrature hybrid and its circuit applications", *IEEE J. Solid-State Circuits*, vol. 26, pp. 1394–1398, Oct. 1991.
- [92] F. Ali & N. Mysoor, "An analog MMIC phase modulator for X-Band satellite transponder applications", *IEEE Microwave and Guided Wave Lett.*, vol. 2, pp. 445–446, Nov. 1992.
- [93] D. E. Dawson, A. L. Conti, S. H. Lee, G. F. Shade, and L. E. Dickens, "An analog X-band phase shifter", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, pp. 6–10.
- [94] D. C. Boire, G. St. Onge, C. Barratt, G. B. Norris, and A. Moysenko, "4:1 bandwidth digital five bit MMIC phase shifters", in *IEEE Microwave and Millimeter-Wave Monolithic Circuits Symp. Dig.*, 1989, pp. 69–73.
- [95] J. I. Upshur & B. D. Geller, "Low-loss 360° X-band analog phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1990, vol. 1, pp. 487–490.
- [96] D. C. Boire, J. E. Degenford, and M. Cohn, "A 4.5 to 18 GHz phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1985, pp. 601–604.
- [97] S. Lucyszyn & I. D. Robertson, "Two-octave bandwidth monolithic analog phase shifter", *IEEE Microwave and Guided Wave Lett.*, vol. 2, pp. 343–345, Aug. 1992.
- [98] B. Ulriksson, "Continuous varactor-diode phase shifter with optimized frequency response", *IEEE Trans. Microwave Theory Tech.*, vol. 27, pp. 650–654, July 1979.
- [99] E. C. Niehenke, V. V. DiMarco, and A. Friedberg, "Linear analog hyperabrupt varactor diode phase shifters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1985, pp. 657–660.

- [100] R. V. Garver, "360° varactor linear phase modulator", *IEEE Trans. Microwave Theory Tech.*, vol. 17, pp. 137–147, March 1969.
- [101] S. Nam, C. W. Park, F. M. Ghannouchi, E. Allamando, and I. D. Robertson, "Wideband monolithic millimeter-wave phase shifter with minimum insertion loss variation", in *IEEE 2000 MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 3, pp. 1577–1580.
- [102] H. A. Atwater, "Reflection coefficient transformations for phase-shift circuits", *IEEE Trans. Microwave Theory Tech.*, vol. 28, pp. 563–568, June 1980.
- [103] M. Lajugie, H. Brouzes, and F. Villain, "Digital vector control chip set at Q band for communication phased array", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1998, vol. 3, pp. 1577–1580.
- [104] A. J. Slobodnik, Jr., R. T. Webster, and G. A. Roberts, "A monolithic GaAs 36 GHz four-bit phase shifter", *Microwave J.*, vol. 36, pp. 106–111, June 1993.
- [105] M. Tuckman, "I-Q vector modulator—the ideal control component?", *Microwave System News & Communications Technology*, pp. 105–115, May 1988.
- [106] A. E. Ashtiani, S.-I. Nam, A. d'Espona, S. Lucyszyn, and I. D. Robertson, "Direct multilevel carrier modulation using millimeter-wave balanced vector modulators", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 2611–2619, Dec. 1998.
- [107] D. S. McPherson, I. D. Robertson, and S. Lucyszyn, "A W-band vector modulator and its application to software radar for automotive collision avoidance", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2000, vol. 3, pp. 1423–1426.
- [108] A. E. Ashtiani, T. Gokdemir, A. Vilches, Z. Hu, I. D. Robertson, and S. P. Marsh, "Monolithic GaAs/InGaP HBT balanced vector modulators for millimeter-wave wireless systems", in *IEEE Radio Frequency Integrated Circuits Symp. Dig.*, 2000, pp. 187–190.
- [109] J. B. Hacker, R. E. Mihailovich, M. Kim, and J. F. DeNatale, "A Ka-Band 3-Bit RF MEMS True-Time-Delay Network", *IEEE Trans. Microwave Theory Tech.*, vol. 51, pp. 305–308, Jan. 2003.
- [110] G.-L. Tan, R. E. Mihailovich, J. B. Hacker, J. F. DeNatale, and G. M. Rebeiz, "Low-Loss 2- and 4-Bit TTD MEMS Phase Shifters Based on SP4T Switches", *IEEE Trans. Microwave Theory Tech.*, vol. 51, pp. 297–304, Jan. 2003.
- [111] J. Smit & H. P. J. Wijn, *Ferrites*, New York: Wiley, 1959.
- [112] D. L. Bergman & J. P. Wesley, "Spinning charged ring model of electron yielding anomalous magnetic moment", *Galilean Electrodynamics*, vol. 1, no. 5, pp. 63–67, Sept.-Oct. 1990.
- [113] D. L. Bergman, "Spinning charge ring model of elementary particles", *Galilean Electrodynamics*, vol. 2, no. 2, pp. 30–32, March-April 1991.
- [114] R. A. Waldron, *Ferrites—An Introduction for Microwave Engineers*, London: D. Van Nostrand, 1961.
- [115] A. J. Baden Fuller, *Ferrites at Microwave Frequencies*, London: Peter Peregrinus, 1987.
- [116] R. F. Soohoo, *Theory and Application of Ferrites*, Englewood Cliffs, NJ: Prentice-Hall, 1960.
- [117] K. J. Standley, *Oxide Magnetic Materials*, 2nd ed., London: Oxford University Press, 1972.

- [118] A. Goldman, "Magnetic Ceramics", in *Electronic Ceramics—Properties, Devices, and Applications*, L. M. Levinson (ed.), chap. 3, New York: Marcel Dekker, 1988.
- [119] A. Goldman, *Modern Ferrite Technology*, New York: Van Nostrand Reinhold, 1990.
- [120] R. Valenzuela, *Magnetic Ceramics*, Cambridge, UK: Cambridge University Press, 1994.
- [121] S. K. Koul & B. Bhat, *Microwave and Millimeter Wave Phase Shifters: Volume II, Dielectric and Ferrite Phase Shifters*, Dedham, Massachusetts: Artech House Inc., 1991.
- [122] C. R. Boyd, Jr., "A dual-mode latching reciprocal ferrite phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 18, pp. 1119–1124, Dec. 1970.
- [123] L. R. Whicker & C. R. Boyd, Jr., "A new reciprocal phaser for use at millimeter wavelengths", *IEEE Trans. Microwave Theory Tech.*, vol. 19, pp. 944–945, Dec. 1971.
- [124] C. R. Boyd, Jr., "A 60 GHz dual-mode ferrite phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1982, pp. 257–259.
- [125] X. Yansheng & J. Zhengchang, "Dual-mode latching ferrite devices", *Microwave J.*, vol. 29, pp. 277–285, May 1986.
- [126] W. E. Hord, C. R. Boyd, and D. Diaz, "A new type of fast-switching dual-mode ferrite phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 1219–1225, Dec. 1987.
- [127] W. E. Hord, "Microwave and millimeter-wave ferrite phase shifters", *Microwave J.*, vol. 32, 1989 State of the Art Reference, pp. 81–94, Sept. 1989.
- [128] E. Gimonet & B. Chan Song Lint, "Optimization and design of millimeter wavelength phase shifters", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1979, pp. 373–375.
- [129] E. Schlömann, "Theoretical analysis of twin-slab phase shifters in rectangular waveguide", *IEEE Trans. Microwave Theory Tech.*, vol. 14, pp. 15–23, Jan. 1966.
- [130] R. W. Babbitt & R. A. Stern, "Fabrication and performance of ferrite phase shifters for millimeter frequencies", *IEEE Trans. Magn.*, vol. 15, pp. 1744–1746, Nov. 1979.
- [131] J. F. McIlvanna, "Monolithic phased arrays for EHF communications terminals", *Microwave J.*, vol. 31, pp. 113–125, March 1988.
- [132] R. W. Babbitt & R. A. Stern, "Non-reciprocal ferrite phase shifters for millimeter applications", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1978, pp. 94–96.
- [133] H. Luo, Y. Liang, J. Yuan, and Q. Qiu, "A Ka band compressed grooved waveguide ferrite phase shifter", in *Proc. Int. Conf. Millimeter Wave and Far Infrared Science and Technology*, Aug. 1996, pp. 150–153.
- [134] W. E. Hord, F. J. Rosembaum, and C. R. Boyd, "Theory of the suppressed-rotation reciprocal ferrite phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 16, pp. 902–910, Nov. 1968.
- [135] Cernex Inc., CA. Ferrite phase shifters. [Online]. Available: <http://www.cernex.com/PRODUCTS/Phase-shifter-PDF/Ferrite%20Phase%20Shifters.pdf>
- [136] R. W. Babbitt & R. A. Stern, "Millimeter wave ferrite devices", *IEEE Trans. Magn.*, vol. 18, pp. 1592–1594, Nov. 1982.

- [137] V. V. Meriakri & M. P. Parkhomenko, "Millimeter wave phase shifters based on ferrite dielectric waveguide", in *Proc. 12th Int. Conf. Microwaves and Radar, MIKON'98*, vol. 2, May 1998, pp. 514–517.
- [138] C. E. Patton, "Hexagonal ferrite materials for phase shifter applications at millimeter wave frequencies", *IEEE Trans. Magn.*, vol. 24, pp. 2024–2028, May 1988.
- [139] X. Zuo, H. How, P. Shi, S. A. Oliver, and C. Vittoria, "Development of high frequency ferrite phase shifter", *IEEE Trans. Magn.*, vol. 37, pp. 2395–3397, July 2001.
- [140] X. Zuo, H. How, P. Shi, S. A. Oliver, and C. Vittoria, " Zn_2Y Hexaferrite ($\text{Ba}_2\text{Zn}_2\text{Fe}_{12}\text{O}_{22}$) single-crystal microstripline phase shifter", *IEEE Trans. Magn.*, vol. 38, pp. 3493–3497, Sept. 2002.
- [141] E.-B. El-Sharawy & C. J. Koza, "Dual-ferrite slot line for broadband high-nonreciprocity phase shifters", *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 2204–2210, Dec. 1991.
- [142] E.-B. El-Sharawy, S. Gaglione, and D. Partlow, "Low cost ferrite toroid phase shifter for broadband applications", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1999, vol. 3, pp. 1157–1160.
- [143] S. N. Stitzer, "Modeling a stripline ferrite phase shifter", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1997, vol. 2, pp. 1117–1120.
- [144] N. Sherwani, Q. Yu, and S. Badida, *Introduction to Multichip Modules*, New York: Wiley, 1995.
- [145] Merrimac Industries, Inc., NJ. "A three-channel vector modulator with reference channel", *Microwave J.*, vol. 41, pp. 128–130, Nov. 1998.
- [146] DuPont Microcircuit Materials, NC. "Cost-effective solutions for high density interconnect and RF modules using low temperature cofired ceramic materials", *Microwave J.*, vol. 44, pp. 140–150, April 2001.
- [147] M. A. Skurski, M. A. Smith, R. R. Draudt, D. I. Amey, S. J. Horowitz, and M. J. Champ, "Thick-film technology offers high packaging density", *Microwaves & RF*, vol. 38, pp. 77–86, Feb. 1999.
- [148] G. Slama, "Low-temp co-fired magnetic tape yields high benefits", *Power Electronics Technology*, pp. 30–34, Jan. 2003.
- [149] Lamina Ceramics, NJ. "New substrate technology for high performance circuits and systems", *Microwave J.*, vol. 45, pp. 104–106, Dec. 2002.
- [150] Northrup Grumman Corp., CA. Ferrite Production and Development. [Online]. Available: <http://sensor.northrupgrum.com/es/ltcc/FerriteProductionDevelopment.pdf>
- [151] Mini-Circuits, Blue Cell Technology, NY. "High performance, low cost, surface-mount microwave mixers", *Microwave J.*, vol. 43, pp. 156–158, April 2000.
- [152] G. Heftman, "Chip antenna reduces cell-phone dimensions", *Microwaves & RF*, vol. 38, p. 182, Nov. 1999.
- [153] H. Uchimura & T. Takenoshita, "A ceramic planar 77 GHz antenna array", in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1999, vol. 2, pp. 453–456.
- [154] J. O. Litzenburger, M. Clénet, G. A. Morin, and Y. M. M. Antar, "Study of a waveguide antenna implemented in laminated material", Defence Research and Development Canada, Ottawa, ON, Tech. Rep. TR 2002-132, Dec. 2002.

- [155] H. Uchimura, T. Takenoshita, and M. Fujii, "Development of a 'laminated waveguide'", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 2438–2443, Dec. 1998.
- [156] J. Kassner & W. Menzel, "A drop-on band-pass filter for millimeter-wave multichip modules on LTCC", *IEEE Microwave Guided Wave Lett.*, vol. 9, pp. 456–457, Nov. 1999.

Chapter Three — Characterisation of LTCC materials and transmission media at microwave frequencies

The microwave properties of LTCC materials are often regarded as trade secrets and are rarely fully disclosed by LTCC suppliers, even though these are crucial to the proper design of microwave circuitry. The objective of this chapter is to identify and to implement methods by which LTCC materials and transmission media may be characterised in terms of loss and permittivity. Many different approaches may be followed for characterisation purposes. In this chapter, only *in situ* characterisation methods will be considered, meaning that the material will be characterised by measuring the response of *circuits* that have been built from the material itself. This contrasts with *ex situ* techniques, in which a blank sample of the material is inserted into testing apparatus that has been designed specifically for the measurement of a certain material property. The *in situ* techniques described herein have been conceived to use wafer probing equipment and a network analyser, items that should be commonly available to a microwave engineer. The characterisation of a given transmission medium therefore begins with the measurement of scattering parameters; from these, the properties of the LTCC material system itself are then deduced.

Only two types of transmission media have been considered in this work, the first being microstrip line. Microstrip is a popular type of transmission line and is commonly used for characterisation purposes. The first half of this chapter, Section 3.1, deals uniquely with microstrip lines. The remainder of the chapter, Section 3.2, presents the second medium, laminated waveguide, which is the implementation of a rectangular

waveguide inside an LTCC stack. Laminated waveguide has been chosen in anticipation of the ferrite phase shifters that will be implemented using this medium.

3.1 Microstrip characterisation techniques

The characterisation of transmission lines may be broadly subdivided into non-resonant and resonant techniques. The term *non-resonant* refers to methods in which the line is operated in its transmission mode, coupling power directly from input to output. Although several methods may be envisioned, only three are presented in Sections 3.1.1–3.1.3, which are the Absolute Magnitude, Line Comparison, and Thru-Reflect-Line methods. Resonant methods characterise the line by analysing a resonance, and a new microstrip ring resonator is used for this purpose in Section 3.1.4.

3.1.1 Absolute Magnitude method for the loss characterisation of microstrip

The Absolute Magnitude method is the most rudimentary of the three non-resonant characterisation techniques, for it uses the response of only one transmission line of length d to determine the loss. Theoretically, the measured transmission scattering parameter should suffice for characterisation since for a transmission line, $s_{21} = e^{-\gamma d}$ yields the desired complex propagation constant $\gamma = \alpha + j\beta$. Unfortunately, several factors limit the practicality of this method. Firstly, a transition is usually required between the measurement equipment and the transmission line under test. The measured s-parameter matrix $[s']$ is therefore that of the cascade shown in Fig. 3.1, where $[T]$ is the s-matrix of each transition and $[s]$ is the desired scattering matrix of the transmission line.

Fig. 3.1 *Measurement of a transmission line embedded between two transitions.*

For sake of simplicity, the two transitions are assumed to be identical although in reality they are only approximately so. The planar transmission to be measured is microstrip whereas the measurement system probes are usually implemented in coplanar waveguide (CPW), thus microstrip-to-CPW transitions are required, as illustrated in Fig. 3.2.

Fig. 3.2 *Coplanar waveguide-to-microstrip transition.*

The centre CPW probe contacts the microstrip conductor whereas the outer CPW probes contact vias that lead directly to the microstrip ground plane. Higher-order modes exist in the vicinity of the transition, contributing reactance to the transition's scattering matrix and thereby corrupting the measured phase of the transmission line, which masks the lossless propagation constant β of the line. Without this constant, the relative permittivity of the substrate cannot be determined; for this reason, only the magnitude of the transmission coefficient of the line is considered, hence the name *Absolute Magnitude*.

If the reflections from the transitions are high, internal reflections between the two transitions will establish resonance at certain frequencies, further corrupting the observed data. Conceptually, such a condition would likely be caused by each transition reactively loading the ends of the transmission line with a susceptance B , as shown in Fig. 3.3.

Fig. 3.3 *Reactively end-loaded transmission line.*

This type of resonance is recognisable since it occurs for line lengths that are odd multiples of $\lambda_g/4$. Measurement artefacts are also caused by mismatch between the transmission line characteristic impedance Z_0 and the measurement system impedance Z_S , which also causes resonant ripples in the measured scattering coefficients. Using the theory of infinite reflections [1], it can be shown that a terminated, mismatched line generates the reflection and transmission coefficients given by:

$$\Gamma = \frac{\Gamma_1(1 - e^{-2\gamma d})}{1 - \Gamma_1^2 e^{-2\gamma d}} \quad (3.1)$$

$$T = \frac{(1 - \Gamma_1^2)e^{-\gamma d}}{1 - \Gamma_1^2 e^{-2\gamma d}} \quad (3.2)$$

where Γ_1 is the magnitude of the partial mismatch between Z_0 and Z_S , given by $\Gamma_1 = (Z_0 - Z_S)/(Z_0 + Z_S)$. Impedance mismatch occurs at even multiples of $\lambda_g/4$, that is, at half-wavelength multiples. The final source of error in the measurement is attributable to the ohmic contact between the measurement probe tips and the pads on the circuit, which is never ideal, inducing additional loss through the transitions.

Consider the measured response shown in Fig. 3.4, which is that of a microstrip of width $W = 210 \mu\text{m}$ and length $d = 18.3 \text{ mm}$ with a substrate height $h = 223 \mu\text{m}$ using Heraeus CT2000 LTCC (c.f. Appendix A).

Fig. 3.4 Measured (a) reflection and (b) transmission s -parameter magnitudes of a Heraeus CT2000 microstrip transmission line of width $W = 210 \mu\text{m}$, length $d = 18.3 \text{ mm}$, and substrate height $h = 223 \mu\text{m}$.

The reflection response is consistent with that of an impedance mismatch. Using (3.1) with a partial mismatch of $\Gamma_1 = 0.07$ (-23.1 dB) and an effective length of $d = 18.1 \text{ mm}$ yields the simulated response of Fig. 3.5, which is superimposed on the averaged value of the measured s_{11} and s_{22} magnitudes for clarity.

Fig. 3.5 *Simulated and averaged measured reflection coefficients for the Heraeus microstrip line.*

With the Γ_1 term identified, the exponential $e^{-\gamma d}$ may be extracted from the transmission magnitudes using the following quadratic, which is a re-arrangement of (3.2):

$$T \Gamma_1^2 (e^{-\gamma d})^2 + (1 - \Gamma_1^2) (e^{-\gamma d}) - T = 0 \quad (3.3)$$

where T is the averaged measured transmission coefficient magnitude. The attenuation constant of the line is extracted directly from the exponential's magnitude, yielding the results displayed in Fig. 3.6(a), where a five-point smoothing operation has been employed to remove spurious spikes from the data.

Fig. 3.6 Attenuation constant of the Heraeus microstrip line using the Absolute Magnitude method (a) for a line length of 18.3 mm, (b) for three different line lengths: 1.80 mm, 2.65 mm, and 18.3 mm.

Fig. 3.6(a) indicates that the attenuation of the line ranges between 0.42–0.90 dB/cm across the band, which is known to be somewhat high for a microstrip line on LTCC. The lossy transitions are clearly to blame here, a hypothesis that is confirmed by Fig. 3.6(b), which plots the results of the Absolute Magnitude method for three different lengths of lines: 1.80 mm, 2.65 mm, and 18.30 mm. As expected, shorter lines appear to have higher attenuation constants because there is less line length over which the finite loss of the transitions may be distributed. As such, the Absolute Magnitude method is unreliable.

3.1.2 Line Comparison method for the loss characterisation of microstrip

As previously mentioned, the lossy transitions from CPW to microstrip, introduce additional loss into the measured transmission coefficients, causing the Absolute Magnitude method to falsely associate this loss with the attenuation of the line. However, if two different lengths of the *same* line are available, the difference between their transmission magnitudes should be equal to $e^{-\alpha\Delta}$, where Δ is the difference in length

between the two lines, and where it is assumed that the transitions in both lines are identical. To use this Line Comparison method, the transmission coefficients of both lines are first independently corrected for any mismatch using the correction method presented in the previous section. The transmission magnitude of the longer line is then assumed to be the product of that of the shorter line and $e^{-\alpha\Delta}$, from which the α estimate is obtained. Fig. 3.7 plots the attenuation constant obtained by comparing the original $d = 18.30$ mm Heraeus microstrip line to one that is $\Delta = 16.5$ mm shorter.

Fig. 3.7 Attenuation constant of the Heraeus microstrip line using the Line Comparison method with $\Delta = 16.5$ mm.

Although the curve of Fig. 3.7 has the same ripples as those in Fig. 3.6, its slope is much lower. The estimate of the attenuation constant now ranges between 0.36–0.55 dB/cm.

3.1.3 Thru-Reflect-Line (TRL) calibration method for characterising microstrip

Vector network analyser calibration methods have been conceived specifically for removing the effects of transitions. A popular calibration sequence is the TRL calibration method, in which at least three planar standards provide the information required to de-embed the transitions [2], [3]. The calibration method also yields an estimate of the

attenuation of the line. For the TRL calibration to work properly, the overall loss of the longer Line standard must obviously be greater than that of the shorter Thru standard. Unfortunately, repeated experimental measurements have found that the loss of the Thru standard is oftentimes *higher* than that of the Line, which is a most curious result and is discussed later. Fig. 3.8 shows the measured transmission coefficients of Thru and Line standards for the microstrip geometry listed in Table 3.1.

Table 3.1 *Parameters of the Thru and Line calibration standards for the Heraeus and Dupont LTCC material systems.*

LTCC system	dielectric constant ϵ_r	substrate height h [μm]	line width W [μm]	Thru standard length [μm]	Line standard length [μm]
Heraeus CT2000	9.1	223	210	1800	2650
Dupont 943	7.5	220	270	6000	7450

Fig. 3.8 *Transmission magnitudes of the microstrip Thru and Line standards for (a) the Heraeus CT2000 and (b) the Dupont 943 material systems listed in Table 3.1.*

For both systems, the loss of the Thru surpasses that of the Line standard at several frequencies. This fouls the calibration, which interprets the line as having gain at those

frequencies, a result that obviously negates any attempt at extracting attenuation data based on this calibration technique. The loss of the transmission line is so low that the additional length of the Line standard (ideally 90°) does not dramatically alter the insertion loss. The following factors may be contributing to the problematic results of Fig. 3.8: (i) that the transitions of the Thru and Line standards are not identical due to the loose LTCC tolerances, (ii) that the Thru standards are not long enough for evanescent modes to have died off, and (iii) that the loss associated with the probes contacting the pads is not repeatable. Consider the first factor (i): unlike thin film processing, the screen printing used in LTCC limits the resolution to line widths and spacings of approximately $100\text{ }\mu\text{m}$. This in turn leads to loose transition tolerances, which are illustrated in Fig. 3.9.

Fig. 3.9 *Illustration of the dimensional tolerance of LTCC processing: (a) a CPW-to-microstrip transition as designed, and (b) a photograph of the manufactured transition.*

The edge of an LTCC lines has a typical variation of at least $\pm 10\text{ }\mu\text{m}$. To the right of Fig. 3.9(a), the spacing between the ground pads and the centre conductor is $100\text{ }\mu\text{m}$, as is the centre conductor line width. This is obviously not respected in the manufactured transition in Fig. 3.9(b). The poor edge definition of the pads and lines means that the

Thru and Line standards do not have identical transitions, which inevitably impacts the TRL calibration sequence. The second cause (ii) for the poor TRL calibration is that the Thru is not long enough to allow evanescent modes to decay sufficiently. While this might be true for the relatively short Thru standard of the Heraeus system of Table 3.1, the Dupont 943 Thru standard is two wavelengths long at its highest frequency of 40 GHz. Further, HFSS simulations of the transitions have been performed to determine the rate at which higher order modes decay in the microstrip line. The lengths used for the Thru standards should have been adequate. The final cause (iii) for the TRL calibration problem is related to the repeatability of the contact between the probes and the printed probe pads. As opposed to thin films, printed LTCC lines have roughness values on the order of hundreds of nanometres. This roughness may lead to probing repeatability problems, an example of which is shown in Fig. 3.10, which plots the transmission coefficients of two successive measurements of the same Heraeus Thru standard.

Fig. 3.10 *Successive transmission measurements of the same Heraeus microstrip Thru standard of Table 3.1.*

The two curves are identical except for a finite level shift of approximately 0.3 dB. While this level of discrepancy may not be significant for other measurements, it seriously impacts the reliability of the TRL calibration, especially with regards to the attenuation estimate. This repeatability problem also affects the reliability of the Absolute Magnitude and Line Comparison methods. In essence, it means that the probe placement must be repeated several times before the levels can be trusted, and then either one measurement or an average of several measurements must then be taken. The uncertainty associated with the transmission magnitude levels of the calibration standards seriously restricts the use of any non-resonant method for LTCC loss characterisation.

3.1.4 Microwave characterisation using microstrip ring resonators

Resonators have long been used to characterise the microwave properties of materials [4]. Instead of relying on s-parameter magnitudes directly, the loss of the circuit is found by measuring the bandwidth of a resonance, as explained in Appendix C. The bandwidth measurement is insensitive to uniform magnitude level offsets such as the one illustrated in Fig. 3.10 and should therefore yield more reliable data than the non-resonant characterisation methods. The only disadvantage to the resonant method is that data is only collected at discrete resonant frequencies, requiring several different resonators to be measured if a fine frequency coverage is desired.

For planar circuitry, resonator configurations such as rings, half wavelength lines, and others have been used to characterise both the attenuation of the line and the substrate permittivity, which are obtained via a single swept frequency measurement [5]–[7]. The simple geometry of these resonators, combined with the relative ease of their analysis, contribute to their popularity, especially for benchmarking purposes. The resonator

configuration considered herein consists of a the two-port co-linear fed ring, illustrated in Fig. 3.11, where r is the mean radius, W is the microstrip line width, and S is the length of the coupling gaps.

Fig. 3.11 *Layout of a two-port, co-linear fed microstrip ring resonator.*

Analytically, the one-port, gap-coupled, lumped element model introduced by Yu will be expanded to the case of a two-port ring [8]. The lumped element model is an attractive one because it does not require complex simulation tools, as will be discussed in Section 3.1.4.1. The resonator theory leading to the characterisation of the microstrip line is presented in Section 3.1.4.2, whereas Section 3.1.4.3 presents the manufactured ring and the evaluation of its attenuation constant. Capacitors are used in the lumped element model to represent the coupling gaps; in Section 3.1.4.4, a method is presented by which these capacitor values are extracted using a commercial 2.5D electromagnetic solver, *Ansoft Ensemble*. The frequency pushing effect of the coupling gaps is analysed in Section 3.1.4.5 and a correction is then applied to the observed resonant frequencies of the ring, from which the dielectric constant of the substrate is finally obtained. Finally, the lumped element model of the two-port ring resonator will be invoked using the material parameters that have been experimentally obtained, thereby demonstrating the accuracy of the model.

3.1.4.1 Analytical behaviour of the microstrip ring resonator

Before considering the entire circuit, the input impedance of the ring alone shall be obtained. A previously derived model for the unloaded ring resonator is shown in Fig. 3.12(a) [8].

Fig. 3.12 *Circuit models of the unloaded ring resonator, (a) ring decomposed into two sections, (b) transmission line model.*

In Fig. 3.12(a), the ring is conceptually split into two sections of lengths L_1 and L_2 , with ports designated at the junctions. The sum of the two lengths must equal the circumference of the ring, $L = 2\pi r$. To obtain the unloaded ring input impedance Z_R , the input port is driven while the output port, which is arbitrarily chosen, is kept open-circuited. In [8], ABCD transfer matrices are derived for the individual line sections and their parallel connection is then expressed in the form of an admittance matrix, similar to the derivation presented in [7]. Finally, the admittance matrix is converted to an impedance matrix to obtain the ring's unloaded input impedance [8]:

$$Z_R = \frac{1}{j\omega C_R} \quad (3.4)$$

where

$$C_R = \frac{2Y_0[1 - \cos(2\pi\beta r)]}{\omega \sin(2\pi\beta r)} \quad (3.5)$$

and where Y_0 is the characteristic admittance, ω is the radian frequency, and β is the lossless propagation constant. The lengthy derivation leading to (3.4) is a good academic example of how transfer matrices may be used to express an input impedance, however a far simpler technique is to consider the ring as a parallel connection of two open-circuited transmission lines, as shown in Fig. 3.12(b). Given the anticipated co-linear feed, the length of each line section is $L_1 = L_2 = L/2$. Transmission line theory then allows the input impedance of the unloaded ring to be *immediately* expressed, without using (3.5), as:

$$Z_R = \frac{Z_0}{2j} \cot \frac{\beta L}{2} \quad (3.6)$$

where Z_0 is the characteristic impedance. It is straightforward to prove that (3.4) and (3.6) are identical. The unloaded ring model has recently been expanded to include loss, but the derivation again uses complicated transfer matrices [9]. Using (3.6), the derivation may be avoided since loss may be easily included by using the hyperbolic cotangent:

$$Z_R = \frac{Z_0}{2} \coth \frac{\gamma L}{2} \quad (3.7)$$

Equations (3.6) and (3.7) are by far the simplest presented to date for the unloaded impedance of a co-linear fed ring. Using these, one finds that the unloaded ring resonator presents an input impedance that cycles between a series and a parallel resonance as a function of frequency, presenting the former when the circumference is an odd multiple of $\lambda_g/2$ and the latter when the multiple is even, where λ_g is the guided wavelength. This is summarised by:

$$L_{OUS} = (2N - 1)\lambda_g / 2 \quad (3.8a)$$

$$L_{OUP} = N\lambda_g \quad (3.8b)$$

where N is the order of the resonance. A resonator must couple to an external circuit if it is to be sensed. In doing so, parasitic elements are introduced that perturb the unloaded behaviour of the ring. It has been shown that the parasitic elements caused by the gap between the microstrip feed line and the ring may be modelled as an L-network of two capacitors, as shown in Fig. 3.13(a) for the one-port loaded ring [8]. As illustrated in Fig. 3.11, the present work uses the ring as a two-port loaded resonator, therefore the model has been extended to that shown in Fig. 3.13(b).

Fig. 3.13 Loaded ring resonator models for (a) one-port gap coupling and (b) co-linear two-port gap coupling.

Conceptually, the two-port model is justified since increasing the gap to infinity on any one side must yield the same one-port model of Fig. 3.13(a). A simple field analysis of the two-port ring reveals that an unloaded series resonance always produces a voltage null at the output port, meaning that a transmission null will always occur for series resonant modes. A lightly loaded two-port resonator already has near-negligible transmission levels, so these nulls will be buried in the noise of the analyser. This is illustrated in Fig. 3.14, where the transmittance of the two-port resonator illustrated in Fig. 3.13(b) is plotted versus the circumference of the ring.

Fig. 3.14 Magnitude of s_{21} for the loaded two-port ring resonator of Fig. 3.13(b) versus the ring circumference at 5.7 GHz. The microstrip width is $W = 280 \mu\text{m}$, the dielectric constant is 6.15, and the substrate height is $h = 187 \mu\text{m}$. The coupling gap length is $S = 50 \mu\text{m}$ and the attenuation of the microstrip is set to 0.1 dB/cm.

Only the parallel resonance modes of (5.8b) have voltage maxima at the output port, resulting in transmission peaks at those resonant frequencies. This clearly indicates that the ring's fundamental resonance *must* be chosen to coincide with the parallel resonant mode, or $L = \lambda_g$.

Given that the unloaded ring of Fig. 3.12(a) is operating in a parallel mode, one would expect the input impedance to resemble that of a lightly loaded (undercoupled) parallel resonator, similar to the one shown on the Smith Chart in Fig. 3.15(a).

Fig. 3.15 Smith Chart input impedance loci of undercoupled resonators, (a) parallel and (b) series.

In practice, this response will *not* be observed because the capacitive L-network of the input port inverts the input impedance, resulting in the reflection response shown in Fig. 3.15(b), which is consistent with the *series-resonant* circuit of Fig. 3.16.

Fig. 3.16 *Series model of the two-port ring resonator without parasitics, as seen by the external circuitry.*

Note however that the L-networks are not ideal inverters and will introduce parasitic reactance to the resonator, causing the locus to skew slightly towards the capacitive side of the Smith Chart. This ultimately shifts the resonant frequency, a topic that will be addressed in Section 3.1.4.5. Therefore, even though the unloaded ring itself is theoretically operating at what should be a parallel resonance, it appears to the external world as a series-connected, series resonator. This is not surprising, since as the coupling is loosened, the input impedance must tend toward that of an open circuit, meaning that the resonance should appear on the right hand side of the Smith Chart.

3.1.4.2 Characterisation method using the microstrip ring

In this section, the experimental characterisation method is presented, where both the loss and permittivity are obtained from the same microwave measurement. The observed transmission peaks correspond *roughly* to the unloaded ring's parallel resonant frequencies, derived from (3.8b) and given by:

$$f_{0UP} = \frac{cN}{L_{0UP}\sqrt{\epsilon_{eff}}} \quad (3.9)$$

where ϵ_{eff} is the effective dielectric constant and c is the speed of light. Given the physical dimensions of the microstrip, any simple synthesis tool, such as *Agilent LineCalc*, may then be used to convert the effective dielectric constant to the substrate's relative permittivity. As previously mentioned, the capacitive L-networks are non-ideal impedance inverters and will cause the observed resonant frequency f_0 to be lower than the unloaded resonant frequency f_{0UP} . Hence the observed resonant frequency must be corrected prior to it being used in (3.9) to extract the dielectric constant. Failure to do this will cause the relative permittivity of the substrate to be overestimated.

The equation relating the total attenuation of the transmission line to the unloaded quality factor has been rigorously derived by Hsieh [9]. However, given that the unloaded ring model of Fig. 3.12(b) consists of a pair of half wavelength transmission lines, the attenuation of the line may be simply stated as [1:p.342]:

$$\alpha_T = \frac{\pi N}{Q_0 L} \quad (3.10)$$

where Q_0 is the unloaded quality factor of the resonator. Experimentally, the simplest way to evaluate the unloaded quality factor is according to the formula [10]:

$$Q_0 = \frac{Q_L}{1 - s_{21}^{res}} \quad (3.11)$$

where s_{21}^{res} is the magnitude of s_{21} in linear units at resonance and Q_L is the loaded quality factor, the latter given by:

$$Q_L = \frac{f_0}{\delta_L} \quad (3.12)$$

where δ_L is the double-sided loaded bandwidth, measured at a level 3 dB below s_{21}^{res} . More elaborate methods of assessing a resonator's quality factor are presented in Appendix C.

3.1.4.3 Physical circuits and loss characterisation

A total of three rings have been fabricated and tested. Two rings have been built using the Ferro A6-S system (c.f. Appendix A) while the other one uses Dupont 943. Table 3.2 lists the ring parameters.

Table 3.2 *Measured dimensions of the three microstrip ring resonators.*

ring name	substrate height h [μm]	microstrip width W [μm]	ring radius r [mm]	gap length S [μm]
Ferro1	187	315	3.96	100
Ferro2	187	315	4.42	100
DP943	220	270	3.70	135

The *Ferro1* and *Ferro2* rings are designed to have fundamental resonant frequencies near 5.7 and 5.2 GHz, respectively, whereas the *DP943* ring's fundamental should be near 5.8 GHz. The thickness of the microstrip conductors is approximately 8 μm for both material systems. The measured DC conductivity is $\sigma = 5.0(10^7) \pm 8\%$ S/m for the silver Ferro conductor CN33-391 whereas the measured conductivity of the Dupont silver conductor Ag-16A is $\sigma = 6.25(10^7)$ S/m, a value 1.3% greater than that of elemental silver, hence it is reasonable to assume that this trace is made from pure silver, $\sigma = 5.97(10^7)$ S/m. A feed line leading to one of the resonators is shown in Fig. 3.17, where the CPW probe transition is also visible to the far right. The substrate in the picture is slightly translucent, making the microstrip ground plane also visible as a shadow.

Fig. 3.17 Microscope photograph of one of the coupling gaps: the vertical arc is a section of the ring whereas the horizontal line is the feed line, leading to the CPW probe pads.

The measured transmission magnitudes of two of the rings are shown in Fig. 3.18, where the system has been calibrated to the probe tips only. The frequency span has been broken into separate bands and the frequency step sizes are sufficiently small such that the attenuation calculation of (3.10) always yields an uncertainty of less than 10%.

Fig. 3.18 Measured transmission magnitudes of the (a) Ferro1 and (b) DP943 ring resonators, showing seven and six resonance peaks, respectively.

Seven resonant frequencies are clearly visible in Fig. 3.18(a) whereas six are barely distinguishable above the noise floor in Fig. 3.18(b). Resonant bandwidth is marginally affected by the frequency pushing effect of the capacitive loading and so (3.10) leads directly to the total attenuation of the microstrip lines, which is plotted in Fig. 3.19.

Fig. 3.19 Measured attenuation of the (a) Ferro and (b) Dupont 943 microstrip lines versus frequency.

The total attenuation of both the Ferro and Dupont microstrip lines are nearly identical up to 30 GHz, ranging from 0.1 dB/cm near 5 GHz to approximately 0.35 dB/cm at 30 GHz. The average roughness of the conductors has also been measured using a profilometer. The average roughness R_a for the Ferro conductor is 700 nm while that of the Dupont conductor is 780 nm. The three factors that influence the total loss of a microstrip line are its DC conductivity, its roughness, and the loss tangent of the substrate. With the first two factors and the total attenuation known, the loss tangent of the substrate may be found using any microstrip synthesis tool such as *LineCalc*. This has been done for both LTCC systems and the results are plotted in Fig. 3.20.

Fig. 3.20 Calculated loss tangents of the (a) Ferro A6-S and (b) Dupont 943 substrates.

Owing to the high total attenuation of the first two data points of the *Ferro2* ring, this resonator yielded a very high loss tangent with respect to all of the other data points. Ignoring these, the loss tangent of the Ferro system appears to be generally lower than $2(10^{-3})$ across the 5–40 GHz band. The loss tangent of the Dupont 943 substrate is slightly higher, with most data points being consistently above $1.5(10^{-3})$. The noisiness of the loss tangent data suggests that microstrip ring resonators are perhaps not the best means for extracting this parameter in low loss systems.

3.1.4.4 Lumped element model capacitance evaluation of the gap-coupled ring

As previously mentioned, the observed resonant frequency must be corrected prior to its use in (3.9) to remove the frequency pushing effects of the parasitic loading capacitors C_g and C_p of Fig. 3.13(b). These capacitors must be evaluated before their parasitic effects can be removed. In previous work [8], the capacitor values were based on Gupta

et al.'s expressions for an end-to-end microstrip gap discontinuity, illustrated in Fig. 3.21(a) [11].

Fig. 3.21 *Microstrip gap discontinuities: (a) a two-port end-to-end, (b) a three-port T, (c) the equivalent lumped element model for the T-gap.*

Although this is a good first-order approximation, the interface of the feed line to a large-diameter ring more closely resembles a T-gap discontinuity, as shown in Fig. 3.21(b). This three-port T-gap may be readily simulated using a commercial 2.5D EM solver such as *Ansoft Ensemble*. Based on the simulated s-parameters, the derivations of Appendix B show that the model capacitors of Fig. 3.21(c) may be extracted using:

$$C_g = \frac{2Y_0|s_{21}|}{\omega\sqrt{|1+s_{11}|^2 - |s_{21}|^2}} \quad (3.13a)$$

$$C_p = \frac{Y_0[1-s_{11}]}{j\omega[1+s_{11}]} - \frac{2C_g Y_0}{2Y_0 + j\omega C_g} \quad (3.13b)$$

where s_{11} and s_{21} are the s-parameters from the 2.5D simulator and Y_0 is the characteristic admittance. Note that the complex value of s_{11} is used in (3.13), meaning that s_{11} must be de-embedded to the edge of the gap prior to its use. Equation (3.13) is general and is applicable to any microstrip T-gap.

The expressions for the end-to-end microstrip gap model, as derived by Gupta *et al.*, were curve-fitted to data that were in turn calculated using quasi-static methods [11], meaning that the resulting model does not include radiative effects. In keeping with this, the 2.5D simulation of the T-gap has been done by enclosing the structure in a perfect electric conducting (PEC) cavity. For the Ferro system, the sides of the cavity are 1700 μm long and the lid is $5h = 935 \mu\text{m}$ above the substrate; for the Dupont system, the cavity sides measure 2070 μm and the lid is $5h = 1060 \mu\text{m}$ above the substrate. An HFSS eigenmode simulation estimates the cavity modes to be well beyond the maximum frequency of interest. Fig. 3.22 plots the C_g extracted capacitor values with respect to those calculated using Yu's equations [8], the latter being constant versus frequency with $C_g = 7.53 \text{ fF}$ for the Ferro system and $C_g = 7.47 \text{ fF}$ for the Dupont system.

Fig. 3.22 C_g capacitor values for (a) a 100- μm microstrip gap using the Ferro A6-S parameters and (b) a 135- μm gap using the Dupont 943 parameters, both extracted from 2.5D simulations of a microstrip T-gap and expressed with respect to the values obtained by Yu's closed form equations [8].

The extracted capacitor values are close to those suggested by Yu, notwithstanding the frequency dependence. Next, Fig. 3.23 plots the C_p extracted capacitor values with

respect to those calculated using Yu's equations, the latter again being constant versus frequency with $C_p = 4.81$ fF for the Ferro system and $C_p = 6.10$ fF for the Dupont system.

Fig. 3.23 C_p capacitor values for (a) a 100- μm microstrip gap using the Ferro A6-S parameters and (b) a 135- μm gap using the Dupont 943 parameters, both extracted from 2.5D simulations of a microstrip T-gap and expressed with respect to the values obtained by Yu's closed form equations [8].

The corrections to be applied to Yu's C_p estimates are almost identical for both cases. In the work that follows, the extracted capacitor values will be used instead of Yu's estimated ones.

A good model of the ring and of its coupling gaps should not only predict the resonant frequency, but should also predict the magnitude of the transmission and reflection coefficients. To date, simulated coefficient magnitudes have not been corroborated with actual measurements [8]. Fig. 3.24 plots both the simulated and the observed resonant transmission peaks and reflection troughs of the Ferro rings, where the simulated data have been obtained using the lumped element model of Fig. 3.13(b).

Fig. 3.24 Measured and calculated (a) transmittance and (b) reflectance of the Ferro A6-S microstrip rings at resonance. The calculated values are obtained using the lumped element ring model with capacitance values extracted from the T-gap simulations.

Good agreement is obtained between the model and the measured data up to the fourth order resonance, near 20 GHz, where reflectance errors of less than 0.056 and transmittance errors of less than 0.022 are observed. Divergence above the fourth order resonance may be due to the presence of radiation in the physical circuits. The simulated and measured reflection and transmission magnitudes for the Dupont 943 resonator are shown in Fig. 3.25.

Fig. 3.25 Measured and calculated (a) transmittance and (b) reflectance of the Dupont 943 microstrip rings at resonance. The calculated values are obtained using the lumped element ring model with capacitance values extracted from the T-gap simulations.

Below 20 GHz, reflectance errors of less than 0.022 and transmittance errors of less than 0.0081 are observed. Based on these observations, the model appears to be better suited to s-parameter magnitudes for frequencies below 20 GHz.

3.1.4.5 Resonant frequency correction

The frequency pushing of the coupling capacitors will now be assessed using the lumped element model with the extracted capacitor values. Computed polar input reflection loci of the unloaded and two-port loaded rings are shown in Fig. 3.26, where the dashed outer locus is that of the *unloaded* ring while the solid inner locus is that of the two-port gap-coupled *loaded* ring.

Fig. 3.26 Input reflection loci of ring resonators on the Smith Chart. The dashed line represents the unloaded ring and is the same in both graphs, the arrow marking the unloaded parallel resonance. The solid line represents the two-port loaded ring for the cases of (a) tight coupling and (b) loose coupling.

The loaded ring is more tightly coupled in Fig. 3.26(a) than it is in Fig. 3.26(b), meaning that the gap size S would be larger for the loaded ring of Fig. 3.26(b). A series resonance f_s and a parallel resonance f_p can be defined for the loaded ring at points where the input reactance goes to zero [7], [8]. This happens at two frequencies on the loaded ring's locus in Fig. 3.26(a). If the impedance inverters were ideal, then the arrow in Fig. 3.26(a), which represents the parallel resonance of the unloaded ring, would occur at the same frequency as the *observed* resonance (square symbol) of the loaded ring, the latter being the minimum point of the reflection magnitude. Note, however, that the minimum reflection magnitude occurs at a frequency that is below the unloaded parallel resonant frequency. The minimum reflection coefficient occurs at a location on the locus that is slightly capacitive, in fact, the entire loaded locus is skewed slightly to the capacitive side—this is the effect of the loading capacitors, which act as non-ideal impedance inverters. This means that the observed resonant frequency, represented by the square in

Fig. 3.26(a), does not correspond exactly to what has been *defined* as being the resonance, that is, when the input reactance X_{IN} is zero.

As the coupling is reduced, the diameter of the loaded reflection locus decreases and shifts more towards the edge of the Smith Chart, as shown in Fig. 3.26(b), which is a locus that is representative of the physical circuit under test. Not surprisingly, this causes the resonant frequencies f_s and f_p to approach each other, since the length of the locus above the $X_{IN} = 0$ line shrinks as the coupling is reduced. This has also been noted in [7], although the mechanism leading to this observation has not been explained. It must be said that this phenomenon is due solely to the coupling capacitors and is not an inherent property of the ring itself.

If the coupling is further reduced, there will come a point where the two previously defined resonant frequencies coincide, that is, the locus will only touch the $X_{IN} = 0$ line at one frequency. Reducing the coupling further still, the locus will no longer cross the $X_{IN} = 0$ line. Conceptually this is reassuring, since increasing the coupling gap means that the input feed line begins to resemble an open-ended microstrip, which is inherently capacitive. This leads to a dilemma, since the previously held definition of resonance no longer identifies *any* resonant points. And yet the transmission magnitudes on the VNA will still show transmission peaks (albeit small ones) that are consistent with resonance, since there will still be a location on the locus that is closer to the centre of the Smith Chart than the rest.

For the two-port loaded resonator, the notion of being able to identify distinct series and parallel resonance points defined by $X_{IN} = 0$ must be abandoned. Instead, the resonant frequency should be redefined as corresponding to a peak/trough in the

transmission/reflection magnitudes, which is *always* visible on the VNA. But this latter approach is problematic because (3.9), which leads to the evaluation of the dielectric constant, applies only when the *unloaded* ring is at its parallel resonance whereas the observed resonance on the VNA does not occur at this same frequency. It is this difference that will now be assessed and is termed the *frequency correction* rule. A rule will be sought that will allow the observed resonant frequency f_0 to be converted to the unloaded ring's resonant frequency f_{0UP} . The rule is obtained by using the lumped element model of Fig. 3.13(b).

Simulated comparisons show that the *unloaded* parallel resonant frequency is up to 0.46% higher than the observed *loaded* resonant frequency (as defined by peaks/troughs in the transmittance/reflectance) throughout the first seven harmonics that are used in this work. The exact value depends on the materials and gap size used for a given ring resonator. Table 3.3 lists the resonant frequencies and permittivity values thus obtained for the *Ferro1* resonator (a similar table is obtained for the *Ferro2* resonator), and Fig. 3.27(a) plots the relative permittivity versus frequency using the data from both of these resonators.

Table 3.3 *Frequency correction and permittivity of the Ferro1 ring resonator.*

N	1	2	3	4	5	6	7
f_0 observed [GHz]	5.70	11.40	17.00	22.60	28.11	33.64	39.07
f_{0UP} [GHz]	5.73	11.45	17.07	22.69	28.21	33.74	39.17
ϵ_{eff} Equ. (3.9)	4.428	4.435	4.490	4.518	4.567	4.597	4.643
ϵ_r	6.17	6.14	6.18	6.17	6.19	6.18	6.19

As for the Dupont 943 system, the appropriate corrections are applied in Table 3.4, and the relative permittivity is plotted in Fig. 3.27(b).

Table 3.4 Frequency correction and permittivity of the Dupont 943 ring resonator.

N	1	2	3	4	5	6
f_0 observed [GHz]	5.64	11.29	16.83	22.34	27.79	33.20
f_{oup} [GHz]	5.67	11.34	16.90	22.41	27.86	33.34
ϵ_{eff} Equ. (3.9)	5.201	5.195	5.264	5.319	5.377	5.408
ϵ_r	7.59	7.51	7.55	7.55	7.56	7.52

Fig. 3.27 Relative permittivity obtained from (a) the Ferro ring resonators and (b) the Dupont 943 ring resonator, using frequency correction.

With extrema of 6.14 and 6.22, the average relative permittivity of the Ferro system is 6.19, which is 4.9% higher than the nominal value listed for this material system [12]. Failure to correct for the capacitive loading of the circuit would have increased the average relative permittivity to approximately 6.25, a value that has been confirmed by an external laboratory [13]. The average relative permittivity of the Dupont 943 system is 7.55, with extrema of 7.51 and 7.59 across the band.

3.1.4.6 Microstrip ring lumped element model resonant frequency prediction

The analysis of the microstrip ring concludes by testing the lumped element model using the material parameters that have been obtained from the measurements. The

resulting resonant frequency error is shown in Fig. 3.28, where resonance is defined by the peak/trough of the transmittance/reflectance and not by $X_{IN} = 0$.

Fig. 3.28 Difference between the observed and calculated resonant frequencies, the latter simulated using the experimentally obtained material parameters and the lumped element model, for (a) the Ferro A6-S system and (b) the Dupont 943 system.

The agreement between the calculated and observed resonant frequencies is excellent. The overall highest of 15 MHz occurs at 39.1 GHz while the overall highest relative error of 0.11% occurs at 5.7 GHz. For a simple lumped element model to corroborate observations throughout the Ka-band is quite remarkable. Using the new relative permittivity values marginally affects the calculated reflectance and transmittance, which remain as those shown in Figs. 3.24 and 3.25.

3.2 Laminated waveguide characterisation techniques

The second transmission medium to be characterised is laminated waveguide (LWG), which is the implementation of three-dimensional, dielectric-filled rectangular waveguide (RWG) in LTCC [14]. Canonical RWG is a very low loss medium, hence LWG may potentially offer less loss than planar transmission lines, a hypothesis that will be verified

in what follows. For laminated waveguide to be widely accepted as a medium, the behaviour of its complex propagation constant γ must be well understood. Although the propagation constant of the LWG is usually within a few percent of the RWG value, more precise modelling is nonetheless required, especially if LWG is to be used for cavity resonators. In Sections 3.2.1–3.2.3, three theoretical methods are presented for determining the imaginary propagation constant β of a LWG. In Section 3.2.4, the methods are applied to predict the resonant frequency of LWG cavities, and their accuracy is assessed by comparing the predictions with measured results. Section 3.2.5 concludes by presenting the experimental loss characterisation of LWGs based on cavity resonance analyses.

3.2.1 LWG propagation characterisation using the grid impedance method

With the narrow walls of the LWG composed of rows of vias, it is difficult to assess how well bound the TE_{10} mode is to those walls, as depicted in the H-plane cross section of Fig. 3.29.

Fig. 3.29 *Hypothetical H-plane view of wave-fronts travelling down a laminated waveguide.*

The imaginary propagation constant $\beta = \beta_z$ of the LWG is expected to equal that of a RWG having an effective width a^* , which is not necessarily given by the via centre-to-

centre width a of the LWG. The crux of the problem therefore lies in determining the extent to which the fields extend beyond the via walls of the LWG.

Instead of using Maxwell's equations directly, it is simpler to represent a RWG's fundamental TE_{10} mode as a superposition of two perpendicularly polarised plane waves, as shown in Fig. 3.30 (where *perpendicular* is with respect to the page).

Fig. 3.30 Two interfering plane waves with perpendicular polarisation to the xz -plane.

The launch angle θ is subtended between the z -axis and the unit vector of the wavenumber, $\hat{k}$. The net electric field in space is given by the sum of the two waves:

$$\vec{E} = \hat{y}E_0 e^{-jkz \cos \theta} \sin(kx \sin \theta) \quad (3.14)$$

where $k = \omega\sqrt{\mu\epsilon}$. The propagation constant along z in (3.14) is $\beta = \beta_z = k \cos \theta$. Equation (3.14) admits nulls that are co-linear to the z -axis, where $kx \sin \theta = 0, \pm\pi, \pm2\pi$, etc. Since conducting walls may be placed along electric field nulls without perturbing the fields, the single half-wave variation of the TE_{10} mode is bounded by walls at $kx \sin \theta = 0$ and π . Defining these x -locations as 0 and a , the condition for the TE_{10} mode inside a RWG becomes:

$$a = \frac{\pi}{k \sin \theta} \quad (3.15)$$

which fixes the remaining unknown, θ . Broad metal walls may be placed arbitrarily in the xz -plane without perturbing the electric field because this plane is perpendicular to the electric field.

The first theoretical characterisation method, called the *grid impedance method*, is original work that is based on the theory of the RWG *Transvar* directional coupler, in which a row of wires is used to couple two co-linear waveguides [15]. Given that rows of wires form the LWG narrow walls, the *Transvar* circuit model of a row of wires illuminated by a perpendicularly polarised plane wave may also be used to represent the narrow walls of the LWG. When the radius of the wires r is much smaller than their pitch p , the wires may be modelled as a single inductor, as first proposed by Macfarlane in 1946 [16]. When this restriction is not met, the more general model proposed by Marcuvitz may be used, shown in Fig. 3.31 [17].

Fig. 3.31 (a) Perpendicularly polarised plane wave incident on a row of wires and (b) its equivalent circuit with respect to propagation in the x -direction.

Note that the equivalent circuit is referenced with respect to propagation in the x -direction, where the angle of incidence is given by $\phi = \pi/2 - \theta$ and the characteristic impedance is $Z_x = \eta/\cos\phi$, where $\eta = \sqrt{\mu/\epsilon}$ is the wave impedance of the medium. The equations for the inductive and capacitive circuit elements are repeated here as [17]:

$$\frac{X_a}{Z_x} = \frac{p \cos \phi}{\lambda} \left\{ \ln \frac{p}{2\pi r} + \frac{1}{2} \sum_{\substack{n=-\infty \\ n \neq 0}}^{\infty} \left[\left(n^2 + \frac{2np}{\lambda} \sin \phi - \left[\frac{p \cos \phi}{\lambda} \right]^2 \right)^{-1/2} - \frac{1}{|n|} \right] \right\} \quad (3.16)$$

$$\frac{X_b}{Z_x} = \frac{p \cos \phi}{\lambda} \left(\frac{2\pi r}{p} \right)^2 \quad (3.17)$$

which are valid over the range $p(1 + \sin \phi)/\lambda < 1$, or more restrictively, $p < \lambda/2$. Further, the equations are strictly valid only for small obstacles: $r \ll \lambda$ and $p > 10r$, over which range the error of the model is estimated to be less than 10% [17].

Once the reactance values of the T-network are known, network analysis is applied to determine the Z_x -normalised input impedance to the vias, z_L , which is defined at the plane passing through the centres of the vias, marked “T” in Fig. 3.31. The surface impedance of the conductive RWG narrow wall has the form $Z_s = R_s(1 + j)$, hence the complex z_L value may be interpreted as the input impedance of a conducting narrow wall that is offset by some distance Δa from the T-plane, as shown in Fig. 3.32.

Fig. 3.32 *Equivalent rectangular waveguide having an effective width $a^* = a + 2\Delta a$.*

The normalised input impedance is given by:

$$z_L = r + jx = \frac{z_s + j \tan \beta_x \Delta a}{1 + j z_s \tan \beta_x \Delta a} = \frac{z_s + j \pi}{1 + j z_s \pi} \quad (3.18)$$

where $z_s = r_s(1 + j)$, $r_s = R_s/Z_x$, and $\pi = \tan\beta_x\Delta a$. The real and imaginary parts of (3.18) yield the simultaneous equations $r = r_s[1 + \pi(r + x)]$ and $r_s(r\pi - x\pi - 1) = \pi - x$, which, when combined, lead to the quadratic equation:

$$\pi^2(x + r) + \pi(1 - r^2 - x^2) + (r - x) = 0 \quad (3.19)$$

the solution of which leads directly to the evaluation of Δa :

$$\Delta a = \frac{\arctan(\pi)}{\beta_x} = \frac{\arctan(\pi)}{k \sin \theta} \quad (3.20)$$

The sign of Δa is usually positive for $p > 10r$, indicating that the effective width of an equivalent RWG, $a^* = a + 2\Delta a$, is larger than the physical width of the LWG. When the new width value is used in (3.15) the launch angle θ is altered, so the above procedure must be iterated until the effective width converges. Fortunately, very few iterations are required before the change in Δa becomes imperceptible.

3.2.2 LWG propagation characterisation using numerical methods

It should be emphasised that the equivalent lumped element model for the LWG walls of the grid impedance method has been obtained from the application of a variational method [17], which is one of many numerical methods that could be used to find the impedance of the row of vias. Although a moment method has also been formulated to analyse LWGs, the results cannot be easily generalised [18]. Only recently have the results of a boundary-integral method been curve-fitted to obtain a closed-form expression for the equivalent width of a LWG [19]. The following equation may be used when the via pitch obeys the inequalities $p < 8r$ and $p < \lambda/2$ [19]:

$$a^* = a - \frac{4r^2}{0.95p} \quad (3.21)$$

which, when used, is reported to yield a precision of better than $\pm 5\%$ for the TE_{10} cutoff frequency of the guide. It is of interest to note that the guide width appears to be *smaller* than a when $p < 8r$, according to (3.21), whereas the grid impedance method has shown that the guide width appears to be *larger* than a when $p > 10r$.

3.2.3 LWG propagation characterisation using a commercial solver

When the expense is not prohibitive, commercial solvers may be used to complement in-house numerical methods for solving three-dimensional electromagnetic problems. One such package, *Ansoft HFSS*, has been used previously to model LWGs [20], [21]. A LWG operating in the two-port transmission mode is easily modelled in HFSS using a few unit cells of the guide. The phase of the transmission constant is then used, along with the length of the guide d , to determine the imaginary propagation constant according to the formula:

$$\beta = \frac{\text{angle}(s_{21})}{d} \quad (3.22)$$

where the angle function returns the phase of s_{21} in radians.

Thus far only the imaginary component β of the complex propagation constant $\gamma = \alpha + j\beta$ has been considered. In addition to its bulk dielectric loss and conduction losses on its broad walls, the LWG will suffer from conduction losses in the vias and radiation loss through the apertures that make up its narrow walls. The latter two sources of loss are very difficult to model. Although the real part of (3.18) is interpreted as loss, calculations have shown that the grid impedance method overestimates the narrow wall loss of the LWG. This is partly due to the fact that the grid impedance method assumes that an identical parallel plate guide of height b exists on the other side of the narrow

walls, whereas in reality no such guide exists. The grid impedance method also ignores conduction losses in the vias because its formulation assumes that they are perfectly conducting. The numerical method of [19] also fails to mention loss, thus the only remaining tool for the prediction of loss is HFSS.

3.2.4 Resonant frequencies of $TE_{10\delta}$ modes in LWG cavities

A number of LWG cavity resonators have been fabricated in the Heraeus CT2000 and Dupont 943 LTCC systems. These resonators will serve to validate the analytical methods for calculating β . Before proceeding, the grid impedance method must be extended so that it may model cavities. Using four interfering plane waves instead of two, it may be shown that the resulting net electric field in space is:

$$\vec{E} = \hat{y}E_0 \sin(kz \cos \theta) \sin(kx \sin \theta) \quad (3.23)$$

where again, nulls occur along lines that are co-linear to the z -axis, when $kx \sin \theta = 0, \pm\pi, \pm2\pi$, etc. Hence the TE_{10} mode is again enclosed by narrow walls according to (3.15). Nulls also occur co-linear to the x -axis when $kz \cos \theta = 0, \pm\pi, \pm2\pi$, etc. If the physical cavity is chosen to exist along the positive z -axis, the $TE_{10\delta}$ mode requires the length of the guide in z to be given by:

$$d = \frac{\delta\pi}{k \cos \theta} = \frac{\delta\pi}{\beta_z} \quad (3.24)$$

where δ is the order of the mode. Resonance occurs when (3.23) and (3.24) are satisfied simultaneously using the same launch angle θ . The combination of those equations yields the resonance condition for θ :

$$\theta_{res} = \arctan\left(\frac{d}{\delta a}\right). \quad (3.25)$$

Once again, the effective RWG width extension Δa is found using the method described previously using (3.20). In addition, an effective RWG *length* extension Δd is found simultaneously using (3.20) but by replacing ϕ with θ , since the incident angle on the narrow wall is ϕ whereas the incident angle on the end wall is θ , as shown in Fig. 3.33.

Fig. 3.33 *Laminated waveguide cavity with equivalent RWG width and length extensions Δa and Δd , respectively.*

The new effective width a^* and length d^* are substituted into (3.25) and the process is iterated until Δa and Δd are converged. Again, very few iterations are required.

Two-port laminated waveguide $TE_{10\delta}$ cavities have been constructed using low loss Dupont 943 and Heraeus CT2000 LTCC tape, and their circuit parameters are listed in Table 3.5.

Table 3.5 *Circuit parameters of twelve LWG cavity resonators.*

	relative permittivity ϵ_r	waveguide height b [μm]	waveguide width a [mm]	via radius r [μm]	cavity length d [mm]	via pitch p [μm]
Dupont 943	7.5	600	a_1 : 3.00 a_2 : 2.65 a_3 : 2.10	100	d_1 : 7.60 d_2 : 6.40 d_3 : 5.00	600
Heraeus CT2000	9.1	233	2.50	100	d_1 : 5.00 d_2 : 6.50 d_3 : 10.50	p_1 : 500 p_2 : 750 p_3 : 1000

A total of three resonators have been fabricated in the Dupont system, identifiable by matching the subscripts on a and d , for instance, (a_1, d_1) is one resonator. A total of nine resonators have been fabricated in the Heraeus system, corresponding to all permutations of the d and p variables shown in Table 3.5. The two-port cavities are excited by small waveguide probes consisting of vias that descend into the guide through a small circular apertures in the upper broad walls of the cavities. Each probe emerges at the surface of the LTCC module where it is contacted by coplanar probes, as shown in Fig. 3.34.

Fig. 3.34 *E-plane cross sectional view of the waveguide excitation showing the top layer coplanar probe pads.*

Over the frequencies of interest, the via pitch condition of $p < \lambda/2$ is satisfied for all of the resonators. The Dupont system has $p = 6r$, hence its equivalent width and length are expected to be better approximated by (3.21). Fig. 3.35 displays the percentage error between the measured and calculated resonant frequencies when the correction of (3.21) is used, and when no correction is used (when $a^* = a$ is used directly).

Fig. 3.35 *Percentage error between the measured and calculated resonant frequencies for the three Dupont 943 LWG resonators, □ - calculated using (3.21), ○ - calculated without any correction.*

A positive error indicates that the calculated resonant frequency estimate is too high, in other words, the estimated guide dimensions are too small. Conversely, a negative error indicates that the estimated guide dimensions are too large. Referring to Fig. 3.35, although the magnitude of the relative error of the uncorrected calculation is small, the use of (3.21) further improves it, to well within 1%. There does not appear to be a strong correlation between the resonant frequency error and the resonator's a or d parameters. The grid impedance method, not shown, yields a much higher relative error since $p = 6r$ is beyond that method's required condition of applicability, $p > 10r$. Similar results have been noted for the three $p_1 = 5r$ resonators fabricated using the Heraeus CT2000 system.

For the other Heraeus resonators, the equivalent widths and lengths of the $p_2 = 7.5r$ resonators are expected to be better approximated by (3.21), whereas those of the $p_3 = 10r$ resonator should be in better agreement with the grid impedance method. Fig. 3.36 displays the relative error between the measured and calculated resonant

frequencies, the latter calculated by (3.21), by the grid impedance method, and without any correction, for the $p_2 = 7.5r$ Heraeus cavities.

Fig. 3.36 Percentage error between the measured and calculated resonant frequencies for the three Heraeus CT2000 LWG resonators having $p = 7.5r$, $\triangle$ - calculated using the grid impedance method, $\square$ - calculated using (3.21), and $\circ$ - calculated without any correction.

As expected, given the $p = 7.5r$ condition, (3.21) better estimates the actual resonant frequency than does the grid impedance method. However, it is the *uncorrected* resonant frequency that now offers the best estimate, to within $\pm 1\%$ over the entire band. This suggests that the TE_{108} modes are confined near the centres of the vias forming the narrow walls. Again, no strong dependence on the resonator's d parameter is visible, rather, the error appears to be more a function of frequency. Finally, Fig. 3.37 displays the relative error for the three $p = 10r$ Heraeus resonators.

Fig. 3.37 Percentage error between the measured and calculated resonant frequencies for the three Heraeus CT2000 LWG resonators having $p = 10r$, $\triangle$ - calculated using the grid impedance method, $\square$ - calculated using (3.21), and $\circ$ - calculated without any correction.

As expected, the grid impedance now offers the best resonant frequency estimate across the band, since $p = 10r$ is now within this method's range of applicability and is outside that of (3.21) ($p < 8r$). It should be noted however that the relative errors at low mode orders are comparatively high, well beyond the $\pm 1\%$ previously achieved.

Based on these results, the equivalent RWG guide dimension correction factor appears to depend strongly on the p/r ratio and to a lesser extent on the frequency of operation. Negligible dependence was found on the width and length dimensions d and a of the resonator. As the frequency of operation (mode number) increases, the fields are better contained within the confines of the guide—this is evidenced by noting the slope of the error of the uncorrected frequency estimates of Figs. 3.35 through 3.37. When the pitch is small (low p/r ratio), the fields are contained within a region that is smaller than the physical centre-to-centre spacing of the vias. By increasing the pitch, the fields become less bound to the physical structure and extend beyond the centre-to-centre via spacing.

3.2.5 LWG attenuation constant and substrate loss characterisation

Having characterised the imaginary component of the propagation constant, attention is now focussed on its real part, the attenuation constant of the LWG. Given that propagation in a LWG closely resembles that in a RWG, this section begins with a review of RWG loss, which is then shown to be applicable to LWG structures. Methods are then presented by which the radiation quality factor Q_r of a LWG may be obtained from finite element simulations. Measured cavity resonances are then used to determine the loss and the quality factors of laminated waveguides.

The loss of a RWG may be expressed in the form of an attenuation constant α or as a quality factor Q . The fields within a RWG can be derived analytically, making the evaluation of its loss a straightforward task. Using the perturbation method, the conductor loss of a length of RWG in the transmission mode (matched ports) is given by [1:p.147]:

$$\alpha'_c = \frac{4R_s}{abZ_{TE}} \left[\frac{b\pi^2}{2a^2\beta^2} + \frac{a}{4} + \frac{\pi^2}{4a\beta^2} \right] \quad (3.26)$$

where R_s is the surface resistance and $Z_{TE} = \eta k/\beta$ is the longitudinal (z) wave impedance, where $\eta = (\mu/\epsilon)^{1/2}$ is the wave impedance, β is the longitudinal propagation constant, and k is the TEM wavenumber. Equation (3.26) represents the conductor loss in the transmission mode only and does not represent the conductor loss of a cavity at resonance.

Fig. 3.38 A rectangular waveguide cavity.

Referring to Fig. 3.38, it may be shown that the total conductor loss of a RWG cavity at resonance may be expressed as the following:

$$P_c = \frac{abdE_0^2}{4Z_{TE}} \left[\frac{2R_s}{dZ_{TE}} + \alpha'_c \right] \quad (3.27)$$

where E_0 is the maximum electric field strength inside the cavity. The additional term inside the brackets represents the conductor loss of the end reflector walls, which has not been included in α'_c since a transmission line has no such end walls. It is convenient to define an equivalent attenuation constant that includes all conductor losses in the cavity at resonance, which must be given by:

$$\alpha_c = \frac{2R_s}{dZ_{TE}} + \alpha'_c \quad (3.28)$$

The fact that the first term is divided by the length of the cavity d explicitly shows how the finite end wall losses are being distributed over the length. It is of interest to expand (3.26) and (3.28) to explicitly show the dependency on cutoff, which may be done using the set of formulae:

$$\begin{aligned} F &= \sqrt{1 - (f_c/f)^2} & (a) \quad \beta &= \frac{2\pi f \sqrt{\epsilon_r} F}{c} & (b) \quad k &= \frac{2\pi f \sqrt{\epsilon_r}}{c} & (c) \\ a &= \frac{c}{2f_c \sqrt{\epsilon_r}} & (d) \quad d &= \frac{c}{2f \sqrt{\epsilon_r} F} & (e) \quad Z_{TE} &= \frac{\eta_0}{\sqrt{\epsilon_r} F} & (f) \end{aligned} \quad (3.29)$$

where δ is the desired $TE_{10\delta}$ mode number and η_0 is the wave impedance of free space. Substitution of these into (3.28) yields the following expanded expression for the equivalent attenuation constant:

$$\alpha_c = \frac{8R_s \epsilon_r}{c\eta_0} \left(\frac{fF^2}{2\delta} + \frac{f_c^3}{2f^2F} + \frac{cF}{8b\sqrt{\epsilon_r}} + \frac{cf_c^2}{8bf^2\sqrt{\epsilon_r}F} \right) \quad (3.30)$$

It may be shown that the conductor quality factor Q_c of the cavity obeys (3.31), which has an identical form to the dielectric quality factor Q_d , given by (3.32) [1:p.133]:

$$Q_c = \frac{k}{2\alpha_c} \cdot \frac{k}{\beta} \quad (3.31)$$

$$Q_d = \frac{1}{\tan \delta} = \frac{k}{2\alpha_d} \cdot \frac{k}{\beta} \quad (3.32)$$

It may further be shown that k^2/β has a minimum at $ff_c = \sqrt{2}$, which is why α_d also has a minimum at this frequency. Interestingly, $ff_c = \sqrt{2}$ corresponds exactly to a launch angle of $\theta = 45^\circ$, which means that both plane waves that make up the TE_{10} mode are orthogonal to each other. Defining a total attenuation constant $\alpha_T = \alpha_c + \alpha_d$, the unloaded quality factor $Q_0 = Q_{cd}$ of the RWG resonator becomes:

$$Q_0 = \left(\frac{1}{Q_c} + \frac{1}{Q_d} \right)^{-1} = \frac{k^2}{2\alpha_T \beta} = \frac{\pi\sqrt{\epsilon_r}}{c\alpha_T} \cdot \frac{f}{\sqrt{1 - (f_c/f)^2}} = \frac{\pi\sqrt{\epsilon_r}}{c\alpha_T} \cdot \frac{f}{F} \quad (3.33)$$

where f_c is the cutoff frequency of the guided mode, in this case the TE_{10} mode. Equation (3.33) is also applicable to TEM lines, where $\beta = k$, yielding the familiar expression

$$Q_0^{TEM} = \frac{k}{2\alpha_T} = \frac{\pi\sqrt{\epsilon_r}}{c\alpha_T} \cdot f \quad (3.34)$$

Since $\beta < k$ for a TE mode, given a same total loss α_T , a TE mode resonator will yield a slightly higher Q than that of a TEM mode resonator. Ignoring dielectric loss variations,

the conductor loss in a TEM line varies as $\sqrt{f}$, hence its Q increases as $\sqrt{f}$. Owing to the complexity of the conductor loss in a TE mode, the conductor loss in such a cavity does not obey such a simple rule. Substituting (3.30) into (3.33) and ignoring dielectric losses yields

$$Q_c = \frac{\pi f \eta_0}{8R_s \sqrt{\epsilon_r}} \cdot \left(\frac{fF^3}{2\delta} + \frac{f_c^3}{2f^2} + \frac{cF^2}{8b\sqrt{\epsilon_r}} + \frac{cf_c^2}{8bf^2\sqrt{\epsilon_r}} \right)^{-1} \quad (3.35)$$

Given a certain frequency of operation f , (3.35) suggests that a lower cutoff f_c (which also tends $F \rightarrow 1$) will yield a higher Q . Hence the guide should be operating at a frequency that is as far away from cutoff as is possible, which normally implies that it should be operating just below the TE_{20} cutoff frequency. This does not mean that the guide losses decrease as a function of frequency, but rather, as a function of frequency, the rate increase of stored energy is greater than the rate increase of loss. It is also of interest to note that (3.35) does not tend to zero as $f \rightarrow f_c$, but the limit is given by:

$$\lim_{f \rightarrow f_c} Q_c = \frac{\pi \eta_0}{4R_s \sqrt{\epsilon_r}} \cdot \left(1 + \frac{a}{2b} \right)^{-1} \quad (3.36)$$

which decreases as the cutoff frequency is lowered (increasing a). It is straightforward to prove that (3.36) is identical to the conductor quality factor of a parallel plate TM_0 resonator of length a , as it is expected to be, since the x -directed waves inside a RWG at cutoff are identical to the TM_0 parallel plate mode.

Consider a silver RWG that is filled with a dielectric of $\epsilon_r = 7.5$ having a loss tangent of $2.5(10^{-3})$. Its resulting conductor and dielectric quality factor Q_{cd} is plotted in Fig. 3.39(a).

Fig. 3.39 Conductor and dielectric quality factors of a RWG when the loss tangent is (a) constant at $\tan \delta = 0.0025$ and (b) linearly increasing 0.0025–0.0033 over 15–70 GHz.

A constant loss tangent causes Q_{cd} to monotonically increase as a function of frequency. Also confirmed is that a guide having a lower cutoff will yield the highest Q at a given frequency, and that this same guide has the lowest Q at cutoff. If the loss tangent is given a linear frequency dependence, the graph of Fig. 3.39(b) results, where $\tan \delta$ now increases from 0.0025 at 15 GHz to 0.0033 at 70 GHz. Instead of a monotonic increase, Q_{cd} now displays a maximum.

Laminated waveguides do not behave exactly as RWGs do because the loss on their narrow walls is due to via conductor loss and radiation through the walls. The absence of analytical formulae for this loss requires a finite element simulator to be used to predict it. The simplest geometry to simulate consists of a few unit cells of the LWG in its transmission mode. Although complete LWG cavities may also be simulated, they require very large amounts of memory, especially at higher order modes.

To find the quality factors in the transmission mode, the structure is simulated with and without dielectric and conductor losses, allowing $Q_0 = Q_{cdr}$ and Q_r to be isolated separately by examining the real part of the transmission coefficient, $e^{-\alpha l}$. Although simulating LWGs in the transmission mode yields an estimate of Q_r , there is no guarantee that this value will be applicable to a LWG cavity, which presents a standing wave to the medium external to its narrow walls and not a travelling wave. Hence a LWG cavity may radiate differently than a section of LWG transmission line. This is akin to a travelling wave antenna radiating differently than a standing wave antenna. Therefore the cavity simulation method, significantly slower, simulates an entire LWG resonator that is excited by some arbitrarily-placed, idealised probe. The loaded Q of the cavity is then determined from the swept-frequency s-parameters and the external loading of the probe is removed using (c.f. Appendix C):

$$Q_0 = Q_{cdr} = \frac{Q_L}{1 + |s_{11}^{res}|} \quad (3.37)$$

where s_{11}^{res} is the simulated reflection coefficient at resonance. HFSS further provides the total power radiated by the structure, P_{rad} , which is equal to the accepted power times the radiation efficiency $P_{rad} = \eta_{rad} P_{acc}$, where the accepted power is given by $P_{acc} = P_{avail} \left(1 - |s_{11}^{res}|^2\right)$. By exciting the port with an available power of one watt (the HFSS default), the radiation efficiency may be easily calculated via

$$\eta_{rad} = \frac{P_{rad}}{1 - |s_{11}^{res}|^2}. \quad (3.38)$$

The remaining power delivered to the circuit must be that due to conductor and dielectric losses, or $P_{cd} = (1 - \eta_{rad})P_{acc}$. At resonance, the unloaded quality factor must be equal to $Q_{cdr} = Q_0 = 2\omega_0 W_e / P_{acc}$, which results in the following equations for Q_{cd} and Q_r :

$$Q_{cd} = \frac{2\omega_0 W_e}{P_{acc}(1 - \eta_{rad})} = \frac{Q_0}{1 - \eta_{rad}} \quad (3.39)$$

$$Q_r = \frac{2\omega_0 W_e}{P_{acc}\eta_{rad}} = \frac{Q_0}{\eta_{rad}} \quad (3.40)$$

The radiation Q must be much greater than Q_{cd} if the LWG cavity is to be used for material characterisation purposes. The transmission mode and cavity simulation methods will now be compared. Under investigation is a LWG cavity made from Dupont 943 LTCC, the parameters of which are listed in Table 3.6.

Table 3.6 *LWG circuit parameters for radiation loss simulations.*

ϵ_r	waveguide width a [μm]	waveguide height b [μm]	loss tangent $\tan\delta$	metal conductivity σ [S/m]	via pitch p [μm]	via radius r [μm]
7.5	2300	660	0.0025	6.1×10^7	600	100

The length of the guide for the transmission mode simulations is $d = 2400 \mu\text{m}$, while the length of the cavity for the resonant simulation is $d = 5700 \mu\text{m}$, the longer length being required to establish resonance. Both structures, as simulated, are illustrated in Fig. 3.40.

Fig. 3.40 *LWG structures for radiation loss simulations: (a) cavity mode and (b) transmission line mode.*

H-plane symmetry is invoked in both cases to reduce the problem size, whereas E-plane symmetry is further invoked in the transmission line simulation. One of the main advantages of the transmission mode simulation is that data may be easily gathered over a broad frequency range without modifying the geometry. Attenuation and quality factors obtained using the transmission line method are shown in Fig. 3.41.

Fig. 3.41 Attenuation constants and quality factors extracted using the transmission line simulation method with the parameters of Table 3.6 and a guide length $d = 2.4$ mm.

The conductor loss is very nearly identical to that of a RWG, meaning that the vias do not significantly alter α_c . This means that the RWG conductor loss equations are valid approximations for the LWG. The radiation attenuation constant is a small fraction of the total attenuation, representing at most 5.5% of the total loss of the LWG in the transmission mode. The radiation Q also displays a maximum at $\sqrt{2}f_c \approx 37$ GHz. The dielectric Q is not plotted in Fig. 3.41(b) since its value remains constant at $1/\tan\delta = 400$, again identical to that of a RWG. For sake of comparison, cavity resonance analyses are presented in Table 3.7 for the second and third order resonance modes.

Table 3.7 Resonant cavity quality factor analysis results using the parameters of Table 3.6 and a guide length of $d = 5.7$ mm.

order N	res. frequency f_0 [GHz]	loaded Q_L	unloaded Q_0	rad. efficiency η_{rad} [%]	radiation Q_r	conductor Q_c
2	31.11	172	255	7.4	3453	890
3	37.93	240	289	7.4	3897	1418

Similar to the radiation attenuation constant in the transmission mode, the radiation efficiency remains constant as a function of frequency, although the percentage of radiation loss is slightly higher than the previous value of 5.5%. This directly impacts the radiation Q , which, although lower, still displays a higher value near the $\sqrt{2}f_c$ frequency. The Q_c values of 890 and 1418 are approximately those of the transmission mode simulation, 1135 and 1400, respectively, even though the cavity simulation has extra conductor loss at the end walls. Note, however, that end wall loss only represents a very small fraction of the total conductor loss, even in RWGs. The radiation Q is significantly lower in the cavity simulation than it is in the transmission simulation, hence the former must always be used to assess the radiation properly.

Measuring the radiation Q of a physical LWG resonator is a very difficult task since it entails the measurement of radiation patterns in an anechoic chamber over a broad swath of pattern cuts. Instead, HFSS simulations will be used to estimate the radiation. The experimental results will then be corrected using the simulated values of Q_r . Four two-port cavity resonators have been fabricated in Dupont 943 and their parameters are listed in Table 3.8, and a typical response of the resonators is plotted in Fig. 3.42.

Table 3.8 Dupont 943 LWG cavities and their resonant frequencies.

resonator name	width a [mm]	length d [mm]	order δ	observed f_{res} [GHz]	Q_0 (measured)
<i>res1</i>	3.00	7.60	3	28.58	294
			4	34.44	270
			5	40.79	254
<i>res1a</i>	<i>ibid.</i>	<i>ibid.</i>	3	28.54	280
			4	34.39	279
			5	40.70	257
<i>res4</i>	2.65	6.40	2	27.21	282
			3	33.37	247
			4	40.49	241
<i>res2</i>	2.30	5.70	2	31.04	251
			3	37.85	262

Fig. 3.42 Reflection and transmission magnitude of the Dupont 943 res4 resonator.

The material is assumed to have a relative permittivity of 7.5, a loss tangent of 0.25%, and silver conductors. The via radius is 100 μm and the via pitch is 600 μm . The unloaded quality factor of the resonator is found by first measuring Q_L and then removing the external loading via (c.f. Appendix C):

$$Q_0 = Q_{cdr} = \frac{Q_L}{1 - |s_{21}^{res}|} \quad (3.41)$$

where s_{21}^{res} is the transmission coefficient at resonance. The frequency sweep step size is 17.5 MHz, which yields a relative uncertainty of Q_L on the order of $\pm 6\%$ near 30 GHz. This uncertainty is improved upon by using a linear interpolation to measure the bandwidth.

In order to compare the LWG Q_0 values with those of RWG, the radiation loss of the LWG must be removed. HFSS cavity simulations have been performed to assess the radiation Q of the cavities at their resonant frequencies. The Q_r values obtained range in the thousands and have been used to correct the experimental LWG Q_0 values, although the change in Q that this precipitated was no more than 20. Fig. 3.43 displays the resulting conductor and dielectric Q values and attenuation constants of the LWG cavities as discrete data points.

Fig. 3.43 Simulated RWG and measured LWG conductor and dielectric (a) Q ; (b) attenuation constant.

Fig. 3.43(a) shows that there is a decreasing trend to the Q_{cd} values, which is indicative of an increasing loss tangent. Also plotted, as lines in Fig. 3.43, are simulations of RWG cavities having silver conductors and a loss tangent that increases linearly from 0.0026 at 26 GHz to 0.0036 at 45 GHz. Given the almost equal values obtained in Fig. 3.43(a), no distinction is made between the various guide widths; this distinction is more clearly illustrated in Fig. 3.43(b), which plots the resulting attenuation constants, expressed in dB/cm. The simulated and measured results are in very good agreement. The Q_{cd} metric is dominated by the dielectric losses and is therefore very sensitive to the loss tangent. Because of this, good agreement between the simulated and measured values is obtained within a loss tangent window of only $\pm 2(10^{-4})$. Such a tight range of loss tangent values makes the cavity method superior to the microstrip ring resonator for dielectric loss evaluations. Fig. 3.44 is a superposition of the loss tangent obtained for Dupont 943 using ring resonators and LWG cavities.

Fig. 3.44 Dupont 943 loss tangent obtained from a ring resonator and by LWG cavities.

Only one ring resonator measurement was available for the Dupont 943 system and its noisy response makes curve fitting very difficult. The eleven points that have been used to obtain the fitted LWG curve are significantly less noisy and admit a loss tangent uncertainty of only $\pm 2(10^{-4})$, thereby confirming their higher reliability.

3.3 Summary

This chapter has presented many different techniques for the characterisation of transmission lines and resonators, which lead directly to the evaluation of the LTCC material parameters. Beginning with non-resonant characterisation techniques for microstrip, it has been found that the transitions between the measurement probes and the transmission line under test significantly perturb the transmission coefficient, adding unwanted phase and loss to the measured data. The Absolute Magnitude method has no way of compensating for this additional loss and always overestimates the attenuation of the transmission line. The Line Comparison method corrects for the increased loss by

comparing the insertion loss of two lines of differing lengths; however, loose LTCC tolerances and poor measurement repeatability cast doubts on this method's reliability. The same issues plague the TRL calibration method, which has yet to prove itself capable of properly estimating the attenuation of LTCC microstrip lines.

Using ring resonators, the microwave properties of microstrip lines have been obtained from measurements in the 5–40 GHz range. The simplicity of the method and of the ring resonator itself lend themselves well for general use. The relative permittivity of the Ferro A6-S substrate ranges from 6.14 to 6.19 and the total attenuation increases from 0.11 to 0.42 dB/cm up to 40 GHz for 50 Ω microstrip lines. These values are consistent with the material specifications and have been corroborated by an external source. The relative permittivity of the Dupont 943 ranges between 7.51 and 7.59 whereas its loss increases from 0.11 to 0.43 dB/cm up to 33 GHz. Theoretically, an extended lumped element model for the two-port co-linear fed ring resonator has been successfully applied. The most simple derivation of the unloaded ring's input impedance has been presented, and a method for extracting the capacitance values of the gap-coupled resonator from simple 2.5D simulations has also been presented. The method not only leads to excellent resonant frequency predictions, but also leads to good reflectance and transmittance values. For the first time, the frequency pushing effects of the lumped element coupling circuits have been assessed and used to correct the observed resonant frequency. Finally, the mechanism responsible for the input impedance behaviour of the loaded ring has also been discussed.

The second half of this chapter has focussed on laminated waveguides, and analytical techniques have been used to estimate the equivalent RWG dimensions of a LWG.

The techniques are often capable of estimating LWG cavity resonant frequencies to within 1% of the measured values. When the via pitch is smaller than 7.5-times the via radius, $p < 7.5r$, the correction formula given by (3.21) yields the best estimate of the equivalent RWG dimensions, which are smaller than the physical ones. When $8r < p < 10r$, the fields are bound very near to the actual centre-to-centre spacing of the physical guide, meaning that no correction need be applied to find the equivalent dimensions. When the via pitch is large, $p > 10r$, the grid impedance method yields the best estimate.

The conductor and dielectric losses of laminated waveguide have been found to be nearly identical to that of rectangular waveguide when the via pitch is in the range of $p \leq 6r$, allowing RWG formulae to be used to estimate these values. Laminated waveguide suffers from an additional radiation loss component that can only be calculated by using FEM simulations, and a technique for doing so has been presented. When applied to physical measurements of Dupont 943 cavities, the radiation loss correction minimally affects the unloaded quality factor, which is dominated by dielectric losses. This makes the analysis of LWG cavities a superior method for assessing dielectric loss tangents. The loss tangent of the Dupont 943 substrate is estimated to increase linearly from $2.6(10^{-3})$ to $3.6(10^{-3})$ over the 26–45 GHz range, yielding LWG attenuation constants of 0.40–0.55 dB/cm, which are similar to the microstrip line values.

3.4 Chapter Three references

- [1] D. M. Pozar, *Microwave Engineering*, 1st ed., Reading, MA: Addison-Wesley, 1990.
- [2] G. F. Engen & C. A. Hoer, "Thru-reflect-line: an improved technique for calibrating the dual six-port automatic network analyzer", *IEEE Trans. Microwave Theory Tech.*, vol. 27, pp. 987–993, Dec. 1979.

- [3] R. B. Marks, "A multilayer method of network analyzer calibration", *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1205–1215, July 1991.
- [4] H. M. Barlow & A.L. Cullen, *Micro-wave Measurements*, London: Constable and Company, 1950.
- [5] T. C. Edwards & M. B. Steer, *Foundations of Interconnect and Microstrip Design*, 3rd ed., Chichester: John Wiley & Sons, 2000, sect. 9.7.
- [6] J. Carroll, M. Li, and K. Chang, "New technique to measure transmission line attenuation", *IEEE Trans. Microwave Theory Tech.*, vol. 43, Jan. 1995, pp. 219–222.
- [7] K. Chang, *Microwave Ring Circuits and Antennas*, New York: John Wiley & Sons, 1996, chap. 2.
- [8] C.-C. Yu & K. Chang, "Transmission-line analysis of a capacitively coupled microstrip-ring resonator", *IEEE Trans. Microwave Theory Tech.*, vol. 45, Nov. 1997, pp. 2018–2024.
- [9] L.-H. Hsieh & K. Chang, "Equivalent lumped elements G , L , C , and unloaded Q 's of closed- and open-loop ring resonators", *IEEE Trans. Microwave Theory Tech.*, vol. 50, Feb. 2002, pp. 453–460.
- [10] D. A. Rudy, J. P. Mendelsohn, and P. J. Muniz, "Measurement of the RF properties with series resonant microstrip elements", *Microwave J.*, March 1998, pp. 22–41.
- [11] K. C. Gupta, R. Garg, and I. J. Bahl, *Microstrip Lines and Slotlines*, Dedham, MA: Artech House, 1979, chap. 3.
- [12] A6-S Tape, LTCC Tape System, Ferro Electronic Materials specification data sheet, revision 1000, year 2000.
- [13] R. Kulke *et al.*, "Investigation of ring-resonators on multilayer LTCC", contribution to workshop "Advances in ceramic interconnect technologies for wireless, R.F. and microwave applications", *IEEE MTT-S Int. Microwave Symp., IMS 2001*.
- [14] H. Uchimura, T. Takenoshita, and M. Fujii, "Development of a laminated waveguide", *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 2438–2443, Dec. 1998.
- [15] K. Tomiyasu & S. B. Cohn, "The transvar directional coupler", *Proc. IRE*, vol. 41, pp. 922–926, July 1953.
- [16] G. G. Macfarlane, "Surface impedance of an infinite parallel-wire grid at oblique angles of incidence", *J. IEE*, vol. 93, part IIIA, no. 10, pp. 1523–1527, 1946.
- [17] N. Marcuvitz, *Waveguide Handbook*, MIT Radiation Laboratory Series, vol. 10, New York: McGraw-Hill, 1951, pp. 285–289.
- [18] J. Hirokawa & M. Ando, "Single-layer feed waveguide consisting of posts for plane TEM wave excitation in parallel plates", *IEEE Trans. Antennas Propagat.*, vol. 46, pp. 625–630, May 1998.
- [19] Y. Cassivi, L. Perregrini, P. Arcioni, M. Bressan, K. Wu, and G. Conciauro, "Dispersion characteristics of substrate integrated rectangular waveguide", *IEEE Microwave and Wireless Comp. Lett.*, vol. 12, pp. 333–335, Sept. 2002.
- [20] D. Deslandes & K. Wu, "Integrated transition of coplanar to rectangular waveguides", in *IEEE MTT-S Int. Symp. Dig., IMS 2001*, vol. 2, pp. 619–622.
- [21] J. Bray & L. Roy, "Laminated waveguide couplers in LTCC technology", in *Proc. Symp. on Antenna Technology and Applied Electromagnetics, ANTEM 2000*, pp. 527–530.

Chapter Four — Characterisation of the experimental

Ferro ferrite LTCC tape

Ferrite is a crystalline magnetic oxide that is used in a multitude of circuit applications. The suitability of a particular ferrite circuit element depends largely on the operating frequency, ranging from low frequency transformers and inductors, to RF chokes, and finally to microwave circulators and phase shifters [1]–[3]. Perhaps the greatest barrier to more elaborate ferrite circuit designs is that the material has remained incompatible with standardised, high volume planar integrated circuit fabrication, even though extensive research has been conducted on the deposition of ferrite films [4]–[6]. Given that this scenario is unlikely to change, an attractive alternative for the integration of ferrite is in the packaging material. The dimensional scale of a package is often orders of magnitude larger than the chips housed within it, a property that facilitates microwave design since the length of practical ferrite devices is on the order of a wavelength.

Three major LTCC suppliers were contacted in 2001 regarding the availability of ferrite materials, namely Dupont, Ferro, and Heraeus. A Dupont representative indicated that the company had no commercial interest in ferrite, and the same reply was received from Heraeus, although in 2002 it was revealed that Heraeus had in fact begun internal investigations into ferrite tape. Only Ferro Corporation replied that ferrite LTCC tape was available, although it was an experimental material that had not been marketed nor even fully characterised. Given that Ferro still had a limited quantity of this ferrite tape in stock, the company graciously donated two rolls of it for evaluation purposes. Sufficient materials and resources were available for only one 5×5 cm panel to be fabricated,

which meant that all of the circuit elements had to be included in one single design layout.

This chapter presents the characterisation methods for determining the suitability of ferrite LTCC for microwave phase shifter applications. Saturation and remanent magnetisation, substrate loss tangents, and dielectric constant are the essential material properties. The circuits that are used to characterise these properties have been built using the actual ferrite LTCC under test. This *in situ* methodology differs from other, more complicated *ex situ* methods that normally use a blank sample to determine its properties [7]. A recent example of such a method for millimetre-wave characterisation uses free-space magneto-optical effects [8]. While this remains a powerful technique, the required test bench is very elaborate and includes a very large electromagnet. Contrarily, the methods presented herein emphasise simplicity, requiring little more than a network analyser and an oscilloscope. While the more complicated methods are able to isolate permittivity and permeability, microwave circuitry is affected only by their product, hence in many practical applications the two need not be known separately.

The processing of the experimental ferrite tape is similar to the standard method used for Ferro's A6-S LTCC system. The use of standard processing techniques is essential for reducing the cost of high volume fabrication, and Section 4.1 provides an overview of the ferrite LTCC fabrication process. Section 4.2 presents the magnetostatic characterisation of the material, that is, the evaluation of the ferrite hysteresis curve, which has been obtained using a compact solenoidal transformer embedded entirely within the ferrite LTCC. Knowledge of the hysteresis properties of the ferrite is crucial since the gyromagnetic interaction between the ferrite and a microwave signal is directly

proportional to the magnetisation of the material. The temperature dependence of the magnetostatic properties has also been measured. Sections 4.3 and 4.4 present the microwave characterisation of the ferrite, including the measurement of the combined relative permittivity/permeability and of the loss tangent of the tape at frequencies ranging from 6 to 40 GHz. These properties have been obtained using a microstrip ring resonator and laminated waveguide cavities. Finally, the conclusion summarises the findings.

4.1 Ferrite tape processing

The fabricated test panel consists of ten layers of ferrite, each layer having a green thickness of 142 μm (5.6 mils). Although no conductor pastes are specifically made for the Ferro F180 experimental ferrite, the silver paste materials of the Ferro A6-S LTCC system have been recommended and have been used, including via fill CN33-343, inner conductor CN33-398, and surface conductor CN33-391. The punching of the 150- μm diameter via holes in the tape sheets was successful, as was the via filling process. However, the silver conductor paste binder and solvent systems are not quite compatible with the ferrite tape, as is understandable with developmental systems. Specifically, the drying of the paste solvent caused the underlying tape sheet to wrinkle. This has been almost eliminated by drying the paste at room temperature. The layer stack was laminated isostatically and fired at 850 °C using the material supplier's recommended profile, as shown in Fig. 4.1.

Fig. 4.1 *Firing profile of the experimental Ferro F180 ferrite LTCC.*

The temperature increases by 2 °C/min up to the burnout temperature of 450 °C and is held there for two hours, during which the volatile organic binders contained in the tape are vented. The temperature is then increased by 6–8 °C/min up to the sintering temperature of 850 °C, which is held for a minimum of ten minutes before cooling. Prototypes of fired ferrite parts have been diced and visually inspected under a microscope, photographs of which are shown in Fig. 4.2.

Fig. 4.2 *Microscope photographs of prototype ferrite LTCC modules.*

The surface and cross-section of the illustrated sample are in excellent condition: the surface conductors, buried conductors, and vias are all well defined. Further, no delamination between the layers is apparent. Given the successful prototyping, an entire 5×5 cm panel was subsequently fabricated, containing the characterisation structures that are described forthwith.

4.2 Measurement of the ferrite hysteresis curve

A hysteresis measurement involves plotting the magnetic flux density B as a function of the magnetic field H and is denoted by $B(H)$. For linear applications below 50 MHz, a hysteresis measurement is required to determine a ferrite's initial relative permeability, nonlinearity, and hysteresis loss. At microwave frequencies the ferrite is usually operated at remanence or at saturation, so key hysteresis properties include the saturation magnetisation, squareness, and coercivity. A clear advantage of using LTCC is that conductors may be printed within the ferrite, permitting windings to be constructed deep inside the ferrite material. A transformer is constructed by wrapping two windings around a common core and may be used to determine the magnetostatic properties of the ferrite [9:p.110]. In the following subsection, a theoretical description of hysteresis measurements using a transformer is provided. A transformer design that adheres to the LTCC design rules will then be presented. Experimental measurements of the ferrite transformer are subsequently presented, along with the results of the magnetostatic characterisation of the ferrite material.

4.2.1. Theoretical operation of the transformer for hysteresis measurements

The simplified circuit that will be used for hysteresis measurements is shown schematically in Fig. 4.3, where the transformer core is filled with the ferrite under test.

Fig. 4.3 *Simplified transformer circuit for $B(H)$ measurements. Connection to the external circuit elements is done through wafer probes.*

The primary side consists of a signal generator and an audio power amplifier capable of delivering at least 500 mA of rms current. The resistor R_1 has a value of $3.5 \, \Omega$ and consists of three half-watt parallel resistors. The limit of 1.5 W restricts the primary current to values below 650 mA rms. When a current I_p flows through the primary winding of an ideal transformer, Ampere's Law may be invoked to yield the magnetic field H inside the transformer core:

$$H = \frac{N_1 I_p}{\ell} = n_1 \frac{V_H}{R_1} \quad (4.1)$$

where N_1 is the number of turns on the primary, ℓ is the length of the core, n_1 is the primary turn density expressed in turns per metre [t/m], and V_H , which is directly proportional to H , is the voltage across R_1 to be applied to the horizontal deflection of an oscilloscope. The flux density inside the core may be expressed as:

$$B = \mu_{eff} n_1 I_p \quad (4.2)$$

where μ_{eff} is the effective linear permeability of the core material, a variable that is no longer constant in the context of hysteresis (non-linear) measurements. If one assumes that the flux density inside the core's cross-sectional area A is uniform, the total flux, given by:

$$\Phi = AB \quad (4.3)$$

may be used with Faraday's Law to yield the electromotive force induced across the secondary winding:

$$V_1 = N_2 \frac{d\Phi}{dt} \quad (4.4)$$

where $d\Phi/dt$ is the temporal rate of flux change in the core and N_2 is the number of turns on the secondary. Note that V_1 will *not*, in general, be sinusoidal and hence AC voltmeter readings of V_1 should be avoided. This negates the calibration procedure presented in [9:p.111]. Upon amplifying and integrating V_1 , the output voltage V_v becomes directly proportional to the B-field,

$$V_v = G_v \int V_1 dt = G_v N_2 \Phi = G_v N_2 AB \quad (4.5)$$

which is then applied to the vertical deflection of the oscilloscope. The Lissajou figure that is obtained by displaying V_v versus V_H is directly proportional to $B(H)$, the desired hysteresis curve. In the strictest sense, a $B(H)$ curve should be static, that is, measured at DC; however, the dynamic $B(H)$ curve will be representative of the static $B(H)$ curve if the frequency is sufficiently low, in the range of fifty to a few thousand hertz.

4.2.2. Theoretical transformer design in an LTCC environment

Having described the operation of a transformer in the context of a $B(H)$ measurement, attention is now focussed on the design of a transformer using LTCC tape. With ferrite permeating the entire medium, designing a closed magnetic circuit becomes a trivial matter. Although a toroidal transformer core could be envisioned, a solenoidal core has been implemented for sake of manufacturing simplicity. To maximise the field strength within the core, (4.1) indicates that the turn density n_l within the solenoid should be as high as possible; however, the design rules of the LTCC process must be satisfied simultaneously. Consider a solenoid consisting of a series of single-turn loops around a single tape layer, illustrated in Fig. 4.4.

Fig. 4.4 *Single-loop winding having three turns per unit cell length.*

The single tape layer restricts each loop of a straight solenoid to having only three of its four segments, the fourth being sacrificed to connect to the next loop. The final depth transition in Fig. 4.4 emerges from the same corner whence the winding began, therefore four loops are required to form a unit cell. Only three current loops are ever encircled using Ampere's Law, hence the unit cell has three turns per unit cell length. If at least three tape layers are available, a double-loop winding may be conceptualised, as shown in Fig. 4.5.

Fig. 4.5 *Double-loop winding having seven turns per unit cell length.*

Although a true double-loop consist of eight segments, the eighth segment has again been used for the depth transition. Applying Ampere's Law to this structure shows that there are now seven turns per unit cell, the latter again consisting of four double-loops. The general winding relationships are given by:

$$\begin{aligned} N_L &= 2x - 1 \\ n &= 4x - 1 \end{aligned} \tag{4.6}$$

where N_L is the number of layers required, x is the number of loops at a fixed cross-section location, and n is the turn density, which rises twice as quickly as the number of tape layers required. The length of the unit cell depends only on the LTCC design rules and is the same regardless of the number of loops x chosen.

A solenoidal transformer is constructed by wrapping the primary and secondary windings around a common core. A conceptual transformer cross-section, explicitly showing five ferrite tape layers, is shown in Fig. 4.6.

Fig. 4.6 *Five layer solenoidal transformer cross-section: (a) conceptual, (b) design rule adherence.*

The black lines represent the double loop of the primary whereas the single white loop is the secondary. The mathematical derivations require that the B-field of the secondary winding be uniform throughout its cross-section and this is achieved by burying the secondary well within the primary winding. Vias should not be stacked in the LTCC process, so the modified cross-section of the solenoid is shown in Fig. 4.6(b).

4.2.3. Practical solenoid design considerations

Having described the desired structure, details concerning the actual circuit are now considered. It is imperative that the output voltage levels be above detectable thresholds regardless of the unknown properties of the ferrite material.

The LTCC ferrite panel consists of ten layers of 50×50 mm tape, each having a green thickness of $142 \mu\text{m}$ (5.6 mils). Although the shrinkage of this experimental tape is unknown, Ferro has suggested that it will be similar to the Ferro A6-S tape system, which has a 24% shrinkage of thickness during firing. This being the case, the fired thickness of the blank tape is $108 \mu\text{m}$. Similar to the A6-S tape, it will be further assumed that conductors leach into the tape by $4 \mu\text{m}$ per side. Using the smallest via diameter maximises the circuit density; however, a via diameter of $\varnothing = 150 \mu\text{m}$ has been chosen

because 100 μm vias, the minimum diameter, are difficult to punch. Vias require catch pads that are 50 μm larger than the via diameter, hence the minimum line width becomes 200 μm . The minimum via pitch on any single layer is $p = 2.5\varnothing = 375 \mu\text{m}$, meaning that the minimum pitch along the *length* of the solenoid must also be 375 μm . The unit cell therefore measures $4 \times 375 = 1500 \mu\text{m}$ and the turn densities are $n_1 = 4667 \text{ t/m}$ on the primary and $n_2 = 2000 \text{ t/m}$ on the secondary. The width of the secondary winding does not impact the magnetic design and can be freely chosen.

We now proceed with a sample calculation. If 500 mA of rms current can be safely applied to the primary, the peak magnetic field inside the core will be $H_{\text{peak}} = n_1 I_{\text{peak}} = 3300 \text{ A/m}$ (41.5 oersteds [Oe] in cgs units). The *defacto* standard for measuring saturation magnetisation is to apply a magnetic field H that is five times the coercive field value H_c . Lithium, magnesium, and nickel ferrite from suppliers such as *Trans Tech* and *EMS* have coercive fields ranging from 0.5 to 6 Oe, so if the Ferro ferrite is reasonably good, the peak magnetic field of 3300 A/m should be more than sufficient.

Using (4.5), the peak output voltage from the integrator will be $V_v = G_v N_2 A B_s$ and depends on the total length of the solenoid (sets N_2), the gain G_v of the signal amplifier, the area A of the secondary winding, and the saturation flux density of the ferrite B_s . This output voltage will be quite small owing to the small value of A . By increasing the number of secondary tape layers to four, eight out of the ten LTCC layers will be used by the transformer, as shown in Fig. 4.7, which also indicates the placement of the vias.

Fig. 4.7 Cross section of the eight layer LTCC solenoid transformer.

Since via stacking is not recommended, the secondary vias must also be staggered and the resulting area of the secondary is $A = 368.35(10^{-9}) \text{ m}^2$. If a reasonable core length is 2.25 cm (15 unit cells), then 105 turns will exist on the primary with 45 on the secondary. If the secondary had been circular, its effective radius would be 342 μm . The length of the secondary is over 30 times its effective circular diameter, hence the infinite solenoid approximation holds for the secondary. As for the primary, its length is over 12 times the effective circular diameter of its *outer* loop, so the infinite solenoid approximation should again be valid [10].

With 45 turns on the secondary, the peak output voltage becomes $V_v = 16.58(10^{-6})G_v B_s$. Common ferrite saturation flux density values range from 0.05 to 0.48 T. With a signal amplifier gain of $G_v = 30\,000 \text{ V/V}$, the output voltage V_v will be well within the range that is detectable by standard oscilloscopes, which is on the order of 5 mV per division. Given the low frequency of operation (1000 Hz), operational amplifiers may be used for both signal amplification and integration, and the complete circuit schematic is shown in Fig. 4.8.

Fig. 4.8 Complete circuit schematic for the measurement of the ferrite $B(H)$ curves.

The secondary voltage V_1 is amplified by the first 741 op-amp circuit, an inverting amplifier with a nominal gain of $G_1 = -R_3/R_2$. The second 741 circuit, comprised of R_4 , R_B , and C , forms the inverting integrator, obeying the relation:

$$V_v = \frac{-1}{R_4 C} \int V_2(t) dt \quad (4.7)$$

The integrator is susceptible to the accumulation of DC charge if any DC voltage is present at its input, which can cause the output V_v to hit the power supply rails. This problem is mitigated by carefully setting the offset null potentiometers on both op-amps such that V_v contains no DC component. The potentiometers are tied to the negative power supply rail, $V_{EE} = -V_{CC} = -10$ V. The bleeder resistor R_B also helps to discharge any accumulated static charge from the capacitor. Table 4.1 lists the component values.

Table 4.1 Component values and expected gain levels.

R_1	R_2	R_3	R_4	R_B	C	$-G_1$	$1/(R_4 C)$
3.5	21.6	670	3.2	1.0	0.312	31.0	1002
Ω	$k\Omega$	$k\Omega$	$k\Omega$	$M\Omega$	μF	V/V	V/V

Actual gain levels have been measured by replacing V_I with a small sinusoidal signal. The average amplifier gain is $-G_I = 29.9$ V/V and the average integrator coefficient is $1/(R_4C) = 1014$ V/V. The average total gain of the amplifier/integrator cascade is therefore $G_v = G_I/(R_4C) = 30.31(10^3)$ V/V and is sufficient to boost the output voltage to measurable levels. Fig. 4.9 displays the input waveform to the integrator as well as its output and confirms its proper operation.

Fig. 4.9 Measured input (V_2) and output (V_v) waveforms to and from the integrator.

With 15 unit cells, the solenoid transformer measures only 2.25 cm long and 4.6 mm wide, although the module containing it is larger, 3.5 cm long and 12.5 mm wide, to allow sufficient room for flux return paths. These modest numbers become impressive if one considers the total number of segments and vias contained within the transformer, which are listed in Table 4.2.

Table 4.2 LTCC solenoid transformer statistics.

	Number of segments	Number of vias	Total length of segments	Total length of vias
Primary	1305	1380	1.05 metres	.15 metres
Secondary	282	360	0.21 metres	0.04 metres

The primary winding contains over 1300 layer transitions and has a total path length of over 1.2 metres while the secondary has 360 transitions and a total path length of over 25 cm. A screen capture of the solenoid layout with transparent layers is shown in Fig. 4.10 along with a photograph of the fabricated module.

Fig. 4.10 *LTCC solenoid transformer: (a) layout screen capture, (b) photograph of the manufactured module.*

Fig. 4.11 is a photograph of the solenoid module on the probing station chuck, with the primary and secondary connections plainly visible.

Fig. 4.11 *Measurement of the LTCC ferrite solenoid transformer with low frequency probes.*

4.2.4. Measured ferrite permeability and hysteresis using the LTCC solenoidal transformer

The first series of measurements consists of recording the minor $B(H)$ curves, which are determined by incrementally augmenting the peak primary current and measuring the $B(H)$ curve for each current setting. This is done until the saturation, or the major $B(H)$ curve, is observed. A plot of the peak B-field versus the peak H-field for each current setting may be used to determine the permeability of the material. Although permeability is an ill-defined metric when a magnetic material is non-linear, one may still speak of *linear* permeability and *incremental* permeability, as defined in Fig. 4.12.

Fig. 4.12 Linear μ_{lin} and incremental μ_{incr} permeability are defined by the slopes of the peak B versus peak H graph.

For any given operating point indicated in Fig. 4.12 by arrows, the linear permeability is the slope of the line that extends from the origin to that operating point. The incremental permeability is the instantaneous slope of the curve and can be regarded as the permeability experienced if the maximum H-field is increased or decreased beyond that operating point. Minor $B(H)$ curves at two low H-field values are shown in Fig. 4.13.

Each curve, measured at room temperature, is a time-average of 128 triggers and is comprised of 512 data points downloaded from a Tektronix TDS 3014 digital oscilloscope.

Fig. 4.13 Minor $B(H)$ curves for two low-level H -fields: (a) $H_{peak} = 141$ A/m, (b) $H_{peak} = 382$ A/m.

The maximum H -field $H_{peak} = 141$ A/m corresponds to a primary current of 20 mA rms and results in a nearly linear $B(H)$ curve. When H_{peak} approaches 300 A/m, the opening of the $B(H)$ curve becomes detectable, as shown in Fig. 4.13(b). A total of 17 curves have been measured, leading to the major $B(H)$ curve shown in Fig. 4.14.

Fig. 4.14 Major hysteresis curve of the Ferro ferrite material.

The saturation flux density is $B_s = 230$ mT (2300 G), the remanent flux density is $B_r = 136$ mT (1360 G), and the coercive magnetic field is $H_c = 688$ A/m (8.6 Oe). The peak H-field is 3986 A/m, more than 5.7 times H_c .

Fig. 4.15 (a) Peak flux density B_{peak} versus peak magnetic field H_{peak} , (b) linear μ_{lin} and incremental μ_{incr} relative permeability versus peak magnetic field.

Peak B versus peak H is plotted in Fig. 4.15(a). Heating of the module at high current levels is a concern because magnetisation is adversely affected by heat. All curves have been acquired one second after applying the current to the primary. The final data point of Fig. 4.15(a) was obtained with a primary current of over 700 mA rms and the magnetic properties of the sample decayed very quickly at this extreme level. Fig. 4.15(b) displays the relative linear and incremental permeability of the material (relative to μ_0). The relative linear permeability peaks at 97 for a peak magnetic field of 1170 A/m whereas the relative incremental permeability peaks at 118 near the same magnetic field value. The incremental permeability indicates that there is no advantage to operating the material above a magnetic field magnitude of 3000 A/m.

As a verification, the self-inductance of the secondary winding (with the primary open) has been measured at 1 kHz with a 500 mV rms sinusoidal signal using an HP 4263B LCR Meter. Calibrating out the series inductance and resistance of the interconnecting wire, the series inductance of the secondary is 2.75 μH and its series resistance is 5.8 Ω . The self-inductance formula for a long solenoid is given by:

$$L = \frac{\mu_0 \mu_r N^2 A}{\ell} \quad (4.8)$$

Substitution of the variables for the secondary winding into (4.8) yields a relative permeability of 66. The rms voltage of 500 mV delivers a peak current of 122 mA to the secondary winding, which in turn generates a peak magnetic field of $H_{peak} = n_2 I_{peak} = 244$ A/m within its core. The relative permeability of 66 at $H_{peak} = 244$ A/m compares very well to relative linear permeability of Fig. 4.15(b), which is 64.9 at $H_{peak} = 206$ A/m. The measured series resistance of the primary winding is 14.9 Ω , meaning that the primary

dissipates 5.4 W when $I_p = 600$ mA rms, the current required to excite the major $B(H)$ curve.

Fig. 4.16 plots the remanence and coercivity versus the peak magnetic field and confirms that little is gained by operating the ferrite over $H_{peak} = 3000$ A/m.

Fig. 4.16 (a) Remanent flux density B_r and (b) coercive magnetic field H_c versus peak magnetic field at 20 °C.

The probing station, shown in Fig. 4.17, is equipped with a *Temptronic TP0315b ThermoChuck*, permitting temperature dependence measurements. A buried heating element is used to raise the chuck temperature whereas forced coolant is used to lower it. The *Temptronic* rack module contains the compressor, pumps, and the necessary electronics for remote temperature control.

Fig. 4.17 *Thermally controlled chuck and Temptronic controller rack.*

The surface area of the chuck is approximately 20 times that of the ferrite module and it is assumed that both are at the same temperature during the measurements. Fig. 4.18 illustrates the saturation B-field of the major $B(H)$ curve as a function of temperature, with nine data points having been taken between 10 and 120 °C.

Fig. 4.18 *Saturation flux density B_s versus temperature.*

Saturation decreases quickly as a function of temperature between 10 and 100 °C but appears to decrease less quickly above 100 °C. If the material was diamagnetic (non-magnetic), the peak B-field would be on the order of $\mu_0 H_{peak} \approx 5$ mT. The observed peak

B-field is always greater than this value so it appears that the ferrite material becomes *paramagnetic* (slightly magnetic) at a temperature in the vicinity of 110 °C. Fig. 4.19 displays the remanence and coercivity of the ferrite as a function of temperature.

Fig. 4.19 (a) Remanent flux density B_r and (b) coercive magnetic field H_c versus temperature.

The graphs indicate that B_r and H_c also decrease as a function of temperature. The temperature at which a ferrite makes the transition from ferrimagnetic to paramagnetic behaviour is known as the Curie temperature T_c [9]. For the ferrite under consideration, hysteresis ceases and the $B(H)$ curve becomes linear at temperatures above $T_c = 117$ °C. Other researchers, possibly using a similar material, have recently reported the Curie temperature of another ferrite tape to be 120 °C [11].

4.2.5. Magnetostatic characterisation summary

The findings of the preceding paragraphs are, on the whole, positive. It is encouraging to observe that the two manufactured solenoids are flawless, with not one open circuit in

the 1700+ layer transitions contained in each solenoid. Further, this was achieved without any costly special processing.

The magnetostatic properties of the ferrite are average. The saturation flux density of $B_s = 230$ mT places this ferrite midway in the range of possible ferrite saturation values, which generally range from 50 to 500 mT for commercially available ferrite. The measured remanent flux density of $B_r = 136$ mT is also consistent with commercially available ferrite having saturation flux densities in the range of 200–250 mT, such as magnesium and nickel ferrite *TT1-2000*, *TT1-2500*, and *TT2-102* from *Trans Tech*, for which the remanence values range from 139 to 149 mT. The measured coercive magnetic field $H_c = 688$ A/m is higher than the highest H_c of these same commercial ferrite materials, which is 352 A/m. This classifies the experimental material as being a *harder* ferrite, which means that higher current levels will be required to bias the ferrite. The material must still be regarded as a *soft* ferrite though, since its coercivity still lies below 1600 A/m [9]. A hard ferrite is classified as having a coercivity on the order of 80 000 A/m, which is a desirable property of permanent magnets [9].

The most surprising property of the experimental ferrite is its low Curie temperature of $T_c = 117$ °C, which is generally associated with ferrites having saturation flux density values below 100 mT. The same three commercial ferrites listed above have Curie temperatures in the range of 275–570 °C. The low Curie temperature obviously restricts the use of the material to applications that have constant and relatively low ambient temperatures. The low Curie temperature is also problematic because ohmic heating from the bias windings is likely to degrade the performance of the ferrite, especially since the ferrite is relatively hard.

4.3 Microwave characterisation of the Ferro ferrite

Two methods have been used to characterise the dielectric constant and loss tangent of the substrate. The first uses a novel microstrip ring resonator and the second uses laminated waveguide cavities.

4.3.1 Characterisation of the ferrite permittivity, permeability, and loss tangent using ring resonators

As with the previous LTCC material systems, microstrip ring resonators have been used to characterise the permittivity and loss of the ferrite substrate. A critical design aspect of the gap-coupled microstrip ring resonator, illustrated in Fig. 4.20(a), is the length of the coupling gap: it should be long enough to lightly load the ring but not so long as to cause the measured transmission resonance to be lost in the noise of the network analyser.

Fig. 4.20 Two-port co-linear fed ring resonators: (a) gap-coupled, (b) line-connected.

Over the course of this research, a number of fabricated ring resonators have indeed failed to generate detectable transmission levels due to the excessively light loading caused by the minimum gap length allowable in LTCC, which is on the order of 100 μm . This problem has also been observed by other researchers [12]. Given that only one

ferrite panel was to be manufactured, the risk of obtaining undetectable resonance levels with a gap-coupled ring was deemed unacceptable. As an alternative, a two-port ring resonator that is directly connected to its feed lines without coupling gaps was also included on the ferrite panel, as illustrated in Fig. 4.20(b), a configuration referred to as the *line-connected* ring resonator. As suspected, given the high dielectric constant of the ferrite, the gap-coupled ring failed to produce detectable resonance levels, meaning that the line-connected ring had to be used as a contingency. Given its new geometry, the analysis of the line-connected ring is slightly different from that of the gap-coupled ring, and the details of the former are presented in the following subsections.

4.3.1.1 Input impedance of the line-connected ring resonator

A transmission line model has been derived previously for the *unloaded* ring and it has been successfully applied to lightly loaded gap-coupled microstrip resonators [13]. For a colinear-fed two-port, the unloaded ring impedance reduces to:

$$Z_r = \frac{Z_0}{2} \coth(\gamma L / 2) \quad (4.9)$$

where Z_0 is the characteristic impedance, $\gamma = \alpha + j\beta$ is the complex propagation constant, and L is the mean circumference of the ring. An appropriate model of the two-port gap-coupled ring of Fig. 4.20(a) is shown in Fig. 4.21(a), where C_g and C_p are the capacitors associated with the coupling gap [14].

Fig. 4.21 *Circuit model of the two-port colinear-fed ring resonator (a) gap-coupled, and (b) line-connected (oversimplified).*

Given this, a plausible model for the line-connected ring is shown in Fig. 4.21(b), where the parasitic capacitance associated with the T-junction has been neglected altogether. Although the simple model of Fig. 4.21(b) may appear to be reasonable, the circuit does not adequately predict the behaviour of the ring, as will be shown presently. Indeed, it is false to assume that the same unloaded ring impedance given by (4.9) can be applied to the circuit of Fig. 4.20(b), meaning that a new formulation for the line-connected ring impedance must be sought.

As shown in Fig. 4.22, the impedance of the line-connected ring may be derived using an even/odd mode analysis [15].

Fig. 4.22 Excitation of the ring with (a) even and (b) odd modes, and their equivalent transmission line circuits.

The equivalent circuit for the even mode excitation is a parallel connection of two open-circuited $L/4$ transmission lines that presents an even input impedance Z_e and an even reflection coefficient Γ_e given by:

$$Z_e = \frac{Z_0}{2} \coth(\gamma L/4) \quad (4.10a)$$

$$\Gamma_e = \frac{Z_e - Z_0}{Z_e + Z_0} \quad (4.10b)$$

The equivalent circuit for the odd mode is identical except that the transmission lines are now short-circuited, presenting an odd input impedance Z_d and reflection coefficient Γ_d :

$$Z_d = \frac{Z_0}{2} \tanh(\gamma L/4) \quad (4.11a)$$

$$\Gamma_d = \frac{Z_d - Z_0}{Z_d + Z_0} \quad (4.11b)$$

The total input reflection coefficient and input impedance of the line-connected ring are given by:

$$\Gamma_{IN} = \frac{1}{2}\Gamma_e + \frac{1}{2}\Gamma_d \quad (4.12a)$$

$$Z_{IN} = Z_0 \frac{1 + \Gamma_{IN}}{1 - \Gamma_{IN}} \quad (4.12b)$$

whereupon substitution of (4.10)–(4.12a) into (4.12b) yields the input impedance expressed as a function of the even and odd mode impedance values:

$$Z_{IN} = Z_0 \frac{Z_0(Z_e + Z_d) + 2Z_e Z_d}{Z_0(Z_e + Z_d) + 2Z_0^2} = Z_0 \frac{1 + \coth(\gamma L/4) + \tanh(\gamma L/4)}{4 + \coth(\gamma L/4) + \tanh(\gamma L/4)} \quad (4.13)$$

In the absence of loss, it is straightforward to show that (4.13) equals Z_0 when $L = N\lambda_g$, presenting a perfect match and thereby causing a resonant dip in s_{11} at those frequencies, where N is the order of the resonance and λ_g is the guided wavelength. The presence of loss spoils the quality of the match. For odd-numbered resonance modes including the fundamental, the coth term in (4.13) may be neglected near resonance, leaving:

$$\begin{aligned} Z_{IN} &\approx Z_0 \frac{\cosh(\gamma L/4) + \sinh(\gamma L/4)}{4\cosh(\gamma L/4) + \sinh(\gamma L/4)} \\ &= Z_0 \frac{2e^{\gamma L/4}}{5e^{\gamma L/4} + 3e^{-\gamma L/4}} \\ &= Z_0 \frac{2e^{\alpha L/4}(\cos \beta L/4 + j \sin \beta L/4)}{5e^{\alpha L/4}(\cos \beta L/4 + j \sin \beta L/4) + 3e^{-\alpha L/4}(\cos \beta L/4 - j \sin \beta L/4)} \end{aligned} \quad (4.14)$$

Near the fundamental resonance frequency,

$$\beta L/4 = \pi/2 + \Delta\omega\pi/(2\omega_0) = \pi/2 + m \quad (4.15)$$

where ω_0 is the radian resonant frequency, $\Delta\omega = \omega - \omega_0$ is the single-sided bandwidth from resonance, and $m = \Delta\omega\pi/(2\omega_0)$. Given that the sine and cosine arguments are near $\pi/2$, the following substitutions may be made,

$$\begin{aligned}\cos(\pi/2 + m) &= -\sin m \approx -m \\ \sin(\pi/2 + m) &= \cos m \approx 1\end{aligned}\quad (4.16)$$

and if one further assumes that the attenuation is low enough to warrant only the first two terms of the exponential Taylor series,

$$e^{\alpha L/4} \approx 1 + \alpha L/4 \quad (4.17)$$

then substitution of (14) through (16) into (13) yields

$$Z_{IN} \approx Z_0 \frac{2(1 + \alpha L/4)(-m + j)}{5(1 + \alpha L/4)(-m + j) + 3(1 - \alpha L/4)(-m - j)} \approx Z_0 \frac{-2m + j2(1 + \alpha L/4)}{-8m + j2(1 + \alpha L)} \quad (4.18)$$

where the final approximation has been made by neglecting all $m \times \alpha L/4$ cross-terms, which are very small. Finally, (4.18) is inverted to obtain an input admittance, where again all $m \times \alpha L/4$ cross-terms are neglected, as are any second-order powers of m and αL ,

$$Y_{IN} = Y_0 \frac{(4 + 5\alpha L) + j12m}{4 + 2\alpha L} \quad (4.19)$$

and where $Y_0 = 1/Z_0$. Equation (4.19) is similar in form to the input admittance to the terminated, parallel-connected, parallel resonator illustrated in Fig. 4.23.

Fig. 4.23 Terminated, parallel-connected, parallel two-port resonator.

Note that loading the resonator with Y_0 is necessary for the analogy between it and the line-connected ring resonator to be valid. The input admittance to this parallel circuit near resonance is given by [16]:

$$Y_{IN} \approx G_{tot} + j2\Delta\omega C \quad (4.20)$$

where $G_{tot} = G + Y_0$. Equivalence is obtained between (4.19) and (4.20) if the following values are chosen:

$$G = \frac{3\alpha LY_0}{4 + 2\alpha L} \quad (4.21a)$$

$$C = \frac{3\pi Y_0}{\omega_0(4 + 2\alpha L)} \quad (4.21b)$$

which are the lumped-element equivalents of the line-connected ring resonator. As a verification, the unloaded quality factor of a parallel resonator is $Q_0 = \omega_0 C/G$, which upon substitution of (4.21) yields $Q_0 = \pi/\alpha L$. At resonance, $L = \lambda_g = 2\pi/\beta$, resulting in $Q_0 = \beta/2\alpha$, which is the expected quality factor of a TEM transmission line, confirming the derivation. From this we conclude that a quality factor measurement using s_{11} is sufficient to yield the loss of the line.

As a further verification, the reflection coefficient of the line-connected ring given by (4.12a) will be compared to that of the oversimplified model of Fig. 4.21(b), and to a simulation using a 2.5D EM simulator. The closed-form methods use dispersion-corrected data that have been obtained using microstrip synthesis software. The substrate is assumed to have a dielectric constant of $\epsilon_r = 2.2$, a loss tangent of $\tan\delta = 0.002$, a height $h = 0.787$ mm, and a conductor width of $W = 2.4$ mm with half-ounce copper cladding ($t = 17$ μm). The mean circumference of the ring is $L = 43.68$ mm, providing a fundamental resonance at 5 GHz. Fig. 4.24 compares the reflection coefficient magnitudes obtained using all three methods.

Fig. 4.24 Reflection coefficient magnitude of a line-connected two-port ring resonator. The dashed line uses the oversimplified model of Fig. 4.21(b), the solid line uses the even/odd mode model, and the circles are from a 2.5D simulation.

Note how the oversimplified model overestimates the return loss magnitude. The cause is its reliance on an *unloaded* ring impedance, which is negated by the *strong* coupling used by the line-connected configuration. Conversely, the even/odd mode model is in excellent agreement with the 2.5D simulation.

4.3.1.2 Resonance analysis of the line-connected ring

Fig. 4.25 plots the reflection magnitude of the even/odd mode model given by (4.12a) along with that obtained using the equivalent lumped element model using (4.21) in the vicinity of the first-order resonance.

Fig. 4.25 *Reflection magnitude of a line-connected two-port ring near resonance: the solid line uses the even/odd mode model whereas the dashed line uses the lumped element equivalent circuit.*

The lumped element approximation is in excellent agreement with the even/odd mode analysis near resonance, again confirming the derivation. For the terminated, parallel-connected, parallel two-port resonator of Fig. 4.23, Appendix C shows that the *unloaded* and *loaded* quality factors may be obtained by measuring the double-sided bandwidth at the following reflectance levels:

$$|s_{11\ U}| = \sqrt{\frac{2s_{11\ 0}^2}{s_{11\ 0}^2 + 1}} \quad (4.22a)$$

$$|s_{11\ L}| = \sqrt{\frac{s_{11\ 0}^2 + 1}{2}} \quad (4.22b)$$

where $s_{11\ U}$ is the unloaded 3-dB level, $s_{11\ L}$ is the loaded 3-dB level, and where $s_{11\ 0}$ is the magnitude of the reflection coefficient at resonance, all expressed in linear units. Oftentimes, when the loss is low, the unloaded bandwidth is very small and necessitates an inordinate number of measured data points or equivalently, a very small frequency step size. Because of this, the loaded bandwidth is often chosen to be measured instead.

Note that when the match is good, (4.22b) reduces to -3 dB, the so-called loaded 3-dB point for s_{11} . Returning to Fig. 4.25 however, note that the lumped-element model no longer applies at -3 dB: the line-coupled resonator is always sufficiently matched so that it *never* reflects half of the power that it absorbed at resonance. Hence measuring the loaded Q via (4.22b) is impossible, leaving the measurement of the unloaded bandwidth via (4.22a) as the only other choice. When the match is good ($s_{110} \approx 0$), (4.22a) reduces to $+3$ dB above the resonant level (the so-called unloaded 3-dB point). Referring again to Fig. 4.25, the slope of the reflectance is *very* high near resonance, making it difficult to locate $s_{110} + 3$ dB exactly; further, this bandwidth can be exceptionally small for low loss systems. The key to this “3-dB dilemma” is that the standard quality factor equations are derived *only* for the 3-dB, or “half-power” points. It would be beneficial if the bandwidth at *some arbitrary level other than half-power* could be related to the unloaded Q of the resonator. As shown in Appendix C, the unloaded Q of the resonator is related to the unloaded fractional double-sided bandwidth $\delta(p)$ measured at an arbitrary normalised power ratio $0 < p < 1$ via the following equation:

$$Q = \frac{1}{\delta(p)} \sqrt{\frac{1-p}{p}} \quad (4.23)$$

where p is the fraction of the power accepted by the unloaded resonator at frequencies coinciding with the bandwidth $\delta(p)$. Note how (4.23) reduces to the classical result of $Q = 1/\delta$ when $p = 0.5$, the half-power point. It may be further shown that the magnitude of the reflection coefficient for the circuit of Fig. 4.23 at the unloaded normalised power ratio p , in linear units is:

$$|s_{11\ U}| = \sqrt{\frac{s_{11\ 0}^2}{(1-p)s_{11\ 0}^2 + p}} \quad (4.24)$$

Hence measuring the bandwidth at a reflectance level of $s_{11\ U}$ and using the result in (4.23) leads directly to the unloaded quality factor of the resonator, and thereby also leads to the loss of the line. Note how (4.24) reduces to (4.22a) when $p = 0.5$, the 3-dB point. It is of interest to note the behaviour of (4.24) when $s_{11\ 0} \approx 0$ and p are small: when this is the case, $|s_{11\ U}| \approx s_{11\ 0} \times p^{-1/2}$, or in decibels, $s_{11\ U} \approx s_{11\ 0} + 10\log_{10}(1/p)$. Hence to increase the bandwidth measurement levels to 10 dB above the resonant level, one would choose $p = 0.1$; to increase the levels to 20 dB above resonance, choose $p = 0.01$, etc. For the data in Fig. 4.25, choosing $p = 0.001$ leads to a bandwidth measurement near $|s_{11}| = -20$ dB, which is sufficiently wide to alleviate the problems associated with small bandwidths. Further, the lumped-element model still applies near this level. The only caveat to using (4.23) and (4.24) occurs when the resonator match is very good (when the loss is very low), which causes the exact value of $s_{11\ 0}$ to be very small and potentially lost again in the noise of the network analyser.

4.3.1.3 Microstrip line-connected ring design

A practical ring design cycle will now be considered. Unfortunately, the permittivity of the substrate is unknown, although the relative permittivity of ferrite is commonly between 12 and 16. Given that the fired tape layer thickness is approximately 100 μm , a practical height h of the microstrip substrate must be chosen. The substrate height condition to avoid substrate modes is given by $h < 0.1\lambda_{g(\text{TEM})}$. The smallest height given by this equation occurs if we assume a worst case of $\epsilon_r = 16$, which at 40 GHz yields

$h < 263 \mu\text{m}$, meaning that up to two tape layers are permitted. The width of the microstrip W is too narrow when h is only one layer, being approximately $70 \mu\text{m}$ for a 50Ω line. Hence two layers must be used, making the substrate height $h = 2 \times 108 - 8 = 208 \mu\text{m}$, where it is assumed that the conductors will leach into the tape by $8 \mu\text{m}$. A microstrip width of $W = 150 \mu\text{m}$ yields impedance values that are close to 50Ω if the relative permittivity is in the range of 12–16.

Three rings have been designed, each having a fundamental frequency that is slightly offset from the others for better frequency coverage at the harmonics. The ring design assumes a dielectric constant of 12 and fundamentals of 4.5, 4.7, and 5.0 GHz. Using the gap-coupled ring circuit model, appropriate ring radii are 3.80, 3.65, and 3.40 mm. The worst case L/W ratio is 178, meaning that $L \gg W$ is satisfied and the curvature of the ring is inconsequential [17]. The nominal coupling gap length is set to $S = 120 \mu\text{m}$, which is very close to the $100 \mu\text{m}$ minimum spacing allowable. Using the gap-coupled circuit model, the estimated transmission peaks are in the range of -55 to -20 dB from 5 to 40 GHz if the attenuation of the line is assumed to be 0.5 dB/cm, which should be acceptable. As a precaution, the 4.7 GHz ring resonator was converted to a line-connected ring.

4.3.1.4 Microstrip line-connected ring measurements

The average measured gap length of the fabricated rings is $S = 135 \mu\text{m}$, $15 \mu\text{m}$ above nominal. Not surprisingly, this small difference is enough to send the transmission peaks of the two gap-coupled resonators into the noise of the network analyser, meaning that two of the three rings fail to provide any useful data. The reflection response of the

contingency line-connected 4.7 GHz ring is shown in Fig. 4.26, where all eight resonance troughs have been numbered.

Fig. 4.26 *Measured reflection coefficient magnitude of the line-connected ring resonator.*

A few of the troughs appear to be corrupt, namely $N = 4, 5$, and possibly 8. The fourth resonance minimum is ill-defined and yields a wide bandwidth, the fifth is exceptionally narrow, and the eighth is somewhat noisy. Ideally, several identical resonators would be measured to obtain averaged values. Nonetheless, a resonance analysis has been performed at the $p = 0.25$ (quarter-power) points, yielding bandwidth measurements at the locations marked by X in Fig. 4.26, where an X also indicates the resonance frequency. The resulting attenuation constants are plotted in Fig. 4.27.

Fig. 4.27 *Total attenuation of the line-connected ring resonator obtained by measuring the quarter-power unloaded bandwidth.*

Ignoring the fourth and fifth resonant frequencies, the total attenuation appears to range between 0.3 and 0.8 dB/cm from 5 to 40 GHz. The maximum value of 0.8 dB/cm (9.2 Np/m) is approximately double the typical maximum attenuation of other LTCC microstrip lines at 40 GHz, which is on the order of 0.4 dB/cm (4.6 Np/m).

Lumped-element model “Libra-style” simulations for a similar material system show that the microstrip T-junctions reduce the resonant frequency by 0.6% over all eight harmonics compared to simulations without the MTEE circuit element, hence capacitive loading can be neglected in this early stage of material characterisation. Therefore (4.25) may be used directly to extract the effective permittivity:

$$\epsilon_{eff} = \left(\frac{cN}{f_0 L} \right)^2 \quad (4.25)$$

The measured mean radius of the ring resonator is $r = 3645 \pm 5 \mu\text{m}$, which is close to the nominal $3650 \mu\text{m}$. The effective permittivity is plotted in Fig. 4.28 for the eight resonant frequencies.

Fig. 4.28 *Effective permittivity versus frequency.*

The curve shows a strong variation below 20 GHz, the cause of which will be discussed presently.

The final step in the characterisation process is to convert the attenuation and effective permittivity into loss tangent and dielectric constant values, which is done using Agilent's *LineCalc* microstrip synthesis program. The measured microstrip width is $W = 140 \mu\text{m}$, $10 \mu\text{m}$ less than the nominal. The previously found values of conductivity and roughness are used for the Ferro silver paste, which are $5.0(10^7) \text{ S/m}$ and 700 nm , respectively. The results are plotted in Fig. 4.29.

Fig. 4.29 (a) Combined relative permittivity and relative permeability, and (b) loss tangent of the experimental ferrite.

Note that the relative permittivity ϵ_r and relative initial permeability μ_{ri} are inextricable using this analysis; as such, Fig. 4.29(a) plots the product of $\epsilon_r \mu_{ri}$. (The term relative *initial* permeability is used here because the ferrite sample is unmagnetised.) Similarly, it is the *sum* of the magnetic and electric loss tangents that is plotted in Fig. 4.29(b). Observe how $\epsilon_r \mu_{ri}$ increases over the frequency range of 4–15 GHz up to a nearly constant value of 11; simultaneously, the loss tangent decreases by nearly an order of magnitude from 0.02 to values near 0.003, notwithstanding the anomalous fourth order resonance.

This behaviour is in fact a reasonable one. Domain wall rotation within Weiss domains is the primary contributor to initial permeability, and a resonance frequency is associated with this rotation [9]. The resonant frequency depends on the shape of the Weiss domain; in a polycrystalline ferrite such as the one under consideration here, a

multitude of crystallite sizes exists, each having its own resonant frequency. Hence the losses associated with this type of resonance are broadly distributed from zero frequency upwards. This is the source of so-called “low-field” losses in an unsaturated ferrite. It can be shown that the maximum possible domain rotation resonance frequency, the frequency at which these losses should tail-off, is given by [9]:

$$f_{\max} = 2\pi\gamma_r M_s \quad (4.26)$$

where γ_r is the gyromagnetic ratio and M_s is the saturation magnetisation (other work has shown that $f_{\max}$ should be increased by a factor of 1.3 [18]). For the ferrite under consideration here, the measured saturation magnetisation is 2300 G, yielding $f_{\max} = 8.4$ GHz [18]. Not surprisingly then, losses are high at the ring’s first resonance and decay as the frequency is increased. Hence the magnetic loss tangent is suspected as being the primary source of the frequency dependence in Fig. 4.29. The relative permeability of ferrite can also known to dip below unity during domain wall resonance, which explains the behaviour of the $\epsilon_r\mu_{ri}$ graph in Fig. 4.29(a) [9].

4.3.2 Characterisation of the ferrite permittivity, permeability, and loss tangent using LWG cavities

Although the microwave data that has been collected using the microstrip ring should, in theory, be applicable to laminated waveguide, two LWG cavities have also been fabricated and tested. The data that these cavities provide will be directly applicable to the phase shifters, since the latter have been implemented using the same LWG medium. A top view of the layout of one of the cavities is shown in Fig. 4.30, where the upper broad wall has been removed to explicitly show the inside structure.

Fig. 4.30 *Top view of a laminated waveguide cavity layout in ferrite. The top broad wall has been removed to show the inside structure and the ground-signal-ground pads have been made transparent to show their underlying vias.*

The ground pads have vias that connect to the (invisible) upper broad wall of the cavity, whereas the centre conductor passes through an aperture in the upper broad wall and descends one layer into the cavity. The vias forming the walls of the cavity are connected on each layer by laminations. Both cavities are composed of four layers of ferrite tape, having a total thickness of $b = 420 \mu\text{m}$. The via diameter is $\varnothing = 150 \mu\text{m}$ and the via pitch is $p = 500 \mu\text{m}$. The first cavity has a via centre-to-centre width of $a_1 = 2350 \mu\text{m}$ and a via centre-to-centre length $d_1 = 5750 \mu\text{m}$, whereas the second cavity has $a_1 = 2000 \mu\text{m}$ and $d_2 = 5000 \mu\text{m}$. Fig. 4.31 plots the transmission s-parameter through both cavities as a function of frequency.

Fig. 4.31 Transmission magnitude of (a) the ferrite cavity having $a = 2350 \mu\text{m}$ and $d = 5750 \mu\text{m}$, and (b) the ferrite cavity having $a = 2000 \mu\text{m}$ and $d = 5000 \mu\text{m}$.

Four resonant frequencies are identifiable for the first cavity whereas three are present for the second cavity, each indicated by an arrow. Ignoring the parasitic reactance of the probes, the combined dielectric/magnetic constant of the material may be extracted via:

$$\mu_r \epsilon_r = \frac{c^2}{4\pi^2 f_{10\delta}^2} \left[\left(\frac{\pi}{a} \right)^2 + \left(\frac{\delta\pi}{d} \right)^2 \right] \quad (4.27)$$

where δ is the order of the resonance and $f_{10\delta}$ is the observed $\text{TE}_{10\delta}$ -mode resonant frequency. The results are plotted in Fig. 4.32 using squares, whereas the circles are the repeated data obtained from the ring resonators in the previous section.

Fig. 4.32 Combined relative permittivity and relative permeability, obtained using laminated waveguide cavities (squares) and ring resonators (circles).

From 22 to 36 GHz, the combined dielectric/magnetic constant ranges from 10.1–10.7 instead of from 10.5–11.4, the latter obtained using the microstrip ring. There are two probable causes for the discrepancy: firstly, the parasitic reactance of both types of resonators has been ignored, and second, it is difficult to estimate an exact value of d due to LTCC shrinkage. Further, referring to Fig. 4.30, the actual value of d can be measured either from the inside edges of the laminations, or from the via-centre to via-centre, a difference of 200 μm . Using the smallest estimate of d causes the dielectric constant estimates to range from 10.1 to 10.9.

Fig. 4.31 indicates that the LWG cavity is a transmission two-port resonator, contrarily to the line-connected microstrip ring resonator, which is an absorption resonator. The equations relating the quality factor to the bandwidth of transmission resonators are also found in Appendix C and have been used to evaluate the unloaded Q of the cavities, which are illustrated in Fig. 4.33(a) as triangles for the first cavity and as circles for the second cavity. The X symbols in Fig. 4.33(a) are simulated RWG values

that have been obtained using silver as the conductor and the loss tangent values shown in Fig. 4.33(b), where triangles and circles again indicate either the first or second resonator, respectively.

Fig. 4.33 (a) Measured unloaded quality factors of the first (triangles) and second (circles) cavities, compared to simulations using RWG formulae (X); (b) loss tangents required to simulate the unloaded Q of the first (triangles) and second (circles) cavities, superimposed on the results from the line-connected ring resonator (squares).

The loss tangent values shown in Fig. 4.33(b) (circles and triangles) indicate that the loss tangent of the substrate is consistently near 0.004 over the 20–40 GHz band, and these values are slightly higher than those obtained using the line-connected ring (squares). Notwithstanding this apparent discrepancy, the primary observation is that the ferrite performs well over the 20–40 GHz band, both as a microstrip substrate and as a LWG dielectric.

4.4 Summary and conclusions of the Ferro ferrite characterisation

Perhaps the most important observation of this chapter is that it is possible to fabricate high quality millimetre-wave ferrite LTCC modules using standard processing

techniques, thereby ensuring that the cost of any future volume production will remain competitive. The material properties of the Ferro ferrite may be generally regarded as being *good*. The low frequency measurements indicate that the ferrite is clearly magnetic, having a measured saturation flux density of $B_s = 230$ mT. The remanent flux density of the ferrite is 136 mT, indicating that the ferrite does not have a good squareness ($B_r/B_s = 0.6$) and that it is unlikely that the ferrite can be used at remanence. Further, the coercivity of the ferrite is quite high, $H_c = 688$ A/m, meaning that high current levels are required for saturation, which is problematic since the Curie temperature is relatively low, $T_c = 117$ °C. It is therefore likely that the heat generated from the high current levels will adversely impact the performance of the ferrite. There is clearly room for improvement in the chemical formulation of this material.

The line-connected microstrip ring resonator has solved the problem of undetectable transmission levels, to which gap-coupled ring resonators are prone. Analyses of the line-connected resonator measurements indicate that the total attenuation of the microstrip line is in the range of 0.3–0.8 dB/cm over the 5–40 GHz band. The loss tangent of the substrate drops from 0.02 to approximately 0.003, and the product of the relative permittivity and relative initial permeability rises from 5.6 to a near constant value of approximately 11 over this same band. The strong frequency dependence observed below 15 GHz is most likely due to domain rotation, and as expected, these effects are altogether absent above 20 GHz. Assuming that $\mu_{ri} \rightarrow 1$ above 20 GHz, the relative permittivity of the ferrite is 11.0 ± 0.5 , which is slightly below the estimated value of 12. Above the fifth order ring resonance ($f > 25$ GHz) the total loss tangent is in the range of 2 to $4(10^{-3})$.

Two laminated waveguide cavities have also been used to characterise the ferrite in the 25–40 GHz range. The dielectric/magnetic constant estimate is 10.1–10.7, which is slightly lower than that obtained using the ring resonator, which was 10.5–11.4. The discrepancy is either due to non-negligible reactive parasitic loading, or to difficulties in assessing the exact effect length and width of the cavity resonators. The loss tangent obtained from the Q measurement of the LWG cavities is near 0.004 over the band, which is slightly higher than that obtained using the ring resonator, which varies between 0.002 and 0.004.

In general, the ferrite substrate performs well in the 20–40 GHz range: the relative permittivity is nearly constant and the loss tangent is fairly low. Note, however, that the loss estimates do not yet include the additive damping factor loss associated with the ferrimagnetic linewidth, which may also be present when the ferrite is biased (magnetised).

4.5 Chapter Four references

- [1] J. Y. Park, L. K. Lagorce, and M. G. Allen, "Ferrite-based integrated planar inductors and transformers fabricated at low temperature", *IEEE Trans. Magn.*, vol. 33, pp. 3322–3324, Sept. 1997.
- [2] Q. Yu, T. W. Holmes, and K. Naishadham, "RF equivalent circuit modeling of ferrite-core inductors and characterization of core materials", *IEEE Trans. Electromagn. Compat.*, vol. 44, pp. 258–262, Feb. 2002.
- [3] G. P. Rodrigue, "A generation of microwave ferrite devices", *Proc. IEEE*, vol. 76, pp. 121–137, Feb. 1988.
- [4] M. T. Quirke, J. J. Barrett, and M. Hayes, "Planar magnetic component technology—a review", *IEEE Trans. Comp. Hybrids, Manufact. Technol.*, vol. 15, pp. 884–892, Oct. 1992.
- [5] H. L. Glass, "Ferrite films for microwave and millimeter-wave devices", *Proc. IEEE*, vol. 76, pp. 151–158, Feb. 1988.

- [6] M. Abe, T. Itoh, Y. Tamaura, Y. Gotoh, and M. Gomi, "Ferrite plating on GaAs for microwave monolithic integrated circuit", *IEEE Trans. Magn.*, vol. 23, pp. 3736–3738, Sept. 1987. R. F. Soohoo, *Theory and Application of Ferrites*, Englewood Cliffs, NJ: Prentice-Hall, 1960, pp. 83–88.
- [7] P. Quéffélec, M. Le Floch, and P. Gelin, "New method for determining the permeability tensor of magnetized ferrites in a wide frequency range", *IEEE Trans. Microwave Theory Tech.*, vol. 48, pp. 1344–1351, Aug. 2000.
- [8] K. N. Kocharyan, M. N. Afsar, and I. I. Tkachov, "Millimeter-wave magneto-optics: new method for characterization of ferrites in the millimeter-wave range", *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 2636–2643, Dec. 1999.
- [9] R. F. Soohoo, *Theory and Application of Ferrites*, Englewood Cliffs, NJ: Prentice-Hall, 1960, pp. 83–88.
- [10] U. S. Inan & A. S. Inan, *Engineering Electromagnetics*, Menlo Park, CA: Addison-Wesley, 1999, p. 514.
- [11] G. Slama, "Low-temp co-fired magnetic tape yields high benefits", *Power Electronics Technology*, pp. 30–34, Jan. 2003.
- [12] L. Devlin, G. Pearson, J. Pitlock, and B. Hunt. 2001, Sept. RF & Microwave Components Development in LTCC. Presented at IMAPS Nordic Conf., Oslo, Norway. [Online]. Available: http://www.cmac.com/mt/LTCC_IMN.pdf
- [13] C.-C. Yu and K. Chang, "Transmission-line analysis of a capacitively coupled microstrip-ring resonator", *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 2018–2024, Nov. 1997.
- [14] J. R. Bray & L. Roy, "Microwave characterization of a microstrip line using a two-port ring resonator with an improved lumped element model", *IEEE Trans. Microwave Theory Tech.*, vol. 51, pp. 1540–1547, May 2003.
- [15] J. Reed & G. J. Wheeler, "A method of analysis of symmetrical four-port networks", *IRE Trans. Microwave Theory Tech.*, vol. 4, pp. 246–252, Oct. 1956.
- [16] D. M. Pozar, *Microwave Engineering*, 1st ed., Reading, MA: Addison-Wesley, 1990, pp. 333–335.
- [17] T. C. Edwards & M. B. Steer, *Foundations of Interconnect and Microstrip Design*, 3rd ed., Chichester, UK: John Wiley & Sons, 2000, section 9.7.
- [18] M. Igarashi & Y. Naito, "Microwave permeability tensor of partially magnetized ferrites", in *IEEE MTT-S Int. Symp. Dig.*, 1976, vol. 1, pp. 269–271.

Chapter Five — LTCC rectangular waveguide phase shifters using via posts made of ferrite paste

Chapter Three has shown how low-loss millimetre-wave laminated waveguides may be readily constructed in LTCC. The simple and practical implementation of these compact, integrated, three-dimensional transmission lines begs the question as to whether or not waveguide *devices* may also be integrated in LTCC. Suppliers of LTCC materials claim that a wide variety of materials are available for the integration of passive devices, such as resistors, capacitors, and inductors. High quality inductors use ferrite materials, and if these can be secured, then it is likely that they could also be used to construct laminated waveguide phase shifters. Rectangular waveguide ferrite phase shifters have fallen out of favour over the past two decades due to their apparent incompatibility with planar circuitry. The continued popularity of LTCC as a packaging technology, combined with the evolution of ferrite materials suitable for use with LTCC, have lead to the feasibility of constructing a phase shifting device made entirely out of packaging material, bypassing the need for expensive monolithic chips.

At the present time, ferrite thick film paste that is compatible with LTCC is regarded as an experimental material and is not widely available. Of the three leading suppliers of LTCC materials (Dupont, Ferro, and Heraeus), only Heraeus offers ferrite thick film pastes but they are not formulated specifically for use with the Heraeus low-loss CT2000 LTCC system. Instead, these pastes are intended for the fabrication of multilayer inductors using a process called *Wet Stack*, which is similar to LTCC processing. A fabrication run of Heraeus CT2000 was scheduled at VTT Electronics in September of 2001, presenting an opportunity to test the compatibility of the Wet Stack ferrite paste

with the CT2000 LTCC material system. Two types of ferrite paste were donated by Heraeus for the purpose, IP9150 and IP9220, where the final three digits represent the relative initial permeability of the ferrite at 1 MHz. The specification data sheet for IP9150 is listed in Appendix A.

This chapter is subdivided into nine sections. The first introduces the Wet Stack ferrite pastes and considers their compatibility with the CT2000 LTCC system. Section 5.2 determines how ferrite paste can be used with LTCC laminated waveguide (LWG) to fabricate millimetre-wave phase shifters. Designs using transverse-bias ferrite-loaded waveguide configurations are shown to be the most conducive to an LTCC integration, and a LWG loaded with rows of ferrite vias is suggested as a suitable prototype. An analytical method for calculating the phase shift through the device is required, and this is presented in Section 5.3 for a LWG loaded with two rows of ferrite vias. In an attempt to improve the phase shift, Section 5.4 presents the field theory of a waveguide that is loaded with four rows of ferrite vias. The primary limitation of using ferrite vias is that they fill but a small portion of the waveguide's total volume. Section 5.5 discusses how the ferrite filling factor may be maximised while simultaneously respecting the LTCC design rules. The amount of phase shift that can be achieved depends on the location of the ferrite vias along the guide's cross-section. In Section 5.6, the ideal location is compared with a figure of merit that is based on the axial ratio of the circular-polarised magnetic fields inside the ferrite vias, and a good correlation between it and the maximum phase shift is obtained. The magnetostatic field within the ferrite vias controls the phase shift, and Section 5.7 presents the practical implementation of the bias fields using closed ferrite toroids inside the LTCC module. The bias arrangement completes the

design, therefore Section 5.8 presents the fabrication and measured results of a number of prototype two-post ferrite-loaded laminated waveguides. The findings of this chapter are then summarised in Section 5.9.

5.1 Wet Stack multilayer ferrite paste and its compatibility with Heraeus CT2000 LTCC

Heraeus Wet Stack thick film ferrite paste is specifically formulated for the fabrication of multilayer inductors operating between 1 and 25 MHz [1]–[3]. Although the printing of Wet Stack inductors is very similar to that used in LTCC fabrication, a notable exception is that *no dielectric tape layers* are present in the former process. The first step consists of the sequential printing and drying of thick film ferrite paste layers onto a temporary rigid carrier. Next, conductor coils are printed and dried. Another layer of ferrite is then printed, this time with apertures for vias, which are subsequently filled before printing the next conductor coil layer. These steps are repeated until all of the coils are defined. More layers of blank ferrite are then printed to complete the stack, which is illustrated in Fig. 5.1.

Fig. 5.1 Wet Stack thick film ferrite layers and solenoidal windings for the fabrication of inductors.

After the final ferrite layer has been dried, the ferrite stack is removed from the carrier, fired, and diced. The view in Fig. 5.1 is akin to a side view of several vertically-oriented

solenoids, and viewing the stack from the top would reveal that the windings form circular loops on every level. The blank ferrite layers at the top and bottom are used for magnetic flux return paths.

The printing of conductors on LTCC tape is accomplished using the same thick film technique, except that the conductors are printed on multiple, separate dielectric tape sheets instead of being printed sequentially on dried ferrite paste. Table 5.1 lists some of the material properties of the different materials.

Table 5.1 *Selected material properties of CT2000 dielectric tape, CT2000 pastes, and Wet Stack IP9xxx ferrites.*

	CT2000 Dielectric Tape	CT2000 Silver Conductor TC2304HQ	CT2000 Silver Via Fill TC7301	IP9xxx Wet Stack Ferrites
Printing Screen	---	290–400 mesh	---	200–325 mesh
Thinner	---	RV 507	HVS 507	RV 100
Emulsion	---	N/A	---	0.5 mils
Drying	---	< 80°C, 10– 20 min	< 80°C, 10– 30 min	100°C, 10 min
Film Thickness	---	10 μm (fired)	---	18–20 μm (dried)
Planar Shrinkage	11.5%*	---	---	12.7–15.3%** (dried to fired)
Lamination	4500 PSI @ 70°C, 5–10 min	---	---	None

* depends on the lamination conditions (temperature and pressure), ** depends on the type of ferrite paste, N/A = data unavailable

Given that the printing method is identical for thick film conductors and thick film ferrite, the selection of an appropriate screen does not pose a problem. If the pastes do not flow adequately through the printing screen, specially formulated thinners, which are also supplied by Heraeus, can be added to lower their viscosity. The term *emulsion* refers to the polymer that is applied to the printing screen to define the printing pattern. If the

emulsion in the mesh is too thin, the printed lines will be poorly defined. A common LTCC thick film printing practice is to increase the thickness of the emulsion beyond that of the screen by approximately 0.5 mils, the same value that is listed for the ferrite material. Vias are filled using a stainless steel stencil instead of a mesh, so emulsion is irrelevant to via-fill paste. The drying step is used to remove most of the solvent (organic flowing agent) from the paste. The drying temperature and airflow must be tightly controlled to avoid surface cracking of the ferrite. The lower drying temperatures used by CT2000 should not pose a serious problem as long as the airflow in the oven is kept at, or below, 8 cubic feet per minute [3]. The *fired* thickness of the LTCC conductor paste is approximately 10 μm , while that of the ferrite is unknown—only the *dried* ferrite thickness of 19 μm is listed. The shrinkage of thickness during the firing of another Heraeus ferrite paste, CL90-7057 (see Appendix A), is approximately 30%, hence the expected fired thickness of the fired Wet Stack ferrite should be approximately 13 μm , not very different from that of the LTCC conductors. The *lateral* shrinkage of the material is another important property. If the materials are mismatched, cracking and even delamination of the LTCC tape may occur. The ferrite pastes shrink by 15.3% during firing. The shrinkage of the CT2000 dielectric tape is lower, nominally 11.5%. Although there is a small mismatch of 3.8%, the following points must be noted: firstly, VTT uses a lamination pressure of 3500 psi as opposed to the recommended pressure of 4500 psi. The Heraeus specifications (Appendix A) state that using a *higher* pressure decreases the shrinkage. Secondly, lamination is not required in the Wet Stack system. *If* the ferrite paste behaviour is similar to that of the tape, the lamination step *might* reduce

its shrinkage. If either or both of these points are true, the mismatch will be lower than 3.8%.

The most significant discrepancy between the CT2000 and the Wet Stack system is the burnout and firing temperature profile, illustrated in Fig. 5.2.

Fig. 5.2 *Firing profiles of Heraeus CT2000 LTCC, of Wet Stack ferrite pastes IP9xxx, and of ferrite paste CL90-7057.*

The peak temperature of both systems is nearly identical, ranging between 850 and 900 °C. This is not coincidental because both systems use silver conductors. Although the melting point of silver is 960 °C, it begins to bead and eventually volatilises if the firing temperature is raised above 900 °C. The two LTCC firing profiles shown in Fig. 5.2 represent the shortest and longest recommended firing profiles of CT2000. The IP9xxx curve has two burnout soaks: one is a 30 minute solvent burnout at 220 °C and the second is a 30 minute binder burnout at 350 °C. The IP9xxx soak at peak temperature lasts for two hours whereas the LTCC soak is only 20–30 minutes. The implication of the short soak time is potentially severe: if the solid state reaction to the ferrite phase is incomplete, the magnetisation properties of the ferrite will be degraded. Heraeus states

that “the time at peak can be varied between one and four hours for minor adjustment of the inductance value” [3]. Also included in Fig. 5.2 is the firing profile of another Heraeus ferrite thick film paste, CL90-7057, which is optimised for printing on alumina substrates. Its relative permeability is 150 at 1 MHz and it is compatible with other Heraeus thick film pastes. According to Fig. 5.2, its firing profile is similar to that used by CT2000. Although a sample of this paste was obtained, it was unfortunately not received in time to be incorporated in the Heraeus fabrication run.

The amount of phase shift that can be achieved in a device is proportional to the saturation magnetisation of the ferrite. However, only the relative permeability of IP9xxx and the quality factor of typical multilayer inductors are provided by Heraeus, which are inadequate metrics for phase shifter applications. Fortunately, a thesis pertaining to the characterisation of multilayer inductors has recently been published in which the major hysteresis loops of IP9xxx ferrite have been measured [4], listing IP9150 and IP9220 as having saturation flux density values of 220 and 250 mT, respectively, which are typical of good ferrite materials.

Although shrinkage mismatch is a potential problem, the only way to determine its impact is to manufacture prototype circuits using the ferrite material. In the next section, potential waveguide phase shifter configurations using ferrite paste are considered.

5.2 Integration of ferrite paste and laminated waveguide for the design of phase shifters

Laminated waveguides must be oriented such that their narrow walls are perpendicular to the LTCC tape layers because the RWG TE_{10} -mode surface currents are co-linear with the narrow-wall vias. The surface current is diffuse on the broad walls,

which must therefore be made from contiguous printed metal. Ferrite paste offers a high degree of design flexibility because it may either be used to print films at the LTCC tape interfaces or used to fill vias in the transverse direction, as shown in Fig. 5.3.

Fig. 5.3 *E-plane cross-section of a laminated waveguide showing ferrite paste used as thick film or as via fill material.*

A reciprocal RWG phase shifter may be constructed by placing longitudinal, longitudinally-biased ferrite inside the guide, a classical example of which is the Reggia-Spencer phase shifter [5]. A Faraday rotation phase shifter also uses longitudinal ferrite, although it requires the guide to be fully-filled with ferrite, with the addition of dielectric-filled sections and absorbent vanes [5]. The added complexity of the Faraday rotation phase shifter makes it less conducive to an LTCC realisation.

Contiguous, longitudinal ferrite can be realised by printing strips of ferrite film down the length of the guide. A Reggia-Spencer phase shifter uses a square ferrite bar (or a cylindrical rod), of sides (or diameter) d , that is centred within an air-filled RWG. For nominal operation, the cross-sectional dimension d of the ferrite must lie within the range [5]:

$$\frac{\lambda_0}{1.71\sqrt{\epsilon_{rf}}} \leq d \leq \frac{\lambda_0}{1.31\sqrt{\epsilon_{rf}}} \quad (5.1)$$

where ϵ_{rf} is the relative permittivity of the ferrite. If d is too small, the phase shift is negligible whereas the insertion loss increases rapidly if d is too large. The finite

thickness of the ferrite film obviously limits the height of the ferrite within the guide. The minimum d in (5.1) may be rearranged to explicitly show the minimum frequency of operation:

$$f \geq \frac{c}{1.71d\sqrt{\epsilon_{rf}}} \quad (5.2)$$

where c is the speed of light. If a typical film thickness of 10 μm is chosen with a dielectric constant of 12, the minimum operating frequency lies within the terahertz range, well beyond the frequencies of interest. Although the present work uses dielectric-filled waveguides instead of air-filled waveguides, HFSS simulations have indicated that widening the ferrite strip does not provide any appreciable phase shift at millimetre-wave frequencies. Another configuration would be to place a series of ferrite vias down the centre of the guide, although providing a longitudinal bias across such a broken magnetic circuit would be extremely difficult. Given these incompatibilities, the Reggia-Spencer design has been abandoned.

Nonreciprocal RWG phase shifters may be constructed by placing transverse-biased, longitudinal ferrite slabs in the guide [5]. A plethora of slab-loaded waveguide designs exists in the open literature, a sample of which is shown in Fig. 5.4 for air-filled guide.

Fig. 5.4 Variants of the slab-loaded nonreciprocal RWG phase shifter, where arrows indicate the direction of magnetic bias: (a) two thin slabs flush with the broad walls, (b) one full-height slab, (c) two antisymmetrically-biased full-height slabs, (d) a twin-toroid with dielectric loading (hashed area).

The structure of Fig. 5.4(a) is effective in high-power applications where the removal of heat from the ferrite is paramount. Increased phase shift may be obtained by using the asymmetrical slab in Fig. 5.4(b), and even more is obtained by choosing the antisymmetrical configuration of Fig. 5.4(c). Closed magnetic circuits are more easily biased, so the configuration of Fig. 5.4(c) may be approximated by using toroids inside the guide, as shown in Fig. 5.4(d). Although the bias field in the outer segments of the toroids (near the narrow walls of the guide) are in the wrong direction, they are nonetheless tolerated because the energy is mostly concentrated in the centre of the guide. Dielectric loading may be used to improve the phase shift of the device.

The configurations of Fig. 5.4 are amenable to an implementation in LTCC if ferrite-filled vias are used to approximate the slabs. Normally, the slabs are made from contiguous ferrite running longitudinally down the length of the guide; the LTCC implementation is expected to negatively impact the phase shift since it reduces the interaction volume between the ferrite and the guided wave. Fig. 5.5 illustrates the

proposed implementation, where the complete magnetostatic circuit is only shown on one side of the LWG.

Fig. 5.5 *Nonreciprocal, transverse-biased ferrite phase shifter implemented using laminated waveguide, ferrite vias, and ferrite thick film.*

Instead of housing the toroids inside the guide, apertures in the broad walls of the LWG lead the ferrite vias to ferrite thick film flux return arms, which then direct the flux into the bias region, consisting of more ferrite vias surrounded by bias windings. A theoretical framework is required to predict the achievable phase shift using ferrite posts instead of slabs, and this is presented in the next section for the two-post case, and then in the subsequent section for the four-post case.

5.3 Two-post ferrite-loaded RWG field derivation and phase shift calculation

The fields of the fundamental TE mode in the ferrite-loaded RWG illustrated in Fig. 5.6(a) may be formulated by using mode matching at the ferrite/dielectric interfaces [6].

Fig. 5.6 *Transverse-biased ferrite RWG phase shifters: (a) dual slab, (b) cylindrical vias, (c) E-plane cross-section through the vias.*

The derivation yields a transcendental equation that must be solved numerically for both the forward β_+ and reverse β_- propagation constants. In a dual-slab ferrite phase shifter, the maximum achievable phase shift is the difference of the calculated β roots, multiplied by the length L of the interaction region:

$$\Delta\phi = (\beta_+ - \beta_-) \cdot L \quad (5.3)$$

In Fig. 5.6(c), a transverse cut through a pair of vias is shown. The distance between the via and the wall c , the via thickness t , and the space between the vias s all depend on the longitudinal position of the cut. If it passes through the centre of the via, c and s assume their minimum values whereas t is at its maximum, which is the diameter of the via, $\varnothing$. The parameters c , t , and s appear explicitly in the transcendental equation for the dual-ferrite slab phase shifter; therefore, it is possible to determine the instantaneous β roots as a function of the location z where the transversal cross-section is taken. After sampling the longitudinal length of the ferrite into N samples, the β roots are determined for each discrete location, and the total phase shift through one pair of vias is then calculated by using the discrete summation:

$$\Delta\phi = \sum_{n=0}^N \Delta\phi(n) = \sum_{n=0}^N \Delta z(n) [\beta_+(n) - \beta_-(n)] \quad (5.4)$$

where Δz , the distance between the samples in z , need not be constant but can be made smaller where the variations of the instantaneous phase shift $\Delta\phi(n)$ are the largest. Given the circular symmetry, the summation is necessary over only half of the via, and doubling it yields the total phase shift through the vias. An example calculation is done for the material parameters shown in Table 5.2.

Table 5.2 *Material parameters for a dielectric-filled, two-post ferrite-loaded RWG.*

frequency f [GHz]	RWG width a [μm]	RWG dielectric constant	ferrite dielectric constant	via diameter $\varnothing$ [μm]	bias H_0 [A/m]	flux density B [mT]	via centre [μm]
35	2300	$\epsilon_r = 9.1$	$\epsilon_{rf} = 12$	150	557	250	$x = 550$

The guide's dielectric constant of 9.1 is typical of the Heraeus CT2000 system, whereas the bias field and flux density are representative of the saturated IP9220 ferrite paste. The guide width of 2300 μm provides a fundamental cutoff frequency of 21.6 GHz and the via centre of 550 μm (with respect to the narrow wall) is optimally chosen to provide the highest degree of phase shift. Fig. 5.7 plots the instantaneous phase shift versus the longitudinal location in z from the via centre.

Fig. 5.7 *Instantaneous phase shift within the via versus the distance from the centre of the via using the parameters of Table 5.2.*

As expected, the phase shift is maximised at the centre of the via and is zero at its edge of $\varnothing/2 = 75 \mu\text{m}$. Using 1000 samples yields a summated phase shift of $0.479 \pm 0.01^\circ$ per via pair. The minimum via pitch along the length of the guide is $2.5\varnothing = 375 \mu\text{m}$ so the phase shift is 0.479° per $375 \mu\text{m}$ or $12.8^\circ/\text{cm}$, a value that is only 32% that of the maximum instantaneous phase shift at the centre of the via. Fig. 5.8 plots the total phase shift versus the placement of the via (with respect to the narrow wall) using the analytical method and compares the results with HFSS simulations.

Fig. 5.8 Phase shift of the two-post ferrite-loaded RWG having the parameters listed in Table 5.2 with a via pitch of $375 \mu\text{m}$, calculated analytically and simulated using HFSS.

The agreement between the two curves is excellent. The small discrepancy is attributable to the 12-sided polygon representation of the cylindrical vias in HFSS. Unfortunately, the finite via pitch significantly reduces the achievable phase shift; in the next section, a field analysis of a four-post ferrite-loaded RWG is derived in an attempt to improve this value.

5.4 Four-post ferrite-loaded RWG field derivation and phase shift calculation

The analytical procedure for determining the phase shift through a four-post via-loaded RWG is the same as that used for the two-post case. Unlike the two-post case, however, the transcendental equation of a four-slab ferrite phase shifter is unknown and must be derived from first principles, where it is again assumed that the fundamental mode is similar to the canonical TE_{10} mode. The E-plane cross-section of half of a four-slab ferrite-loaded RWG is shown in Fig. 5.9, where a magnetic wall has been placed in the middle of the guide.

Fig. 5.9 *E-plane cross-section of half of a four-slab ferrite-loaded RWG, explicitly showing the five field regions I–V.*

For the ferrite regions, Maxwell's equations are given by [6]:

$$\vec{\nabla} \times \vec{E} = -j\omega\vec{\mu}\vec{H} \quad (5.5a)$$

$$\vec{\nabla} \times \vec{H} = j\omega\epsilon\vec{E} \quad (5.5b)$$

where

$$\vec{\mu} = \begin{bmatrix} \mu & 0 & j\kappa \\ 0 & \mu_0 & 0 \\ -j\kappa & 0 & \mu \end{bmatrix} \quad (5.6)$$

is the permeability tensor in a $-y$ -biased ferrite [6], the elements of which are:

$$\mu = \mu_0 \left(1 + \frac{f_0 f_m}{f_0^2 - f^2} \right) \quad (5.7a)$$

$$\kappa = \mu_0 \frac{f f_m}{f_0^2 - f^2} \quad (5.7b)$$

where f is the operating frequency, $f_0 = \mu_0 \gamma_r H_0$ is the gyromagnetic resonance frequency, $f_m = \mu_0 \gamma_r M_s$ is the maximum domain rotation resonance frequency [7], H_0 is the bias field,

M_s is the magnetisation, and γ_r is the gyromagnetic ratio. Assuming propagation in the $+z$ direction with $\gamma = \alpha + j\beta$, the electric and magnetic fields may be expressed as:

$$\vec{E}(x, y, z) = \vec{e}(x, y)e^{-\gamma z} \quad (5.8a)$$

$$\vec{H}(x, y, z) = \vec{h}(x, y)e^{-\gamma z}. \quad (5.8b)$$

For a TE_{m0} mode, $e_z = 0$ and $\delta/\delta y = 0$. Expanding (5.5) leads to $e_x = h_y = 0$, leaving three equations:

$$\gamma e_y = -j\omega\mu h_x + \omega\kappa h_z \quad (5.9a)$$

$$\frac{\partial e_y}{\partial x} = -\omega\kappa h_x - j\omega\mu h_z \quad (5.9b)$$

$$j\omega\epsilon e_y = -\gamma h_x - \frac{\partial h_z}{\partial x}. \quad (5.9c)$$

A solution will be sought for e_y only, requiring the magnetic fields in (5.9a) and (5.9b) to be explicitly given as a function of e_y :

$$h_x = \frac{1}{\omega\mu_e} \left(j\gamma e_y + \frac{\kappa}{\mu} \frac{\partial e_y}{\partial x} \right) \quad (5.10a)$$

$$h_z = \frac{1}{\omega\mu_e} \left(j \frac{\partial e_y}{\partial x} - \frac{\kappa\gamma}{\mu} e_y \right) \quad (5.10b)$$

which upon substitution into (5.9c) yields the wave equation for e_y :

$$\left(\frac{\partial^2}{\partial x^2} + k_f^2 \right) e_y = 0 \quad (5.11)$$

where the x -directed wavenumber in the ferrite is given by

$$k_f = \sqrt{\omega^2 \mu_e \epsilon_0 \epsilon_{rf} + \gamma^2} \quad (5.12)$$

and where

$$\mu_e = \frac{\mu^2 - \kappa^2}{\mu} \quad (5.13)$$

is the scalar permeability of an extraordinary mode in an infinite, transverse-biased ferrite [6], and where ϵ_{rf} is the relative permittivity of the ferrite. The fields in the dielectric-filled regions I, III, and V may be obtained by setting $\mu = \mu_0$ and $\kappa = 0$, leading to the magnetic fields in the dielectric regions:

$$h_x = \frac{j\gamma}{\omega\mu_0} e_y \quad (5.14a)$$

$$h_z = \frac{j}{\omega\mu_0} \frac{\partial e_y}{\partial x} \quad (5.14b)$$

and the wave equation in the dielectric regions:

$$\left(\frac{\partial^2}{\partial x^2} + k_d^2 \right) e_y = 0 \quad (5.15)$$

where the x -directed wavenumber in the dielectric is given by

$$k_d = \sqrt{\omega^2 \mu_0 \epsilon_0 \epsilon_{rd} + \gamma^2} \quad (5.16)$$

where ϵ_{rd} is the relative permittivity of the dielectric. The e_y and h_z fields within the waveguide must be matched such that their tangential fields are equal at the ferrite/dielectric boundaries. Because Maxwell's equations are coupled, this will automatically enforce the remaining boundary condition of equal b_x field components at the interfaces. Although this procedure normally requires ten electric field and ten magnetic field equations, the condition of $e_x = h_y = 0$ reduces the *total* to ten equations. The fundamental mode is constructed in a manner that is similar to that of the dual-slab derivation that is given in [6]. The expressions for the transversal electric fields e_{yi} are:

$$e_{yi} = A \sin k_d x \quad (5.17a)$$

$$e_{yII} = B \sin k_d (x - c) + C \sin k_f (c + t - x) \quad (5.17b)$$

$$e_{yIII} = D \sin k_d (-x + c + t) + E \sin k_d (x - d) \quad (5.17c)$$

$$e_{yIV} = F \sin k_f (x - d) + G \sin k_f (d + t - x) \quad (5.17d)$$

$$e_{yV} = H \cos(a/2 - x) \quad (5.17e)$$

where the subscript denotes the region to which it applies, and where the PEC boundary condition at $x = 0$ requires the sinusoidal dependence of (5.17a). The fields in the other regions have been formulated to facilitate the amplitude matching at the boundaries, which will follow shortly. The longitudinal magnetic fields are also tangential to the boundaries, and are found by inserting (5.17) into (5.10b) and (5.14b) as required in all five regions:

$$h_{zI} = \frac{jAk_d}{\omega\mu_0} \cos k_d x \quad (5.18a)$$

$$h_{zII} = -\frac{\kappa\gamma}{\omega\mu\mu_e} [B \sin k_f (x - c) + C \sin k_f (c + t - x)] \\ + \frac{jk_f}{\omega\mu_e} [B \cos k_f (x - c) - C \cos k_f (c + t - x)] \quad (5.18b)$$

$$h_{zIII} = \left(\frac{jk_d}{\omega\mu_0} \right) (-D \cos k_d (c + t - x) + E \cos k_d (x - d)) \quad (5.18c)$$

$$h_{zIV} = -\frac{\kappa\gamma}{\omega\mu\mu_e} [F \sin k_f (x - d) + G \sin k_f (d + t - x)] \\ + \frac{jk_f}{\omega\mu_e} [F \cos k_f (x - d) - G \cos k_f (d + t - x)] \quad (5.18d)$$

$$h_{zV} = \frac{jk_d H}{\omega\mu_0} \sin k_d \left(\frac{a}{2} - x \right). \quad (5.18e)$$

The next step is to match the fields at all four boundaries, which leads to four electric field equations and four magnetic field equations. The electric tangential boundaries are:

$$x = c, \quad A \sin k_d c = C \sin k_f t \quad (5.19a)$$

$$x = c + t, \quad B \sin k_f t = -E \sin k_d e \quad (5.19b)$$

$$x = d, \quad -D \sin k_d e = G \sin k_f t \quad (5.19c)$$

$$x = d + t, \quad F \sin k_f t = H \cos k_d f \quad (5.19d)$$

and the magnetic tangential boundaries are:

$$x = c, \quad \xi B - C(\zeta - j\tau) = \nu A \cos k_d c \quad (5.20a)$$

$$x = c + t, \quad -\xi C + B(\zeta + j\tau) = \nu(E \cos k_d e - D) \quad (5.20b)$$

$$x = d, \quad \xi F - G(\zeta - j\tau) = \nu(E - D \cos k_d e) \quad (5.20c)$$

$$x = d + t, \quad -\xi G + F(\zeta + j\tau) = \nu H \sin k_d f \quad (5.20d)$$

where the auxiliary variables are given by the following:

$$\nu = \frac{k_d}{\mu_0} \quad \xi = \frac{k_f}{\mu_e} \quad \psi = \frac{\kappa \gamma}{\mu \mu_e} \quad \zeta = \xi \cos k_f t \quad \text{and} \quad \tau = \psi \sin k_f t. \quad (5.21)$$

Equations (5.19) and (5.20) can be rewritten in matrix form, which is included below.

$$\begin{bmatrix} \sin k_d c & 0 & -\sin k_f t & 0 & 0 & 0 & 0 & 0 \\ -\nu \cos k_d c & \xi & -\zeta + j\tau & 0 & 0 & 0 & 0 & 0 \\ 0 & \sin k_f t & 0 & 0 & \sin k_d e & 0 & 0 & 0 \\ 0 & \zeta + j\tau & -\xi & \nu & -\nu \cos k_d e & 0 & 0 & 0 \\ 0 & 0 & 0 & \nu \cos k_d e & -\nu & \xi & -\zeta + j\tau & 0 \\ 0 & 0 & 0 & -\sin k_d e & 0 & 0 & -\sin k_f t & 0 \\ 0 & 0 & 0 & 0 & 0 & \zeta + j\tau & -\xi & -\nu \sin k_d f \\ 0 & 0 & 0 & 0 & 0 & \sin k_f t & 0 & -\cos k_d f \end{bmatrix} \begin{bmatrix} A \\ B \\ C \\ D \\ E \\ F \\ G \\ H \end{bmatrix} = \underline{0} \quad (5.22)$$

where $\underline{0}$ is the null 8×1 matrix. The system has a solution whenever the determinant of the leading matrix is zero. Any numerical analysis software may be used to find the determinant of the matrix and this has been done using Matlab. Before the results are discussed, another solution method is presented. Although it is laborious and extremely

tedious, an algebraic manipulation may be undertaken to reduce the eight boundary condition equations to a single transcendental equation. Assuming a lossless propagating mode $\gamma = j\beta$ and redefining $\psi = \kappa\beta/(\mu\mu_e)$ yields:

$$\mathcal{K} \left\{ \cos k_d e + \frac{\xi^2 \sin k_d e}{\nu \sin k_f t (\nu \sin k_f t \tan k_d f + \nu \sin k_f t - \zeta)} + \frac{\sin k_d e}{\nu} (\psi + \zeta) \right\} - \frac{\nu \sin k_f t}{\xi \sin k_d e} = 0 \quad (5.23a)$$

where

$$\mathcal{K} = \frac{-\xi}{\nu \sin k_f t \cot k_d c + \tau + \zeta} + \cos k_f t + \frac{\sin k_f t}{\xi} (\nu \cot k_d e - \psi) \quad (5.23b)$$

Solutions to this equation exist as a function of β whenever the left hand side of (5.23a) is equal to zero, a condition that can be found using any general root solving algorithm.

Either one of the solution methods may be used to solve the system. Regardless of the one chosen, β is the independent variable and is swept over a broad range. Using the material parameters of Table 5.2, with vias located at 360 μm and 830 μm from the narrow walls, Fig. 5.10 plots the left hand side of (5.23a) versus β .

Fig. 5.10 *Magnitude of the left hand side of the transcendental equation for a four-slab ferrite-loaded RWG having the material parameters given in Table 5.2, with vias located 360 and 830 μm from the narrow walls.*

The transcendental equation is very well-behaved, generating a smooth curve having near-unity maxima over the β range shown. The two roots are easily discernible at $\beta_- = 1797.1 \text{ rad/m}$ and $\beta_+ = 1909.5 \text{ rad/m}$. Next, the same material parameters will be used to find the roots of the determinant of the leading matrix of (5.22), the magnitude of which is graphed in Fig. 5.11.

Fig. 5.11 Magnitude of the determinant of the leading matrix in (5.22) for (a) $-3000 < \beta < 3000$ rad/m, and (b) $1800 < \beta < 2600$ rad/m.

Noting the scale of the vertical axes, the leading matrix in (5.22) is evidently ill-conditioned, yielding enormous determinant magnitudes over much of the β range. This contrasts with the transcendental equation method, which yields values that are no larger than unity over the same range. Fig. 5.11(a) indicates that there are potentially six roots: three positive ones and three negative ones. The first negative and positive roots are asymmetrical and are equal to the roots found using the transcendental equation method: $\beta_- = 1797.1$ rad/m and $\beta_+ = 1909.5$ rad/m. These roots correspond to clear zero-crossings of the determinant. The remaining four potential roots are symmetrical about $\beta = 0$ and do not yield any phase shift, occurring at $\beta_A = \pm 2212.8$ rad/m and $\beta_B = \pm 2489.6$ rad/m. These roots are *not* clear zero-crossings but are merely critical points ($\delta\delta\beta = 0$) occurring at magnitudes very close to zero. The ill-conditioned matrix makes it very difficult to determine whether or not these critical points cross zero because a small change in β near the roots causes a very large change in the magnitude of the determinant. Instead of

searching for the exact root, the validity of the roots may be verified by testing the boundary conditions, since a valid root must simultaneously satisfy all of the boundary conditions given by (5.19) and (5.20). A solution for β at a given frequency yields the x -directed wavenumbers k_f and k_d via (5.12) and (5.16), respectively, which then leads to the solution of all of the auxiliary variables through (5.21). By setting one of the amplitude terms ($A - H$) in (5.17) or (5.18) to a known value, the other seven unknown amplitude terms may be obtained. Because there are eight boundary conditions in (5.19) and (5.20) but only seven remaining unknown amplitude terms, two equations for the final unknown amplitude will result. For the root to be valid, both of these equations must yield the same value for the remaining unknown amplitude. If the magnitude A is set to a known value, the sequence leading to the two equations for H , labelled H_1 and H_2 , is obtained by manipulating (5.19) and (5.20) as shown below:

$$(5.19a): \quad C = A \frac{\sin k_d c}{\sin k_f t} \quad (5.24a)$$

$$(5.20a): \quad B = \frac{1}{\xi} (vA \cos k_d c + C[\zeta + \tau]) \quad (5.24b)$$

$$(5.19b): \quad E = -B \frac{\sin k_f t}{\sin k_d e} \quad (5.24c)$$

$$(5.20b): \quad D = \frac{1}{v} (\xi C - B[\zeta - \tau]) + E \cos k_d e \quad (5.24d)$$

$$(5.19c): \quad G = -D \frac{\sin k_d e}{\sin k_f t} \quad (5.24e)$$

$$(5.20c): \quad F = \frac{1}{\xi} (-vD \cos k_d e + vE + G[\zeta + \tau]) \quad (5.24f)$$

$$(5.19d): \quad H_2 = F \frac{\sin k_f t}{\cos k_d t} \quad (5.24g)$$

$$(5.20d): \quad H_1 = \frac{1}{v \sin k_d f} (-\xi G + F[\zeta - \tau]) \quad (5.24h)$$

Fig. 5.12 plots the percentage difference between H_1 and H_2 near the asymmetrical roots.

Fig. 5.12 Percentage difference between H_1 and H_2 versus the propagation constant near the asymmetrical roots at (a) $-\beta_- = -1797.1$ rad/m and (b) $\beta_+ = 1909.5$ rad/m.

The difference is nil at β_- and at β_+ , confirming the validity of these roots. The percentage difference in the vicinity of the new potential roots is shown in Fig. 5.13.

(a)

(b)

Fig. 5.13 Percentage difference between H_1 and H_2 versus the propagation constant near the symmetrical roots at (a) $\beta_A = 2212.8$ rad/m and (b) $\beta_B = 2489.6$ rad/m.

Fig. 5.13(a) indicates a 100% error at the false root β_A , which means that one of the H amplitudes goes to zero at β_A while the other value does not. A large difference is also noted in Fig. 5.13(b) for the β_B root. The conclusion is that the potential roots suggested by the brute-force matrix method are in fact false—the only roots are the asymmetrical ones. Equation (5.24) allows the fields to be plotted as a function of the cross-section in the guide, which is done in Fig. 5.14.

Fig. 5.14 Normalised field magnitudes of the reverse wave β_- along half the width of the guide: (a) h_z component, (b) e_y (and b_x) component.

The h_z component, shown in Fig. 5.14(a), is maximised at the edge of the guide and obeys its boundary condition of $h_z = 0$ in the middle of the guide. This magnetic field component is tangential to the ferrite/dielectric boundaries and is continuous there. The e_y component shown in Fig. 5.14(b) is also continuous at the boundaries and has a null at the narrow wall of the guide. The b_x field is identical to the e_y distribution and is also continuous at the boundaries. The forward wave's β_+ field distribution is not very different from that shown in Fig. 5.14.

As per the two-post case, the phase shift in a four-post phase shifter is found by discretising the vias in the longitudinal direction according to (5.4). Using the parameters in Table 5.2, the analytical phase shift calculated using the transcendental equation is 20.89 °/cm for vias located at 360 μm and 830 μm from the narrow walls. An HFSS simulation of the same geometry yields a phase shift of 19.63 °/cm when the vias are represented as 12-sided polygons, and this value increases to 20.19 °/cm when the

number of sides is increased to 20. The difference between the latter HFSS value and the analytical one is only 3.4%. The 20.89 °/cm phase shift represents a 63% improvement over the 12.8 °/cm that was obtained using only two ferrite posts, although the value is still only half of what is achievable using a dual-slab RWG phase shifter. The low phase shift is caused by the small volume that the ferrite occupies within the guide, a metric that is optimised in the next section.

5.5 The ferrite filling factor

The amount of achievable phase shift is strongly influenced by the location of the ferrite within the guide's cross-section. The magnetic field modes inside the ferrite regions of the waveguide that produce the non-reciprocal wave behaviour are circularly polarised, hence, a higher axial ratio will result in higher values of phase shift. The second factor that influences the amount of available phase shift is the ferrite filling factor. If the ferrite occupies more of the volume of the waveguide, a greater proportion of the fields will be subjected to the ferrite's gyrotropic influence, and hence more phase shift should result. Because ferrite vias are used to load the waveguide, an inevitable penalty is incurred compared to a waveguide that is filled with a slab of ferrite. However, the filling factor can still be maximised if the parameters are wisely chosen.

Vias cannot completely fill the waveguide in the longitudinal direction because they must obey a practical minimum pitch that is set by the LTCC design rules. Obviously, the minimum pitch in the longitudinal direction should be used to pack as many vias as possible down the length of the guide. As for the transverse direction, illustrated in Fig. 5.15, it is unclear whether small or large via diameters should be used.

Fig. 5.15 *H-plane cross-section of half of a four-post ferrite-loaded RWG.*

The H-fields in the ferrite-loaded waveguide, including the bias fields, are antisymmetric about the mid-point of the guide. Because of this, there should be an even number of posts in the cross-section of the guide. An odd number yields a post in the centre of the guide, which has no effect if its bias is unidirectional. Due to the symmetry, only one half of the waveguide need be considered. To simplify matters, the vias are assumed to be symmetrically arranged across the cross-section of the guide, obeying the same transversal pitch. In anticipation of the replacement of the guide's narrow wall with a row of metal vias, the ferrite vias must be offset from the narrow wall by a minimum distance c . The via pitch is often expressed as a function of the via diameter, $p = P\varnothing$, where the factor P has a minimum P_{min} that is set by the design rules. The volume that is occupied by the ferrite vias is then given by $V_F = \pi b N \varnothing^2 / 4$, where b is the height of the guide. The volume of the unit waveguide cell containing one row of vias is given by $V_U = ab\varnothing P_{min}$, where the longitudinal length of the unit cell has been chosen as the minimum via pitch allowed. The ferrite filling factor (F^3) is therefore:

$$F^3 = \frac{V_F}{V_U} = \frac{\pi N \varnothing}{4aP_{min}} \quad (5.25)$$

which is maximised by choosing the largest possible $N\varnothing$ product, subject to the condition that the minimum via pitch is not violated. This latter condition may be found using the geometry shown in Fig. 5.15, from which the pitch factor P may be isolated:

$$P = \frac{a - 2c - \varnothing}{\varnothing(N + 1)} \quad (5.26)$$

which must obey $P \leq P_{min}$. Rearranging (5.26) leads to the restriction of N , which, as integer, must obey the following floor function:

$$N = \left\lfloor \frac{a - 2c - \varnothing}{\varnothing P_{min}} - 1 \right\rfloor. \quad (5.27)$$

Equation (5.25) is a function of a and $\varnothing$. The width of the guide a cannot be freely chosen since it sets the cutoff frequency, whereas the $\varnothing$ values are limited to those via diameters available from the punching machine. The restriction on N is also a function of c , which has a lower limit since this is the spacing between the vias that make up the narrow wall and the ferrite vias. Further algebraic manipulations of (5.25) and (5.27) do not yield simple results because of the floor function in (5.27). Instead, (5.25) has been plotted versus the parameters a and $\varnothing$, where Fig. 5.16 shows the resulting F^3 for $c = 270 \mu\text{m}$.

Fig. 5.16 Ferrite filling factor F^3 versus via diameter and waveguide width (a) three-dimensional graph, (b) same graph, viewed from above, where shading now indicates the relative value of F^3 .

The discontinuities indicate a step change in the maximum allowable number of ferrite vias, N , and the first few domains of N have been annotated. The filling factor is always small and is generally in the range of 4–8%. For large-diameter vias and small waveguide widths, the minimum pitch requirement is always violated and no ferrite posts are allowed ($N = 0$). This means that a waveguide width of at least $a = 2200 \mu\text{m}$ is required if $200\text{-}\mu\text{m}$ diameter ferrite vias are to be used. If $150\text{-}\mu\text{m}$ vias are used, then only $N = 2$ vias may be used inside the guide over the guide width range shown, and results in a low filling factor. The $100\text{-}\mu\text{m}$ via diameter offers the most flexibility, permitting two, four, or six posts to be used. Although the filling factor is maximised whenever the allowable N undergoes a step change, this condition is restrictive because the minimum pitch is exactly satisfied, meaning that no movement of the posts in the cross-section is allowed. Some movement of the posts is desirable, for as will be shown in the next section, post

locations offering the highest circular polarisation axial ratio (CPAR) of the magnetic field maximise the phase shift.

5.6 Figure of merit: magnetic field circular polarisation axial ratio (CPAR)

The gyrotropic nature of the H-fields inside the ferrite are responsible for modulating the longitudinal propagation constant β . Hence we should expect the amount of the phase shift to depend on the axial ratio of the H-fields inside the ferrite.

Fig. 5.17 *Precession of an electron's magnetic moment m about the bias axis H_0 .*

Given the magnetic bias H_0 , an electron's magnetic moment m will precess about the bias axis until it spirals into alignment with it. When transverse RF magnetic fields are present, they cause the electron's magnetic moment to diverge away from the bias slightly, and its precession is repeated. If the transverse magnetic fields are circularly polarised in the same direction as that of the electron's precession, coupling to the electrons is maximised.

An example CPAR calculation will be undertaken for the symmetrical two-post case using the material system of Table 5.2 by sweeping the lateral position of the via along the cross-section and calculating the resulting phase shift and CPAR at every location. The CPAR of the H-fields is given by $|H_x|/|H_z|$ or by $|H_z|/|H_x|$, the choice of which is dictated by obtaining a value that is less than or equal to unity. Since both field

components vary across the ferrite cross-section, an integral of the CPAR is taken across the E-plane, yielding a better indication as to the quality of the circular polarisation within the cross-section on the whole:

$$\text{CPAR}_X = \int_{x_0-t/2}^{x_0+t/2} \text{CPAR}(x) dx \approx \Delta x \sum_{n=1}^{N_x} \text{CPAR}(n) \quad (5.28)$$

where Δx is the uniform segment length of the discretisation in x , and x_0 is the location of the via. The effective width of the ferrite $t(z)$ that is occupied by the ferrite varies depending on the location of the cross-section in the z -direction, which yields a maximum of $t(z) = \varnothing$ when the z -position crosses through the middle of the via. Because the effective ferrite cross-section varies in as a function of z , so will the magnetic fields, and the integral of (5.28) must be calculated for every location in z . Assuming that the vias are centred at $z = 0$ in the H-plane,

$$\text{CPAR}_{XZ} = \int_{-\varnothing/2}^{\varnothing/2} \int_{x_0-t(z)/2}^{x_0+t(z)/2} \text{CPAR}(x, z) dx dz \approx \Delta x \Delta z \sum_{m=1}^{N_z} \sum_{n=1}^{N_x} \text{CPAR}(m, n) \quad (5.29)$$

Equation (5.29) evaluates the quality of the CPAR throughout the entire via and it is expected to be closely correlated with the level of phase shift. Note that the effect of the filling factor is implicit here: given set values of N_x and N_z , a higher filling factor produces larger Δx and Δz values, weighting the summation accordingly. Although the CPAR is an important metric, if the field magnitudes are negligible at a given point in the ferrite, a high CPAR value will still not produce significant phase shift. Hence the CPAR value is weighted with the magnitude of the transverse H-fields according to:

$$\begin{aligned}
 \text{CPAR_WXZ} &= \int_{-\varnothing/2}^{\varnothing/2} \int_{x_0-t(z)/2}^{x_0+t(z)/2} \left(\sqrt{|h_x(x, z)|^2 + |h_z(x, z)|^2} \right) \text{CPAR}(x, z) dx dz \\
 &\approx \Delta x \Delta z \sum_{m=1}^{N_z} \sum_{n=1}^{N_x} \left(\sqrt{|h_x(x, z)|^2 + |h_z(x, z)|^2} \right) \text{CPAR}(m, n)
 \end{aligned} \tag{5.30}$$

For each z -cut through the via, the two β values are found by searching for the roots of the transcendental equation. The field values in the ferrite sections are then found by solving the amplitude coefficients of the modes, and the weighted CPAR via integral is then calculated. The results for the two β roots are generally not the same. The weighted CPAR figures of merit are plotted in Fig. 5.18 for the phase shifter described in Table 5.2, where the CPAR values have been normalised to the phase shift per centimetre.

Fig. 5.18 Weighted CPAR_WXZ figures of merit for the two-post phase shifter of Table 5.2 versus the offset of the via centres with respect to the waveguide narrow wall. The lower solid line is the weighted CPAR of the forward wave whereas the dashed line is that of the reverse wave. The bold dashed line is the sum of the two CPAR curves, normalised to the phase shift per centimetre, which is plotted as the bold solid line.

The narrow solid line is a plot of the positive (and larger) β_+ root's weighted CPAR integral, while the narrow dashed line is that of the negative root. Note that the location and magnitudes of the CPAR peaks are different. Their sum corresponds to the bold

dashed line, which has been re-normalised to correspond to the phase shift in degrees per centimetre, which is represented by the bold solid line. Note that the sum of the CPAR integrals has a peak very near to that of the peak in phase shift, as expected. The actual phase shift is a slower-varying function than the CPAR integral, so while the CPAR integral yields an approximation to the actual peak in phase shift, no additional information may be inferred from it.

5.7 Magnetostatic bias

The ferrite vias inside the laminated waveguide must be driven by a magnetostatic bias to control the phase shift. Closed magnetic circuits offer lower magnetic reluctance than do open ones, so the former are preferred to minimise the bias current. Two magnetostatic bias schemes are shown in Fig. 5.19, where the light-grey rectangles represent the waveguide cross-section and where the darker loops represent the ferrite.

Fig. 5.19 *E-plane cross sections of a two-post ferrite-loaded rectangular waveguide showing the complete magnetostatic circuit: (a) single toroid, and (b) double toroid. The light grey areas are the waveguide whereas the dark grey lines are the ferrite. The black segments are the DC bias lines, where the direction of the current is indicated by the $\odot$ and $\otimes$ arrowheads.*

The vertical ferrite segments are the via posts whereas the horizontal segments, the flux return arms, consist of printed thick film. The $\odot$ and $\otimes$ arrowheads indicate the required

direction of the bias current in the windings to produce the magnetic flux indicated by the white arrows. In configuration of Fig. 5.19(a), the width of the device is minimised and is equal to that of the guide itself. However, the upper and lower portions of the device require additional tape layers: one tape layer separates the via waveguide broad wall from the inner bias windings, and two additional layers are used to separate the ferrite return arm from the bias windings, meaning that six additional tape layers are required. The configuration of Fig. 5.19(b) uses two separate ferrite toroids on either side of the guide and uses only two extra tape layers, although the device occupies a larger cross-section. The primary advantage to the second configuration is that windings can be located on every layer that is occupied by the waveguide. If required, more bias windings can be accommodated by increasing the spacing between the outer ferrite posts and the waveguide.

In general, reluctance is calculated using the expression $\mathcal{R} = L / \mu A$, where L is the length of the ferrite member, μ is its permeability, and A is its cross sectional area. The reluctance of the thick film ferrite flux return arms is given by $\mathcal{R}_{\text{arm}} = L_{\text{arm}} / \mu t w$, where t is the film thickness and w is the width of the arm. The ferrite vias obey a pitch of p , so each ferrite arm may be considered as having a width $w = p$. The reluctance of the via posts is $\mathcal{R}_{\text{via}} = 4h_{\text{via}} / \mu \pi \varnothing^2$, where h_{via} is the height of the via and $\varnothing$ is its diameter. The ratio of the return arm reluctance to that of the via post is $\pi \varnothing^2 L_{\text{arm}} / 4 t p h_{\text{via}}$. A typical thick film thickness is 12 μm . The length of the return arms will be approximately half the width of the waveguide $a/2$, or 1150 μm if the width is $a = 2300 \mu\text{m}$. A reasonable post height is $h_{\text{via}} = 775 \mu\text{m}$, which is consistent with ten layers of 77.5- μm fired Heraeus CT2000 tape. Using 150- μm diameter vias with a pitch of $p = 2.5\varnothing = 375 \mu\text{m}$ results in a

reluctance ratio of $\pi\phi^2 L_{arm}/4tph_{via} = 5.8$, meaning that the reluctance of the return arm is nearly six times that of the via. This ratio may be halved if two return arms are used on the top and bottom of the toroid.

Not only do the return arms increase the reluctance, but their thinness restricts the amount of magnetic flux that reaches the interaction via inside the waveguide. Using Heraeus IP9220 paste, with a saturation magnetic flux density of 250 mT, two 12- μm thick return arms of width $p = 375\text{ }\mu\text{m}$ only carry 2.25 nWb of flux. The resulting flux density inside a 150- μm diameter via post is 127 mT, only half the saturation value. Hence it is impossible for the toroid to saturate the via that is inside the laminated waveguide. The structure is shown in Fig. 5.20, where the laminated waveguide has been excluded for sake of clarity. A three-dimensional magnetostatic field simulator, *Ansoft Maxwell 3D 5*, has been used to determine the flux density in the toroid, where the initial relative permeability curve of the IP9220 ferrite paste has been used [4].

Fig. 5.20 A three-dimensional view of a unit cell of one of the ferrite toroids, with two upper and two lower flux return arms. A nine-layer double-loop bias winding excites the left via, which transfers the flux through the return arms to the interaction via that is in the waveguide, on the right.

The 100- μm wide bias conductors required 200 mA of bias current to saturate the return arms. The front and back walls of the cell are planes of symmetry, where the magnetic field is forced to be purely tangential. Fig. 5.21 displays the results of the simulation.

Fig. 5.21 Screen captures of the magnetostatic field within the unit cell of a ferrite toroid for a bias current of 200 mA circulating through the nine-layer double loop: (a) within the return arm and (b) in the middle of the interaction via.

The return arm, depicted in Fig. 5.21(a), is close to saturation over most of its cross-section. The via post that is surrounded by the bias lines, although not shown, has a flux density of approximately 125 mT within its centre, confirming the approximation. The interaction via post (the one to be located inside the waveguide), shown in Fig. 5.21(b), has a flux density in the range of 80–95 mT throughout its cross-section, indicating that not all of the flux from the other via reaches it. Indeed, some of the flux lines loop back around the bias lines instead. Fig. 5.22 plots the expected phase shift at 35 GHz using 150- μm diameter vias.

Fig. 5.22 *Two-post ferrite loaded rectangular waveguide phase shift versus via placement at 35 GHz for a guide width of $a = 2300 \mu\text{m}$, a via diameter of $\varnothing = 150 \mu\text{m}$, a ferrite via pitch of $p = 375 \mu\text{m}$, and a flux density of $B = 90 \text{ mT}$.*

The penalty incurred for not fully saturating the ferrite vias is severe: the expected phase shift has been degraded to $4.5^{\circ}/\text{cm}$, an abysmally low value. Nonetheless, a two-post ferrite-loaded phase shifter prototype has been designed. A conceptual top view of the bias coils is shown in Fig. 5.23.

Fig. 5.23 *Top view of an LTCC layer containing laminated waveguide narrow wall vias, ferrite vias, and bias windings.*

Arrows on the conductor windings indicate the required current flow to generate the magnetic field inside the ferrite vias indicated by the white arrowheads $\otimes$ and $\odot$. The two upper rows of ferrite, represented by black dots, form one of the toroids, while the bottom two rows of black dots are those of the second toroid. The solid broad walls of the waveguide and of the return arms of the ferrite toroids are not shown; such a view is typical of an LTCC tape layer near the middle of the guide's height. The via catch pads indicate where the bias current makes a transition to another layer.

Coplanar waveguide probes are to be used to excite the circuit, and the required transition to laminated waveguide is shown in Fig. 5.24.

Fig. 5.24 Top and side view of the transition from coplanar waveguide-to-microstrip, and of the transition from microstrip-to-laminated waveguide using a via probe extending into the laminated waveguide through an aperture in the microstrip ground plane. The via walls are not shown for sake of clarity.

A first transition to microstrip is made at the probe pads, which then leads to a via probe that extends into the LWG through an aperture in the LWG's upper broad wall. The upper broad wall also serves as the microstrip's ground plane. A microstrip tuning stub extends beyond the via probe and its length has been optimised using HFSS to yield a narrowband match at 35 GHz. An identical transition is used at the output of the LWG to connect to the second port of the network analyser. A cross-section of the interaction region is shown in Fig. 5.25 along with the dimensions of the structure.

Fig. 5.25 *Cross-section of the two-post ferrite-loaded laminated waveguide phase shifter, showing the conductors in grey and the ferrite in black. All dimensions are in micrometres.*

The stack consists of fourteen layers of 77.5- μm thick Heraeus CT2000 tape, having a total thickness of 1.09 mm. Double toroid arms are used to improve the flux through the interaction vias. Each of the two bias windings consists of a nine-layer double-loop coil that wraps around the outer ferrite vias. The total longitudinal length of the interaction region is 14 mm, meaning that 29 ferrite via pairs exist along the length of the laminated waveguide.

5.8 The fabricated two-post ferrite-loaded laminated waveguide prototype

A total of six different two-post ferrite-loaded laminated waveguide phase shifters have been fabricated, each having a slightly different placement of the ferrite vias inside the guide. Several problems were encountered concerning the IP9220 ferrite paste. Normally, via paste is more viscous than conductor paste so that it can adhere well to the punched via holes. Given that only one sample of IP9220 was available, it had to be used both for printing and via filling. Although the paste worked well for printing, it was problematic during via filling. Fig 5.26 illustrates the via filling stage.

Fig. 5.26 *LTCC via filling procedure.*

During via filling, a squeegee sweeps over a stainless steel stencil, which rests on top of the green tape sheet. The stencil's apertures are aligned with the via holes in the tape, allowing paste to flow into the latter. The paste is drawn into the holes by suction that is provided by the vacuum under the porous table. A thin sheet of backing paper stops the flow of the paste and is removed after the filling stage. The addition of ferrite vias means that two stencils have been used, one for standard conductive vias and the other for ferrite.

The low viscosity of the IP9220 paste causes it to flow too quickly into the holes, thereby wetting the backing paper and seeping between the tape and the paper. Although thinners are commonly used to reduce the viscosity of pastes, thickeners do not exist. Because of this, many of the vias could not be filled completely, requiring a second pass of the squeegee. The via-filled tape had to be dried quickly, then flipped to remove the backing paper before the paste could cling to it. When this could not be done rapidly enough, the dried ferrite paste caused the backing paper to tack onto the green tape when the paper was peeled from the tape, requiring a scalpel to be used to scrape away the

dried paste and paper remnants from the tape. The blotting is illustrated in Fig. 5.27, which contains photographs of layers that bisect the laminated waveguide and the ferrite vias.

Fig. 5.27 *Photographs illustrating the blotting of the IP9220 ferrite paste on the back side of the green tape sheet after via filling.*

Although these problems lengthened the fabrication time considerably, the technical staff at VTT Electronics were nonetheless able to print and fill all of the required layers with IP9220 paste. During the fabrication, many modules were rejected, primarily due to misalignment of the layers. These modules served as test samples, which were snapped to allow a visual inspection of the waveguide structures. Fig. 5.28 displays the cross-section of two such samples.

Fig. 5.28 *Transverse view of two two-post ferrite-loaded laminated waveguide phase shifter samples made from Heraeus CT2000 LTCC.*

The contrast in the left image has been optimised to highlight the conductor traces. The 100- μm DC bias lines are visible on the outer left and right portions of the image. The two black ferrite toroids, having two return arms printed on their top and bottom portions, are also clearly visible in both images, as is cross-section of the laminated waveguide. Due to the violent snapping that was used to expose the part, some of the ferrite vias have fallen away. Although the waveguide itself is relatively well defined, *via bumps* are present on the upper and lower surfaces, a telltale sign of z-axis shrinkage mismatch between the ferrite via fill and the tape, which strains the tape layer interfaces near the vias and promotes the delamination of the layers. Fig. 5.29 provides a close-up view of two bad delamination sites.

Fig. 5.29 *Worst-case delamination sites around a stacked ferrite via.*

Fig. 5.29 illustrates how a delamination can extend into the fragile 100- μm bias lines, which is likely to damage them, ultimately resulting in open-circuited windings. In

addition to causing stack damage, the ferrite also displayed small cracks in its printed return arms, as illustrated in Fig. 5.30, which is a top view of a fired part that was manufactured without its very top layer of ceramic tape, exposing the printed ferrite.

Fig. 5.30 *Top view of the cracked return arms of two ferrite toroids.*

The upper and lower black strips are the return arms of the toroids, and the bumped ferrite vias are plainly visible. The cracks that run vertically are not problematic since the desired magnetostatic field lines are oriented in the same direction. However, any cracking in the diagonal and horizontal directions causes a sharp increase in the reluctance of the magnetic circuit and lowers the overall magnetisation.

Recognising the fragile nature of the bias lines, four different DC bias probe sites have been included in the design of the phase shifter, each tapping the bias windings at one-third intervals. Table 5.3 lists the conductance test results performed on six prototype phase shifters.

Table 5.3 *Results of the DC bias conductivity tests for the Heraeus phase shifters.*

Module	Full Winding	Tap 1	Tap 2	Tap 3
A	× open	× open	✓ pass	---
B	× open	× open	× open	× open
C	× open	× open	✓ pass	---
D	× open	✓ pass	---	---
E	✓ pass	---	---	---
F	× open	✓ pass	---	---

Of the six different modules, only E has bias lines that are 100% intact. The remaining five have at least one fault in their windings, and module B has at least one fault in all three thirds of its windings, causing it to fail completely. No microwave measurements have therefore been performed on module B. These results are not promising, for broken bias lines are an indication of delamination. Microwave measurements have been taken of remaining modules for bias currents ranging from -250 to $+250$ mA. Unfortunately, all but module A have scattering parameters that are indicative of material failure inside the waveguide. Fig. 5.31 displays the s_{21} transmission amplitudes of the five zero-biased phase shifters.

Fig. 5.31 *Transmission amplitudes of five Heraeus zero-biased two-post ferrite-loaded waveguide phase shifters (A, C–F).*

Apart from module A, the curves display terrible transmission coefficients. They defy analysis and will be hitherto abandoned as failures. Although module A displays a characteristic that is consistent with that of a cavity resonator, its transmission coefficients are simply too low to be of any practical interest. Further, no phase shift has been observed, even when very high bias currents in the range of 700 mA are applied.

5.9 Summary

This chapter has provided the theoretical framework required for the analysis of transverse-biased, ferrite via-loaded laminated waveguides in LTCC. The achievable phase shift of a laminated waveguide with rows of ferrite vias running longitudinally within it is obtained by discretising the waveguide in the longitudinal direction. Ferrite slab-loaded rectangular waveguide theory is then applied at cross-section location to calculate the instantaneous phase shift, which is then integrated to obtain the net phase shift per unit length. Applying this method to an example two-post ferrite-loaded LWG results in a phase shift of 12.8 °/cm at 35 GHz, which is only 32% of that of an equivalent phase shifter using dual ferrite slabs.

The phase shift may be improved by doubling the number of ferrite posts across the guide's cross-section. The absence of an analytical formulation for a four-slab ferrite-loaded RWG has necessitated its derivation from first principles using mode-matching. Two methods have been presented for solving the propagation constant in the device: one requires the roots of a transcendental equation to be found, whereas the other requires the roots of an 8×8 matrix determinant to be found. Although both methods yield identical results, the transcendental equation method is simpler and more reliable. An example calculation has shown that using four ferrite via posts in the guide's cross-section

increases the phase shift to 20.9 °/cm at 35 GHz, which, although being a 63% improvement over the two-post case, is still only 52% of that of an equivalent dual-slab RWG phase shifter.

The primary cause of the low phase shift obtained by using ferrite vias instead of slabs is that the interaction volume of the ferrite is reduced. Subject to the restrictions imposed by the LTCC design rules, the ferrite filling factor may be maximised by judiciously choosing the via diameter and the waveguide width. The improvement obtained is not very large, however, for the filling factor only ranges between 5–8% for the material system considered herein. The placement of the ferrite vias along the guide's cross-section also affects the phase shift, and it has been shown that the phase shift correlates well to the weighted axial ratio of the circular-polarised magnetic fields inside the vias. The phase shift is also maximised by saturating the flux density of the ferrite vias inside the LWG, which are more easily magnetised if they are part of a closed magnetic circuit that extends beyond the bounds of the guide. Such a structure may be implemented as a toroid composed of vertical vias and lateral thick film. Unfortunately, the finite thickness of the film restricts the amount of flux that can be delivered to the ferrite vias, which has a significant impact on the phase shift. For the test case considered in this chapter, three-dimensional magnetostatic simulations indicate that the limited flux density inside a two-post ferrite-loaded LWG reduces the phase shift from 12.8 °/cm to only 4.5 °/cm, a value that is far too small to be of any practical use.

Nevertheless, prototype two-post phase shifters have been manufactured to assess the compatibility of Heraeus IP9220 ferrite paste and Heraeus CT2000 LTCC. Although the materials are not designated for use together, their properties and preparation techniques

were deemed sufficiently similar to warrant the attempt. Unfortunately, IP9220 does not work well as a via-filling paste: its green-state viscosity is too low, and its z -axis shrinkage is insufficiently matched to that of CT2000. The resulting phase shifter prototypes suffer from severe delamination problems, which have caused high insertion losses, open-circuited bias lines, and no apparent phase shift.

Perhaps the most frustrating aspect of this work has been the difficulty in obtaining ferrite pastes suitable for use with LTCC. Although material suppliers are keen to promote the integration of passive devices, including ferrite inductors, as one of the main advantages of LTCC, the availability of special capacitor (high ϵ_r) and inductor (high μ_r) pastes is extremely limited, even for experimental purposes. Currently, the only other ferrite paste that could be potentially used with Heraeus CT2000 is CL90-7057, a sample of which was secured too late to be included in the fabrication of the prototype phase shifters. Even if a ferrite paste having good compatibility with LTCC could be secured, the next problem to overcome is the low amount of flux that is carried by the thick-film members of the toroids. Printing several layers of ferrite to thicken these members is an option, but it is likely to strain the LTCC stack. Ideally, ferrite tape layers would be used for the lateral toroid members, with filled ferrite vias descending into the laminated waveguides. Given that most of the obstacles are currently on the supply side, the design of ferrite-loaded laminated waveguide phase shifters should be suspended pending further improvements to the availability of new ferrite materials.

5.10 Chapter Five references

- [1] J. I. Steinberg, "An alternative technique for producing cofired ceramic circuits and components" in *Proc. National Electronics Packaging and Production Conf.*, NEPCON West '94, pp. 1751–1758.
- [2] C. Murray, J. Flannery, and S. C. Ó Mathúna, "A planar inductor fabricated using co-fired Wet Stack ferrite processing", in *Proc. Capacitor and Resistor Technology Symp.*, CARTS-Europe 1996, pp. 215–220.
- [3] B. C. Smith, J. Steinberg, and G. Yan, "Meeting the challenges of multilayer chip inductor manufacturing", in *Proc. Capacitor and Resistor Technology Symp.*, CARTS 1997, pp. 19–27.
- [4] R. S. Fielder, *Computer Aided Design and Fabrication of Magnetic Composite Multilayer Inductors*, M. Sc. thesis, Virginia Polytechnic Institute and State University, Blacksburg, Virginia, Dec. 2000, p. 35.
- [5] S. K. Koul & B. Bhat, *Microwave and Millimeter Wave Phase Shifters: Volume II, Dielectric and Ferrite Phase Shifters*, Dedham, Massachusetts: Artech House Inc., 1991.
- [6] D. M. Pozar, *Microwave Engineering*, Reading, Massachusetts: Addison-Wesley, 1993, chap. 10.
- [7] R. F. Soohoo, *Theory and Application of Ferrites*, Englewood Cliffs, NJ: Prentice-Hall, 1960, pp. 83–88.

Chapter Six — The ferrite-filled, antisymmetrically-biased

(Ferrite-FAB) rectangular waveguide phase shifter

The previous chapter has shown that rectangular waveguide (RWG) phase shifters using ferrite posts have marginal phase shift values and are difficult to build. Because of this, designs incorporating RWGs and ferrite paste have been abandoned. Fortunately, ferrite is also available in the form of LTCC tape, although at the time of this work, the tape was only available in limited quantities as an experimental material. The use of this new material implies that the entire LTCC module must be comprised of ferrite. To the author's knowledge, a RWG phase shifter that is completely filled with ferrite has never been documented, although work has been done on ferrite-filled *circular* waveguide phase shifters [1]–[3]. Ferrite-filled RWGs with *unidirectional* transverse bias are described in [4] and [5], but the emphasis is either on the guide's cutoff frequency or on its propagation constant: neither reference suggests using the waveguide as a phase shifter. As will be shown in this chapter, a simple modification to the classical ferrite dual-slab RWG phase shifter [6] can significantly improve its achievable phase shift. This observation has led to the development of a new device, called the ferrite-FAB phase shifter.

Section 6.1 of this chapter opens with the definition of phase shift in a nonreciprocal medium. Section 6.2 then reviews the operation and limitations of the ferrite dual-slab RWG phase shifter. Although dual-slab phase shifters are usually built using air-filled guide, Section 6.3 shows that the phase shift can be significantly improved by completely filling the guide with dielectric instead of air. The results also show that the phase shift is

maximised by filling the guide with ferrite. Given this latter observation, Section 6.4 presents the new field theory governing a ferrite-filled, antisymmetrically-biased (ferrite-FAB) rectangular waveguide, and describes the operation of the device. Section 6.5 outlines the design cycle of narrowband 35 GHz ferrite-FAB phase shifter prototypes that are to be fabricated using LTCC ferrite tape, and Section 6.6 presents their measured performance. Section 6.7 attempts to predict the performance by using the ferrite characterisation work of chapter 4. Although this prediction should have normally preceded the design of the phase shifters, the material parameters of the ferrite were only known after the phase shifters and characterisation structures had been fabricated. Section 6.8 concludes by summarising the findings of this chapter.

6.1 Definition of phase shift in a nonreciprocal medium

The general expression for a wave travelling in the $+z$ direction in a lossless medium is $A\cos(\omega t - \beta_z z)$, where A is an amplitude, $\omega = 2\pi f$ is the radian frequency, β is the phase constant, and z is the location along the line. The argument of the cosine, $\omega t - \beta z$, is the absolute phase of the wave. Consider such a wave that is frozen in time, as illustrated in Fig. 6.1(a).

Fig. 6.1 (a) A forward wave travelling in the $+z$ direction, from port 1 to port 2, and (b) a reverse wave travelling in the $-z$ direction, from port 2 to port 1.

Port 1 is the reference and so the origin $z = 0$ is placed there. Selecting time $t = 0$ for convenience, the absolute phase of the signal entering at port 1 is $\omega t - \beta z = 0$ (t and z are zero there) whereas the absolute phase at port 2, where the wave is exiting, is $\angle s_{21} = -\beta\Delta z$. The negative sign means that this was the absolute phase of the signal at port 1 some time in the *past*. A wave that travels more slowly down the guide (having a smaller phase velocity) has a larger phase constant β and will have a more negative $\angle s_{21}$ value.

A wave can also travel in the reverse direction, from port 2 to port 1, as illustrated in Fig. 6.1(b). This reverse wave obeys the expression $A\cos(\omega t + \beta z)$. Now port 2 is considered as being reference and so the origin $z = 0$ is placed there instead. Port 1 lies in the negative z direction, at a distance $-\Delta z$ from port 2. Setting the time to zero again, the absolute phase at port 2 is zero whereas that at port 1 is $\angle s_{12} = \beta(-\Delta z) = -\beta\Delta z$. This is the *same* value that was obtained for $\angle s_{21}$, so the absolute phase of the exiting signal is the *same* regardless of the direction of propagation. This type of medium is called *reciprocal*. An *unmagnetised* ferrite phase shifter is always reciprocal.

Next consider a *nonreciprocal* medium, where the absolute phase of the forward and reverse waves are different, as illustrated in Fig. 6.2(a).

Fig. 6.2 Nonreciprocal propagation (a) with a certain magnetic bias H_0 , and (b) with a reverse bias H_0 .

Fig. 6.2(a) is representative of a *magnetised* ferrite phase shifter that is being biased by a magnetic field H_0 , where the arrows around H_0 indicate the direction of bias. The absolute phase of the forward wave is defined as $\angle S_{21} \equiv -\beta\Delta z + \phi_{fwd}$ whereas that of the reverse wave is $\angle S_{12} = -\beta\Delta z + \phi_{rev}$, where $\phi_{fwd} \neq \phi_{rev}$. The value of ϕ is the *absolute phase shift* that the wave has undergone. A negative value of ϕ means that the wave has been slowed down by the magnetised ferrite.

A key point to note is that the ferrite phase shifters that will be introduced later are symmetrical in their construction, such that the magnetic fields H_0 looking down the phase shifter from port 1 are the *mirror image* of those looking down port 2. This being said, if the magnetic bias H_0 is reversed, as the arrows indicate in Fig. 6.2(b), the H_0 fields looking down port 1 will look like the old fields seen looking down port 2 and vice versa. This causes the absolute phase shift ϕ to reverse itself, causing the forward wave to have an absolute phase of $\angle S_{21} = -\beta\Delta z + \phi_{rev}$ and the reverse wave to have an absolute phase of $\angle S_{12} = -\beta\Delta z + \phi_{fwd}$. The *absolute phase shifts* have been interchanged. Hence the total *relative phase shift* (or simply “*phase shift*”) that the forward wave can undergo,

relative to the unmagnetised state, is $\Delta\phi = \angle s_{21} - \angle s_{12} = \phi_{\text{fwd}} - \phi_{\text{rev}}$, which has the same magnitude as that of the reverse wave. So to measure the phase shift $\Delta\phi$, one does not need to physically reverse the bias: the absolute phases of the forward and reverse waves for one bias direction, $\angle s_{21}$ and $\angle s_{12}$, are all that is required.

The amount of phase shift depends on the length of the phase shifter. It is often more convenient to speak of phase shift per unit length. By assigning new propagation constants to the forward and reverse waves in the presence of the first bias, such that $-\beta_+\Delta z = \angle s_{21} = -\beta\Delta z + \phi_{\text{fwd}}$ and $-\beta_-\Delta z = \angle s_{12} = -\beta\Delta z + \phi_{\text{rev}}$, results in:

$$\beta_+ = \beta - \frac{\phi_{\text{fwd}}}{\Delta z} \quad \text{and} \quad \beta_- = \beta - \frac{\phi_{\text{rev}}}{\Delta z} \quad (6.1)$$

which are the unequal propagation constants of the forward and reverse waves in the presence of bias, respectively. The phase shift per unit length is $\beta_+ - \beta_-$, which yields a positive result when the forward wave is slower than the reverse wave. For an unmagnetised phase shifter, $\beta_+ = \beta_- = \beta$ and the phase shift per unit length is zero.

6.2 Observations of ferrite dual-slab rectangular waveguide phase shifters

As was introduced briefly in Section 2.4.3, antisymmetrical, transverse-biased ferrites have been used extensively in dual-slab RWG phase shifters [6]. In its most fundamental form, the device consists of two full-height ferrite slabs, each of width t , that run longitudinally down the length of the guide, as shown in Fig. 6.3. The magnetic bias H_0 is oriented in opposite directions in each slab, transversally to the direction of propagation in the guide.

Fig. 6.3 A ferrite dual-slab rectangular waveguide phase shifter.

The transcendental equation for the TE modes can be readily derived for this configuration [6]. The solution of the transcendental equation, which is best done numerically, yields the two propagation constants for the guide: β_+ for the forward-travelling wave, and β_- for the reverse wave. In the work that follows, a particular dual-slab phase shifter found in [7] will serve as a reference. The ferrite slabs are assumed to have a relative permittivity of $\epsilon_{rf} = 13$, a remanent magnetisation of $M_r = 142$ kA/m (1786 G), and are located inside an air-filled waveguide of width $a = 2.286$ cm. Fig. 6.4 plots the resulting phase shift at 10 GHz for this device over a range of slab thickness values and for four spacing values.

Fig. 6.4 Dual-slab remanant phase shift versus slab thickness t and slab spacing S at 10 GHz for a ferrite having a relative permittivity of $\epsilon_{rf} = 13$ and a remanant magnetisation of $M_r = 142$ kA/m (1786 G).

Note how the phase shift always displays a broad maximum versus the slab thickness, so widening the slabs much beyond a filling factor of $t/a = 0.12$ does not yield any further benefit. In fact, doing so is disadvantageous: it increases the losses due to the electric and magnetic loss tangents of the ferrite, and it increases the amount of magnetic flux required to saturate the ferrite. Failure to fully saturate a ferrite is known to cause additional low-field losses if the operating frequency is below a certain threshold [8]. Given these problems, it is not surprising that ferrite-filled RWGs have not been pursued as phase shifters. Filling the space between the two slabs with a high permittivity dielectric alters the field configuration within the waveguide and has shown to increase the available phase shift [9], although this complicates the construction of the guide. Further, higher order propagating modes may establish themselves in this new configuration.

6.3 Behaviour of the dual-slab phase shifter in a dielectric-filled waveguide

Previous work has focused on air-filled waveguides. However, using the same equations as those used to generate the curves in Fig. 6.4, if the guide is filled with a dielectric, the phase shift behaviour of the dual-slab phase shifter changes as the relative permittivity of the dielectric ϵ_r is increased. This is shown in Fig. 6.5, where the width a of the X-band guide has been scaled down by the square root of the guide's dielectric constant ϵ_r .

Fig. 6.5 Dual-slab remanant phase shift versus slab thickness t for a slab spacing $S = 0$ at 10 GHz for a ferrite having a relative permittivity of $\epsilon_{rf} = 13$ and a remanant magnetisation of $M_r = 142$ kA/m inside a dielectric-filled waveguide.

As the permittivity is increased, the phase shift at low ferrite filling factors t/a is reduced but the phase shift at high filling factors increases to levels that are well beyond those of an ordinary air-filled dual-slab phase shifter. Interestingly, all of the curves for the $S = 0$ case intersect at a filling factor of 0.25. Fig. 6.5 indicates that if the guide is completely filled ($t/a = 0.5$) with an antisymmetrically biased ferrite, the resulting phase shift will be

maximised at a value that is *more than twice* the maximum value achievable using an ordinary air-filled dual-slab phase shifter (Fig. 6.4).

6.4 Ferrite-FAB field derivation

Although the ferrite-filled, antisymmetrically-biased (ferrite-FAB) phase shift can be estimated using the dual-slab theory, the latter fails when the filling factor t/a is exactly equal to its maximum value of 0.5. Furthermore, a simpler transcendental equation results if the derivation is carried out specifically for the geometry shown in Fig. 6.6.

Fig. 6.6 Ferrite-filled, antisymmetrically-biased (ferrite-FAB) waveguide.

The derivation begins with Maxwell's equations in a y-biased ferrite,

$$\vec{\nabla} \times \vec{E} = -j\omega\vec{\mu}\vec{H} \quad (6.2a)$$

$$\vec{\nabla} \times \vec{H} = j\omega\epsilon\vec{E} \quad (6.2b)$$

where

$$\vec{\mu} = \begin{bmatrix} \mu & 0 & -j\kappa \\ 0 & \mu_0 & 0 \\ j\kappa & 0 & \mu \end{bmatrix} \quad (6.3)$$

is the permeability tensor for a +y-biased ferrite [7], the elements of which are given by:

$$\mu = \mu_0 \left(1 + \frac{f_0 f_m}{f_0^2 - f^2} \right) \quad (6.4a)$$

$$\kappa = \mu_0 \frac{f f_m}{f_0^2 - f^2} \quad (6.4b)$$

where f is the operating frequency, $f_0 = \mu_0 \gamma_r H_0$ is the gyromagnetic resonance frequency, $f_m = \mu_0 \gamma_r M_s$ is the maximum domain rotation resonance frequency, and γ_r is the gyromagnetic ratio. Note that μ is a positive definite quantity whereas the sign of κ depends on the direction of the y -bias. At remanence, $\mu = \mu_0$ and κ is negative if the magnetisation is in the $+y$ direction, as it is between $0 < x < a/2$ in Fig. 6.6. It is assumed that the fundamental ferrite-FAB mode resembles the TE_{10} mode of a canonical rectangular waveguide. By assuming propagation in the $+z$ direction, with $\gamma = \alpha + j\beta$, the electric and magnetic fields may be expressed as:

$$\vec{E}(x, y, z) = \vec{e}(x, y)e^{-\gamma z} \quad (6.5a)$$

$$\vec{H}(x, y, z) = \vec{h}(x, y)e^{-\gamma z} \quad (6.5b)$$

For the TE_{10} mode, $e_z = 0$ and $\partial/\partial y = 0$. Expanding (6.2) then leads to $e_x = h_y = 0$, leaving three equations:

$$\gamma e_y = -j\omega\mu h_x - \omega\kappa h_z \quad (6.6a)$$

$$\frac{\partial e_y}{\partial x} = \omega\kappa h_x - j\omega\mu h_z \quad (6.6b)$$

$$j\omega\epsilon e_y = -\gamma h_x - \frac{\partial h_z}{\partial x} \quad (6.6c)$$

The magnetic field components may be expressed explicitly as a function of e_y by further reducing (6.6a) and (6.6b):

$$h_x = \frac{1}{\omega\mu_e} \left(j\gamma e_y - \frac{\kappa}{\mu} \frac{\partial e_y}{\partial x} \right) \quad (6.7a)$$

$$h_z = \frac{1}{\omega\mu_e} \left(j \frac{\partial e_y}{\partial x} + \frac{\kappa\gamma}{\mu} e_y \right) \quad (6.7b)$$

and substitution of (6.7) into (6.6c) leads directly to the wave equation:

$$\left(\frac{\partial^2}{\partial x^2} + k_x^2 \right) e_y = 0 \quad (6.8)$$

where the x -directed wavenumber is given by:

$$k_x = \sqrt{\omega^2 \mu_e \epsilon + \gamma^2} \quad (6.9)$$

and where

$$\mu_e = \frac{\mu^2 - \kappa^2}{\mu} \quad (6.10)$$

is the scalar permeability of an extraordinary mode in an infinite, transversally-biased ferrite [7]. The H-fields of the fundamental mode are assumed to obey an odd function (as opposed to an even one), which is also consistent with the bias arrangement. Thus the h_z component is assumed to be antisymmetric about $x = a/2$, whereas the h_x field need only be piecewise linear there. This means that a magnetic wall may be placed at $x = a/2$ and the fields need only to be solved between $0 < x < a/2$. Respecting the boundary condition of a perfectly electric conducting (PEC) wall at $x = 0$, the only solution to (6.8) becomes:

$$e_y = A \sin k_x x \quad (6.11)$$

and substitution of (6.11) into (6.7) completes the field derivation:

$$h_x = \frac{A}{\omega\mu_e} \left(j\gamma \sin k_x x - \frac{\kappa k_x}{\mu} \cos k_x x \right) \quad (6.12a)$$

$$h_z = \frac{A}{\omega\mu_e} \left(jk_x \cos k_x x + \frac{\kappa\gamma}{\mu} \sin k_x x \right) \quad (6.12b)$$

By applying the final Dirichlet boundary of $h_z = 0$ at $x = a/2$, (6.12b) yields the transcendental equation for γ , where k_x has been expanded according to (6.9):

$$\kappa\gamma + j\mu\sqrt{\omega^2\mu_e\epsilon + \gamma^2} \cot\left(\frac{a}{2}\sqrt{\omega^2\mu_e\epsilon + \gamma^2}\right) = 0 \quad (6.13)$$

Assuming a propagating mode $\gamma = j\beta$, reversing the bias direction changes the sign of κ in the first term and therefore (6.13) must yield a different solution for β . In other words, a nonreciprocal phase shift is expected. Interestingly, if one assumes that the mode is cutoff, $\gamma = \alpha$, (6.13) admits *no solution*: its first term will *always* be real while its second term remains imaginary. This means that the fields below cutoff do not obey the assumed antisymmetric field configuration. Although (6.13) is a compact equation, its behaviour is nonetheless intriguing. One of the most interesting aspects of (6.13) is that $\beta = 0$ does *not* correspond to the guide's cutoff frequency. This is shown in Fig. 6.7, which is a graph of the magnitude of the left-hand side of (6.13) taken at three different frequencies.

Fig. 6.7 Magnitude of the transcendental equation for a ferrite-FAB phase shifter at remanence having $\epsilon_{\text{eff}} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm. The three lines represent different frequencies: the solid line is taken at 12 GHz, the dotted line at 8 GHz, and the dashed line at 6 GHz.

Over most of the useful operating frequency range of the guide, (6.13) admits both a negative and a positive β root, consistent with the 12 GHz plot in Fig. 6.7. The original assumption according to (6.5) was propagation in the $+z$ direction ($e^{-j\beta z}$), so the positive root $\beta_+ = \beta_1$ is the forward wave's propagation constant. The negative root is representative of propagation in the $-z$ direction and so $\beta_- = -\beta_2$ is the reverse wave's propagation constant, which is *positive* when the second root β_2 is negative. For all frequencies of operation, $\beta_+ > \beta_-$. As the operating frequency is reduced, the curve shifts upwards and to the right, until a critical frequency $f_{\beta 0}$ is reached when $\beta_- = 0$. This frequency is derivable in the absence of bias ($f_0 = 0$, or remanence) by setting $\gamma = 0$ in (6.13), yielding:

$$f_{\beta 0} = \sqrt{\frac{1}{4a^2\mu_0\epsilon} + f_m^2} \quad (6.14)$$

At this frequency, no phase progression is observed in the z -direction along the phase shifter for the reverse wave; however, the reverse wave is still coupling power from the input to the output—it is *not* cut off. Equation (6.9) indicates that the $\beta_- = 0$ condition corresponds to an extraordinary mode bouncing back-and-forth in x between the narrow walls. As the frequency is lowered below $f_{\beta 0} = 8.25$ GHz, (6.13) admits two strictly *positive* roots, meaning that $\beta_- = -\beta_2$ is now negative, a solution consistent with the 8 GHz plot in Fig. 6.7. This means that the phase progression of the reverse wave is in the *positive* z direction, even though it is transferring power in the *negative* z direction. Essentially, the reverse wave phase-fronts appear to be travelling backwards ($+z$), even though the power is being delivered in the proper direction ($-z$). As the frequency is further reduced, the roots will approach each other until they coincide: this is the cutoff

frequency f_c , which occurs when the minimum critical point of the left hand side of (6.13) occurs exactly at zero, which is $f_c = 7.676$ GHz for the case under consideration. For sake of comparison, the cutoff in absence of magnetisation is $f_c = 6.557$ GHz. Operation below cutoff is illustrated by the 6 GHz plot in Fig. 6.7, which fails to admit a root. Maximum phase shift occurs at cutoff, since $\beta_+ - \beta_- = 2\beta_+$.

It is of interest to note that the expressions derived herein reduce to the canonical RWG results if the magnetisation is removed. Assuming a propagating mode $\gamma = j\beta$ and combining (6.9) and (6.4) at remanence leads to:

$$\beta = \sqrt{\omega^2 \mu_e \epsilon - k_x^2} = \sqrt{\omega^2 \mu_0 \epsilon (1 - f_m^2 / f^2) - k_x^2} \quad (6.15)$$

which reduces to the classical RWG propagation constant when no magnetisation exists, $f_m = 0$. Furthermore, if $f_m = 0$, then $\kappa = 0$ via (6.4b), meaning that the boundary condition of $h_z = 0$ at $x = a/2$ in (6.12b) requires $\cos(k_x a/2) = 0$, which for the fundamental mode is satisfied only by $k_x = \pi/a$, the expected RWG cutoff wavenumber. In addition, (6.14) also reduces to the classical cutoff frequency when $f_m = 0$.

The reverse wave always has $\beta_- < \omega\sqrt{\mu_e \epsilon}$, which by (6.9) indicates that k_{x-} is always real-valued. Therefore the reverse wave always has sinusoidal field distributions. Conversely, over most of the operating range, $\beta_+ > \omega\sqrt{\mu_e \epsilon}$, a condition that causes the forward wave's k_{x+} value to be imaginary according to (6.9), which in turn causes the forward fields of (6.12) and (6.13) to adopt hyperbolic distributions. The forward wave reverts to normal sinusoidal field distributions when the operating frequency is lowered beyond the point where $\beta_+ = \omega\sqrt{\mu_e \epsilon}$, or $k_{x+} = 0$. When $k_{x+} = 0$ exactly, the propagation

constant is again identical to that of an extraordinary wave, this time propagating in $+z$. In addition, when this condition is satisfied, the fields of (6.11) and (6.12) reduce to:

$$\begin{aligned} e_y &\approx A'x \\ h_x &\approx -\frac{A'}{\omega\mu_e}(\beta_+x + \kappa/\mu) \\ h_z &\approx -\frac{A'}{\omega\mu_e}((\kappa/\mu)\beta_+x + 1) \end{aligned} \quad (6.16)$$

where A' is a constant. Enforcement of the boundary condition $h_z = 0$ at $x = a/2$ in (6.16) requires $a\kappa\beta_+/(2\mu) = -1$. Combining this latter equation with $\beta_+ = \omega\sqrt{\mu_e\epsilon}$ yields the frequency at which the forward fields revert to sinusoidal distributions (again assuming remanence):

$$f = \frac{f_m^2 a \sqrt{\mu_0 \epsilon}}{\sqrt{f_m^2 a^2 \mu_0 \epsilon - 1/\pi^2}} \quad (6.17)$$

which is near 9.082 GHz for the phase shifter under consideration. The theoretical field magnitudes at 10 GHz are illustrated in Fig. 6.8, which have been confirmed by simulations using HFSS 8.5.

Fig. 6.8 Ferrite-FAB normalised forward and reverse field magnitudes for a phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm: (a) e_y fields, (b) h_x fields, and (c) h_z fields.

As expected, the reverse wave fields in all three graphs have sinusoidal distributions whereas those of the forward wave are approximately linear, indicative of hyperbolic functions. As required, the e_y fields obey their boundary condition of $e_y = 0$ at $x = 0$, whereas the h_z fields obey $h_z = 0$ at $x = a/2$.

Using the same material parameters as before, the theoretical X-band ferrite-FAB phase shift is shown in Fig. 6.9 and is compared directly to that obtained using an HFSS simulation.

Fig. 6.9 Theoretical and HFSS-simulated phase shift versus frequency for a ferrite-FAB phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm.

Both curves are in excellent agreement, confirming the theoretical development. At cutoff, $\beta_+ = -\beta_-$ and the phase shift per unit length ($2\beta_+$) is maximised. As the frequency is increased, the β_- root goes from negative to positive values, and the phase shift gradually tapers off. Fig. 6.10 plots the HFSS-simulated return loss versus frequency for a 16 mm length of magnetised ferrite-FAB waveguide that is fed by identical non-magnetised dielectric-filled waveguide sections at both ends.

Fig. 6.10 HFSS simulated return loss of the ferrite-FAB phase shifter at remanence having $\epsilon_{rf} = 13$, $M_r = 142$ kA/m, and $a = 6.34$ mm, with loss parameters $\tan\delta = 0.002$ and $\Delta H = 7.8$ kA/m (100 Oe).

Apart from the usual increase near cutoff, the ferrite-FAB appears to be well matched to conventional, dielectric-filled RWG, with a return loss of better than 20 dB over most of the useable bandwidth. This is an essential result because it ensures that the magnetised ferrite-FAB will be well matched to the non-magnetised waveguide that feeds the phase shifter portion of the guide.

6.5 Ferrite-FAB prototype design

Given that the entire waveguide is filled with ferrite, a sufficiently strong magnetic bias field is necessary to obtain high magnetisation values throughout its cross-section. This is especially true at low microwave frequencies, where the size of the guide may be considerable. Fortunately, the dielectric constant of the ferrite ϵ_{rf} scales down the guide width by a factor of $\epsilon_{rf}^{1/2}$, thereby reducing the total magnetic flux required. Considerable reductions of size may in fact be achieved since the relative permittivity of ferrite

generally ranges between 12 and 17. If the frequency of operation is above the threshold value of $f > 3f_m/\sqrt{5}$, low-field losses should vanish [8], [10], meaning that the ferrite does not need to be fully saturated for low loss operation. Since f_m is directly proportional to saturation magnetisation, ferrites having high M_s values have the highest threshold frequencies. The highest magnetisation for commonly available ferrites is in the range of 400 kA/m (5000 G), which yields an upper threshold frequency of approximately 19 GHz. Hence propagation through any ferrite above this frequency should be free of low-field losses, permitting the ferrite-FAB to operate as a variable-flux phase shifter.

If the frequency of operation is above the threshold of 19 GHz, the dimensions of the guide are on the order of those commonly used in LTCC circuitry. Attempts were made in 2001 to secure ferrite LTCC tape from three suppliers, namely Dupont, Ferro, and Heraeus. Of the three, only Ferro responded in the positive and donated experimental LTCC ferrite tape, said to be compatible with its A6-S conductor pastes and processing. Unfortunately, the supplier could not confirm the properties of the ferrite, including the dielectric constant and hysteresis curve. Given that sufficient materials were only available for the fabrication of one panel, the design of the prototype phase shifter presented a formidable task.

Laminated waveguide (LWG) has been used to synthesise the RWG inside the LTCC [7]. The height of the LWG is approximately $b = 420 \mu\text{m}$, or four layers of fired tape. The guide's nominal operating frequency is 35 GHz, so the via centre-to-centre width of the guide is $a = 2000 \mu\text{m}$, which yields a fundamental RWG TE_{10} cutoff frequency of $f_c^{10} = 21.6 \text{ GHz}$ if the relative permittivity of the ferrite is assumed to be 12. The via diameter is $\varnothing = 150 \mu\text{m}$ and the via pitch is $p = 500 \mu\text{m}$. Coplanar waveguide probes are to be used

to excite the circuit, so the required transition to laminated waveguide is shown in Fig. 6.11, where a probe descends into the guide approximately $\lambda_g/4$ from the reflector wall, as is typically done in commercial rectangular waveguide-to-coaxial transitions.

Fig. 6.11 Top and side view of the transition from coplanar waveguide-to-microstrip, and of the transition from microstrip-to-laminated waveguide using a via probe extending into the laminated waveguide through an aperture in the microstrip ground plane. The via walls are not shown for sake of clarity.

The transition is identical to the one that was previously described in Fig. 5.24. A first transition to microstrip is made at the probe pads, which then leads to a via probe that extends into the LWG through an aperture in the LWG's upper broad wall. The upper broad wall also serves as the microstrip's ground plane. A microstrip tuning stub extends beyond the via probe and its length has been optimised using HFSS to yield a narrowband match at 35 GHz. An identical transition is used at the output of the LWG to connect to the second port of the network analyser. No magnetic bias is applied in the regions of the transitions. Between the transitions is a length of LWG that is exposed to a

magnetostatic field supplied by windings that run beside and within the guide itself. The windings are illustrated in Fig. 6.12, which is a cross-section of the LWG in the interaction region of the phase shifter.

Fig. 6.12 Cross-sections of the laminated waveguide in the interaction region.

Windings exist on three LTCC layers: two winding columns are external to the waveguide whereas one is buried in the centre of the guide. Figure (a) is the cross-section taken through vias whereas (b) is taken between the vias. The waveguide core is represented by the shaded area, and the dashed lines represent the DC magnetic field.

Although it is unconventional to place windings inside the LWG, the planarity of the layers guarantees that the electric field will be everywhere perpendicular to the windings such that the TE mode should remain unaffected by their presence. The silver waveguide walls are nonmagnetic and increase the reluctance of the magnetic circuit. Manifolds have therefore been introduced in the broad walls of the waveguide to help establish a low-reluctance magnetic field path. HFSS simulations indicate that these manifolds do

not significantly perturb the quasi- TE_{10} mode of the waveguide. The exact dimensions of the device are shown in Fig. 6.13.

Fig. 6.13 Cross-sectional dimensions of the ferrite-FAB interaction region, all units in micrometers. The via diameter is 150 μm and the layer thickness is approximately 105 μm .

Three variations of the ferrite-FAB have been constructed: *FAB A* and *FAB B* are exactly as shown in Figs. 6.11 and 6.12 but are of different lengths, *FAB D* has no manifolds, whereas *FAB C* has manifolds but no internal bias lines, thus allowing the effects of the manifolds and internal bias lines to be determined. The interaction length of the *A*, *C*, and *D* phase shifters is 2.05 cm. The DC bias lines are series-wound and lead to probe pads on the surface of the module. The winding ensures that the current in the outer two winding columns is in the opposite direction of that in the internal winding column.

6.6 Prototype measurements

Fig. 6.14 is a photograph of one of the fabricated modules, containing the two phase shifters *FAB A* and *FAB B*.

Fig. 6.14 Photograph of the fabricated LTCC module containing ferrite-FAB phase shifters 'A' and 'B'.

The GSG pads and microstrip lines are plainly visible on the left and right hand sides. The patterns located on the upper-right hand side of each phase shifter are the DC bias pads. Fig. 6.15(a) plots the magnitude of s_{11} and s_{22} of *FAB A* for twenty one bias levels ranging from -500 to $+500$ mA, whereas Fig. 6.15(b) plots the magnitude of s_{21} and s_{12} . Phase shifter *B* has an identical response.

Fig. 6.15 Measured (a) reflection and (b) transmission of the ferrite-FAB phase shifter 'A' for 21 bias levels ranging from -500 to $+500$ mA. The black lines are s_{n1} whereas the grey lines are s_{n2} , where $n \in \{1, 2\}$.

Although the transitions to the LWG have been optimised to provide a match at 35 GHz, the best match of -13.5 dB appears to occur in the vicinity of 36 GHz and is slightly

nonreciprocal: port one matches at 35.9 GHz whereas port two matches at 36.6 GHz, regardless of the bias level. Over the passband of 36 to 37 GHz, the insertion loss of the phase shifter including the transition loss is 5.8 ± 0.4 dB for all 21 bias levels and is quite stable even though the reflection coefficients have large variations over the same band. Because the two narrowband matches at 35.9 and 36.6 GHz do not significantly improve the transmission coefficients, it is suspected that the power is being absorbed by the resonant matches of the transitions and/or being radiated away by them. Ignoring multiple reflections within the LWG, compensating for the *minimum* return loss of 13.5 dB only reduces the insertion loss to 5.6 dB. However, compensating for the *maximum* (and nearly reciprocal) return loss of 4.0 dB at 37 GHz yields an insertion loss of 3.6 dB. The absence of a good match makes it difficult to obtain more accurate estimates of the insertion loss. Fig 6.14(a) plots the phase shift at 36 GHz versus the bias current, from the degaussed state up to +500 mA, then down to -500 mA.

Fig. 6.16 Measured phase shift of the FAB A shifter: (a) at 36 GHz versus bias, where the bias is varied in the direction of the arrows, and (b) at 500 mA bias versus frequency.

The maximum nonreciprocal phase shift at 36 GHz is 52.8° for the maximum bias current of 500 mA, and the remanent level is 23.3° . Note that the phase shift from the degaussed state to the maximum phase shift level is nearly linear: this indicates that the ferrite within the waveguide has not yet reached saturation, so it is likely that higher phase shift levels can be achieved using this material. However, although more current could be applied, the 125 μm -wide bias lines would probably not sustain higher levels. Furthermore, 3.3 volts are required to generate 500 mA, meaning that over 1.5 W of DC power is being internally dissipated by the windings. Heating of the ferrite degrades the phase shift and must be avoided. A redesign of the biasing arrangement, such as adding more windings, would likely yield better results. Fig. 6.16(b) is a graph of the phase shift at a bias level of 500 mA versus frequency. As predicted (c.f. Fig. 6.9), the phase shift decreases as a function of frequency. Within the passband of 36 to 37 GHz, the mean phase shift is $48.6 \pm 3.5^\circ$.

The reflection and transmission magnitudes of phase shifters *FAB C* and *FAB D* are very similar to those shown in Fig. 6.15, generating their optimum matches between 36 and 37 GHz. The manifolds and internal bias lines did not cause noticeable changes in the s-parameters. The phase-shift characteristics of *FAB C* and *FAB D* are shown in Fig. 6.17.

Fig. 6.17 Measured phase shift of modules *FAB C* (no internal windings) and *FAB D* (no manifolds), (a) at 36 GHz versus bias current, (b) at 500 mA bias versus frequency.

Phase shifter *FAB D* is not adversely impacted by the absence of the manifolds, hence the latter were unsuccessful at coercing more magnetic flux into the waveguide. The maximum phase shift of 61.1° developed by *FAB D* at 36 GHz is in fact slightly higher than the 52.8° developed by of the nominal phase shifter *FAB A*. *FAB D* has the highest phase shift of all four fabricated modules, with a mean value of $61.3 \pm 1.5^\circ$ between 36 and 37 GHz. Phase shifter *FAB C* is severely affected by its lack of internal bias windings, developing only 12.7° of phase shift at 36 GHz, or 20% that of *FAB D*. The external windings are therefore not very efficient at biasing the ferrite within the waveguide.

The most disappointing aspect of the phase shifters is their consistently high insertion loss of approximately 5.5 dB at 36 GHz. *FAB C* and *FAB D* both have approximately the same insertion loss, hence it cannot be attributed to either the manifolds or the bias lines inside the waveguide. Results of the ferrite characterisation indicated that the loss tangent of the substrate at 36 GHz is approximately 0.004. Using RWG formulae and assuming

silver conductors and a dielectric constant of 11, the attenuation constants of the TE_{10} mode at 36 GHz are 0.015 Np/cm due to conductor loss and 0.064 Np/cm due to dielectric loss. The total attenuation of the guide should therefore have been only 0.7 dB/cm, yielding a total expected insertion loss of 1.4 dB. As previously mentioned, an additional loss due to the ferrimagnetic linewidth of the ferrite is expected when it is biased, but the measured insertion loss is nearly constant regardless of bias, so linewidth cannot be the principal loss contributor. Furthermore, linewidth loss is a nonreciprocal phenomenon, whereas the measured s_{12} and s_{21} parameters are nearly indistinguishable. Hence there must be some other source of loss in the phase shifters that has yet to be identified. The only appreciable difference between the LWG cavities that were used to characterise the ferrite and the ferrite-FAB phase shifters is the microstrip-to-LWG transitions that are used in the latter. The LWG cavities were meant only to resonate at their natural frequencies, so no matching was required. The phase shifters, on the other hand, are optimised for a narrowband match at 35 GHz, which is achieved by inducing a resonance between the microstrip stub and the reflector wall shown in Fig. 6.11. The losses in these regions will be amplified due to the resonance, and in addition, any stray radiation from these resonant portions of the phase shifter will also contribute significantly to the insertion loss.

6.7 Validation of the observed phase shift

As previously mentioned, only sufficient materials and resources were secured for the fabrication of one ferrite LTCC panel, which contained both the characterisation structures and the phase shifters. The phase shifters were therefore designed without knowing the material parameters of the LTCC. Now that the ferrite has been

characterised, its material parameters may be used to estimate the expected phase shift. The process begins by using a two-dimensional magnetostatic simulator to find the flux density inside the waveguide produced by the bias lines. The simulator chosen for this task is *Maxwell 2D 9.0* from Ansoft Corporation. The nonlinear initial $B(H)$ curve from Fig. 4.x is the only material parameter required by the simulator. Due to symmetry, only one-quarter of the structure requires simulation, the result of which is shown in Fig. 6.18 for a 500 mA bias level. The cross-section of Fig. 6.18(a) is taken *through* vias whereas that of Fig. 6.18(b) is taken *between* vias in the waveguide's narrow wall.

Fig. 6.18(a) Magnetostatic flux density contour lines over one-quarter of the ferrite-FAB cross-section, through vias, at a bias level of 500 mA. Numbers indicate the flux density of the shades, expressed in teslas.

Fig. 6.18(b) *Magnetostatic flux density contour lines over one-quarter of the ferrite-FAB cross-section, between vias, at a bias level of 500 mA. Numbers indicate the flux density of the shades, expressed in teslas.*

Fig. 6.18(a) illustrates the high reluctance of the silver vias, which prevents most of the external B-field from penetrating into the waveguide. The majority of the flux within the guide emanates from the internal bias lines and are confined to the vicinity of those lines. The net B-field in most of the guide is quite low, ranging from 10–40 mT. The magnetic circuit of Fig. 6.18(b) has less reluctance and allows more of the external B-field to penetrate into the guide. The increase is only marginal, as the flux lines are still strongly bound to the conductors. The minimum field strength within the guide is now approximately 30 mT.

Only the vertical, or y-component, of the B-field contributes to ferrite-FAB operation. To obtain an estimate of the expected phase shift, B_y is integrated over the area of the waveguide, then divided by the total area to obtain an average, according to:

$$\overline{B}_y = \frac{\int (\vec{B} \cdot \hat{y}) dS}{\int dS} \quad (6.18)$$

Given that the via pitch is 500 μm and the via diameter is 150 μm , the vias block the magnetic path for 30% of the guide length, similar to Fig. 6.18(a), whereas no blockage exists for the remaining 70% of the guide length, as shown in Fig. 6.18(b). Letting $\bar{B}_{y1}$ denote the average B_y field in the via-blocked region and $\bar{B}_{y2}$ that in the via-free region, the mean B_y field down the length of the guide may be expressed as $0.3\bar{B}_{y1} + 0.7\bar{B}_{y2}$. Field calculations indicate that $\bar{B}_{y1} = 36.2$ mT and $\bar{B}_{y2} = 45.5$ mT, so the mean B_y field is 42.7 mT. This mean value has been used in the ferrite-FAB theoretical calculation of Section 6.4, yielding the predicted phase shift shown in Fig. 6.19.

Fig. 6.19 Phase shift obtained using the mean flux density from 2D simulations combined with a theoretical analysis, compared to that of the measured FAB A circuit.

The predicted values overestimate the measured values by approximately 20°. This can be attributed to two factors. First, it may be partly due to thermal heating. During the ferrite hysteresis characterisation, the properties were measured approximately one second after applying the bias. The phase shift measurements were downloaded remotely from a network analyser, a process that requires approximately ten seconds. The

relatively high current of 500 mA heats the ferrite module, and owing to the low Curie temperature, quickly degrades the phase shift. The second reason for the discrepancy is the rudimentary averaging process that was used to obtain the mean B_y field via (6.18) for it ignores the distribution of the microwave fields. Nonetheless, the estimated values are at least on the order of the measured values, and the slope of the curves is similar.

It is of interest to note the maximum theoretical performance of the ferrite-FAB. As previously mentioned, the expected insertion loss due to the silver conductors and the loss tangent of 0.004 at 36 GHz is 1.4 dB. If the saturation flux density of 230 mT could have been achieved inside the ferrite, the expected phase shift of the 2.05 cm-long interaction region would be 427.4° , resulting in a figure of merit of $427.4/1.4 = 305^\circ/\text{dB}$. If the ferrite material could be improved to yield a saturation magnetisation of 500 mT, the resulting phase shift and figure of merit would be improved to 990° and $707^\circ/\text{dB}$ at 36 GHz, respectively. Reducing the loss tangent below 0.004 would also improve these metrics. The figure of merit of $707^\circ/\text{dB}$ rivals that of the best documented Ka-band dual-slab air-filled phase shifter, being $750^\circ/\text{dB}$ at 36 GHz, using an optimally-formulated lithium-zinc ferrite with a saturation magnetisation of 460 mT [11]. It should be emphasised however that the ferrite-FAB phase shifter is more compact and offers a higher degree of phase shift per unit length. These qualities are masked by the figure of merit, since the loss of the ferrite-FAB is higher because it is filled with ferrite as opposed to being air-filled. The primary advantage of the ferrite-FAB over traditional dual-slab configurations is its tremendous ease of fabrication using LTCC as compared to conventional methods. The small dimensions and precision required to assemble the waveguides and ferrite toroids make conventional assembly complicated and costly at

millimetre-wave frequencies [11], [12]. Another advantage of the LTCC ferrite-FAB is that it may be readily integrated with onboard chips and even antennas [13]. A patent application for this design was filed with the United States Patent Office in 2003.

6.8 Summary

This chapter has presented the theory and measurements of a new RWG phase shifter, the ferrite-FAB. The theoretical phase shift of the ferrite-FAB is more than double that of a traditional ferrite dual-slab phase shifter having identical material parameters. A new and compact field formulation governing the operation of the ferrite-FAB has been presented and it has been confirmed by three-dimensional finite element simulations. Contrarily to the latter method, the analytical transcendental equation may be solved in a fraction of a second using a simple root solver to yield the predicted phase shift. The field components of the ferrite-FAB have been explicitly given, and the behaviour of the phase progression of the forward and reverse waves inside the device has been explained.

The dimensions of the ferrite-FAB phase shifter are ideally suited to an LTCC fabrication when the operating frequency is near or above 20 GHz, measuring only a few millimetres across and up to a few centimetres long. As such, the devices are conducive to very low cost mass production, at levels that are beyond those achievable by conventional RWG phase shifters, since the latter require mechanical assembly. Furthermore, LTCC offers an integration advantage to the ferrite-FAB because it can be integrated with semiconductor dice, antennas, and other components contained in the same package.

Prototype LTCC ferrite-FAB phase shifters have been successfully manufactured and tested. Unfortunately, a narrowband matching technique has been implemented at 36

GHz, masking the wideband behaviour of the phase shifter. In addition, the microstrip tuning stubs and waveguide reflector walls, used to produce the resonant match, are thought to be responsible for the high measured insertion loss of the prototypes, 5.5 dB (3.6 dB without reflection loss). Theoretical calculations using the measured quality factors of laminated waveguide cavities made from the same ferrite material indicate that the expected insertion loss is only 1.4 dB. In the future, an improved wideband transition from the coplanar measurement probes to the ferrite-FAB phase shifter must be sought.

A maximum measured phase shift of 52.8 °/cm for 500 mA of bias current has been achieved at 36 GHz, resulting in a figure of merit of 14.7 °/dB, which is comparable to the figures of merit of semiconductor phase shifters, as was presented in Chapter 2. The phase shift value is low because of the poor magnetic properties of the ferrite, especially its high coercivity of $H_c = 688$ A/m. The resulting hardness of the ferrite requires relatively high bias currents. The design of the bias windings must also be improved: of the three bias columns used in the prototype, the two columns external to the guide contribute only 30% of the magnetic flux within the guide, whereas the lone internal bias column generates the remaining 70%. Nevertheless, the ferrite-FAB has tremendous potential. Using the same materials, a maximum phase shift of 208 °/cm is theoretically possible if the ferrite within the guide can be fully saturated, thereby yielding a figure of merit of 305 °/dB. Improvements to the magnetic properties of the LTCC ferrite could potentially increase this figure to 710 °/dB, rivalling the best dual-slab designs even though the loss per unit length of the ferrite-FAB is higher.

6.9 Chapter Six references

- [1] C. R. Boyd, Jr., "A dual-mode latching reciprocal ferrite phase shifter", *IEEE Trans. Microwave Theory Tech.*, vol. 18, pp. 1119–1124, Dec. 1970.
- [2] C. R. Boyd, Jr., "A 60 GHz dual-mode ferrite phase shifter", in *IEEE MTT-S Int. Symp. Dig.*, 1982, pp. 257–259.
- [3] C. R. Boyd, Jr., "A latching ferrite rotary-field phase shifter", in *IEEE MTT-S Int. Symp. Dig.*, 1995, vol. 1, pp. 103–106.
- [4] R. F. Soohoo, *Theory and Application of Ferrites*, Englewood Cliffs, NJ: Prentice-Hall, 1960, pp. 207–209.
- [5] B. C. Anderson & Z. J. Cendes, "Solution of ferrite loaded waveguide using vector finite elements", *IEEE Trans. Magn.*, vol. 31, pp. 1578–1581, May 1995.
- [6] P. J. B. Clarricoats, *Microwave Ferrites*, New York: Wiley, 1961.
- [7] D. M. Pozar, *Microwave Engineering*, Reading, MA: Addison-Wesley, 1993, chap. 10.
- [8] J. J. Green & F. Sandy, "Microwave characterization of partially magnetized ferrites", *IEEE Trans. Microwave Theory Tech.*, vol. 22, June 1974, pp. 641–645.
- [9] S. K. Koul & B. Bhat, *Microwave and Millimeter Wave Phase Shifters: Volume II, Dielectric and Ferrite Phase Shifters*, Dedham, Massachusetts: Artech House Inc., 1991.
- [10] M. Igarashi & Y. Naito, "Microwave permeability tensor of partially magnetized ferrites", in *IEEE MTT-S Int. Symp. Dig.*, vol. 1, pp. 269–271, 1976.
- [11] H. Luo, Y. Liang, J. Yuan, and Q. Qiu, "A Ka band compressed grooved waveguide ferrite phase shifter", in *Proc. 4th Int. Conf. on Millimeter Wave and Far Infrared Science and Technology*, 1996, pp. 150–153.
- [12] R. W. Babbitt & R. A. Stern, "Fabrication and performance of ferrite phase shifters for millimeter frequencies", *IEEE Trans. Magn.*, vol. 15, pp. 1744–1746, Nov. 1979.
- [13] J. O. Litzenberger, M. Clénet, G. A. Morin, and Y. M. M. Antar, "Study of a waveguide antenna implemented in laminated material", Defence Research and Development Canada, Tech. Rep. TR 2002-132, Dec. 2002.

Chapter Seven — Conclusions, contributions, and future work

7.1 Conclusions

The following three subsections summarise the findings of the previous chapters. The first subsection presents the state of the art in millimetre-wave phase shifters, the second summarises the results of the LTCC characterisation program, and the final subsection presents a summary of the new ferrite LTCC phase shifters that have been explored in this work.

7.1.1 Millimetre-wave phase shifter state of the art and the potential role for LTCC

Chapter Two of this thesis has provided an extensive overview of current millimetre-wave phase shifter technology. The newest semiconductor phase shifters, using the most modern devices, are still unable to rival the high-frequency performance of older ferrite-loaded waveguide phase shifters. Ferrite designs have figures of merit that are generally higher than 500 %/dB, whereas those of semiconductor designs are only in the range of a few tens of %/dB. Even the newest switching devices, MEMS transducers, can only improve this value to approximately 100 %/dB. Given this wide discrepancy, semiconductor phase shifters are unlikely to outperform their ferrite counterparts in the foreseeable future. The switching speeds of semiconductors, however, are orders of magnitude faster than those achievable by ferrite devices, the latter being on the order of a microsecond, although this has not impaired the development of MEMS switches, which also have switching speeds in the microsecond range. No, the primary disadvantages to ferrite-loaded waveguide phase shifters are their three-dimensional nature and the very

ferrite material itself, which is not easily integrated. In today's planar paradigm, most three-dimensional components are dismissed as being bulky, difficult to manufacture, and difficult to integrate, which is, in fact, quite true of ferrite phase shifters, which are still assembled manually using expensive, miniature components. Although planar ferrite phase shifters have been proposed to remedy this, ferrite films are exotic, and their performance is no better than that achievable using MEMS. True three-dimensional ferrite-loaded waveguide phase shifters have yet to be implemented in an integrated, mass producible environment.

Given that practical magnetic interactions require lengths that are on the order of a wavelength, it is unlikely that ferrite components will ever be realisable in integrated, monolithic semiconductor form. At the other extreme lie the present-day ferrite phase shifters, which are centimetres long and manually assembled. Clearly, what is required is some intermediate medium between these two cases, one that can simultaneously provide the required interaction length while maintaining the desirable properties of integrability and low-cost mass production. New, experimental ferrite LTCC materials are likely candidates for this role. A multilayer LTCC package allows three-dimensional structures, including rectangular waveguide variants, to be built within it. As an IC package, the LTCC module is inherently integrable with monolithic circuitry and is amenable to high volume production. Further, with feature sizes that are orders of magnitude greater than those of ICs, the size of an LTCC package, on the order of a centimetre, is suitable for sustaining magnetic interactions at millimetre-wave frequencies.

7.1.2 Millimetre-wave characterisation of LTCC materials

Low-loss LTCC material systems have only recently been introduced, hence their properties at millimetre-wave frequencies are not well documented. In addition, if experimental ferrite materials are to be used, then these too must be characterised. Chapter Three has presented various *in situ* techniques for the measurement of conductor loss, substrate loss, and substrate permittivity in the microwave and millimetre-wave range. Various non-resonant characterisation methods, including the well-known TRL calibration method, have been found to be inapplicable to probed LTCC circuitry because of the unrepeatability of the probe placement. To alleviate this problem, resonant methods, which are insensitive to magnitude level shifts, have been used instead. Specifically, two-port gap-coupled microstrip ring resonators have been used to characterise both the Ferro A6-S and Dupont 943 LTCC systems, although the minimum LTCC-allowable gap length of 100 μm causes the rings to be very loosely coupled, sometimes causing their response to be lost in the noise of the network analyser. The DC conductivity of the top-layer silver conductors is excellent, being $5\text{--}6(10^7)$ S/m. The microstrip attenuation constant of both systems increases in a log-linear manner, from approximately 0.2 dB/cm at 10 GHz to 0.4–0.5 dB/cm at 40 GHz, which are indicative of good, low-loss systems. Translating these data into substrate loss tangent values yields somewhat noisy results, with those of the Ferro system in the 0.1–0.2% range whereas those of the Dupont system range from 0.15–0.25%. The extracted relative permittivity values of the substrates are very stable versus frequency, 6.14–6.22 for the Ferro system and 7.51–7.59 for the Dupont system. Laminated waveguide cavities have also been used for characterisation purposes. The rows of vias that are used in their narrow walls causes the equivalent waveguide width to

become ambiguous, hence cavity resonators are not well-suited for permittivity measurements. However, the cavity loss is a strong function of the loss tangent, permitting the latter parameter to be extracted accurately. The attenuation constant of the laminated waveguide in the Dupont 943 system is 0.4–0.5 dB/cm over the 35–45 GHz range, which is equivalent to a loss tangent of 0.26–0.36% $\pm 0.02\%$. The low loss tangent of the substrates, combined with excellent conductors and a controlled relative permittivity, make LTCC an excellent medium for the implementation of millimetre-wave passive circuitry.

Chapter Four has presented the full characterisation of the magnetostatic and millimetre-wave properties of an experimental ferrite LTCC tape provided by Ferro Corporation. The processing of the tape is identical to that of the Ferro A6-S system, except that the parts must be diced prior to being fired. Apart from this minor deviance, the tape is fully compatible with standard conductor pastes and can be readily used for high volume fabrication. The magnetostatic properties of the ferrite tape have been measured using an embedded solenoidal transformer, which, although measuring only 2.3 cm long, contains over 1.3 metres of contiguous conductor length. The saturation flux density and saturation magnetic field of the ferrite are $B_s = 230$ mT and $H_s = 3000$ A/m, respectively. The maximum linear relative permeability of the material is 97 and occurs at a magnetic field level of $H = 1170$ A/m. The remanent flux density is $B_r = 136$ mT, yielding a squareness ratio B_r/B_s of only 0.6, meaning that it is unlikely that the material can be used at remanence. The coercive force is $H_c = 688$ A/m, indicating that the ferrite is somewhat hard to magnetise and will require hundreds of milliamps of current for that purpose. Temperature dependence measurement have also been performed and show that

the Curie temperature of the ferrite is quite low, only $T_c = 117^\circ \text{C}$. This is problematic because the high currents required for bias are likely to heat the ferrite, thereby degrading its magnetic properties. In general, the magnetostatic properties of the ferrite are more appropriate for inductors than they are for gyrotropic microwave devices. Nevertheless, the ferrite has also been characterised at millimetre-wave frequencies using resonators. The first set of measurements uses a novel line-connected ring resonator that eliminates problems associated with the very loose coupling of gap-coupled resonators. The second set of measurements uses laminated waveguide cavity resonators. The attenuation of the $50 \, \Omega$ microstrip line ranges from 0.3–0.8 dB/cm over the 5–40 GHz range and is slightly higher than that obtained using conventional LTCC, the latter yielding attenuation levels of approximately 0.5 dB/cm at 40 GHz. The ferrite's product of relative permeability and relative permittivity increases slightly between 22 and 36 GHz, from 10.1–10.7 using cavities, and 10.5–11.4 using ring resonators. As was the case for conventional LTCC, the loss tangent values obtained from the cavity measurements are more stable than those obtained using the ring resonators, ranging from 0.37–0.48% over the 25–36 GHz range. Although the loss values are slightly higher than those of conventional LTCC, the ferrite is nonetheless an appropriate medium for millimetre-wave circuitry.

7.1.3 Ferrite-loaded waveguide phase shifters in LTCC

Another method of introducing ferrite into LTCC is the use of thick film ferrite paste. Three-dimensional structures may be built by printing ferrite traces on different levels and connecting them using ferrite-filled via holes. The theoretical framework for two- and four-post ferrite-loaded rectangular waveguide phase shifters has been derived in Chapter

Five. Using the saturated magnetostatic properties of a potential ferrite paste, the anticipated phase shift per unit length of a 35 GHz two-post ferrite loaded rectangular waveguide is only 12.8 °/cm, which improves to 21°/cm if the number of posts along the guide's cross section is increased to four. These low values are directly attributable to the small volume that the ferrite posts occupy with respect to that of the guide. The finite thickness of ferrite thick-film traces, which is on the order of 10 μm , significantly increases the reluctance of the magnetic circuit and limits its total magnetic flux. A magnetostatic simulation has shown that the thick film limits the flux density of the ferrite posts that load the waveguide, such that the anticipated phase shift of a two-post ferrite-loaded guide is only 4.5 °/cm at 35 GHz. Although prototype two-post phase shifters have been constructed, the ferrite paste has proven to be incompatible with the Heraeus CT2000 LTCC with respect to shrinkage. As a result, the fabricated prototypes show signs of delamination and yield very poor microwave performance. Even if the prototypes had been successfully manufactured, using ferrite paste alone in LTCC is unlikely to yield large amounts of phase shift.

The problems associated with high reluctance are curtailed if the entire medium is composed of ferrite, as is the case for LTCC ferrite tape. However, rectangular waveguide phase shifters that are completely filled with ferrite are not often found in the open literature. Although ferrite-slab dielectric waveguide phase shifters have been demonstrated, these use a small rod of ferrite and, as such, are not integrable in LTCC. Dual-mode phase shifters use ferrite-filled guide but require complicated quarter-wave polarisers and mode suppressors. Instead of these, a new theory for a completely filled ferrite rectangular waveguide phase shifter, the ferrite-FAB, has been elaborated and has

shown to yield excellent amounts of phase shift, more than twice the amount obtainable using conventional dual-ferrite-slab air-filled rectangular waveguides. Theoretically, using the measured properties of the experimental ferrite tape, a phase shift of up to 208.5 °/cm is obtainable with a corresponding figure of merit of 305 °/dB, values that are far superior to the theoretical values of the ferrite-post waveguide configurations. Four ferrite-FAB prototypes have been successfully fabricated and measured, each measuring 3.5 cm long and having an internal interaction length of 2.05 cm. Although reasonably good resonant matches of approximately 13 dB have been observed at 36 GHz, the transitions leading into and out of the phase shifters are narrowband and dissipate power due to their resonance, making insertion loss measurements very difficult. Using this data directly, the best insertion loss estimate is 3.6 dB, a value that is much higher than the anticipated value of 1.4 dB that is based on the measured material properties of the ferrite tape. Measured non-reciprocal phase shifts of up to 61° have been observed between 36–37 GHz and are presently limited by the high levels of required bias current. Although more phase shift can theoretically be obtained by further increasing the bias current, the resulting ohmic heating reduces the phase shift due to the low Curie temperature of the material. Although the prototypes have successfully demonstrated the concept, better biasing arrangements are needed, as are more wideband, non-resonant transitions into the waveguide.

7.2 Thesis contributions

This work has provided numerous contributions to the fields of millimetre-wave ferrite phase shifters and millimetre-wave LTCC circuitry, beginning with an exhaustive overview of millimetre-wave phase shifters, in which the performance of many different

types of phase shifters have been compared using consistent metrics. As such, the review serves as a benchmark reference and as a pertinent introduction to those entering the field.

The microwave properties of LTCC systems are not widely known and can vary depending on the particular fabrication conditions. Non-resonant LTCC characterisation techniques have been shown to be unreliable; using resonant methods, Dupont 943 and Ferro A6-S LTCC have been characterised from 5–40 GHz and both demonstrate excellent microwave properties, having low loss and a stable dielectric constant. The data are very useful to those designing microwave circuitry in these systems, and VTT Electronics is already using the results to promote their LTCC fabrication laboratory. In addition, an experimental Ferro ferrite tape has been rigorously characterised, including its magnetostatic and microwave properties. The results have provided Ferro with useful feedback and have demonstrated the feasibility of the production of ferrite components.

Many contributions have been made with respect to the analysis of passive microwave circuits, especially concerning microstrip ring resonators. The equivalent circuit of an unloaded ring has been improved, and, for the first time, a simple lumped-element equivalent circuit for a gap-coupled two-port ring resonator has been presented and used to estimate the frequency pushing effects of the capacitive gaps. A method has also been elaborated for the evaluation of these capacitors using simple 2.5D simulations (c.f. Appendix B).

Given that gaps do not always couple sufficient power to the resonator, a new configuration for the ring, termed line-connected, has been disclosed, where the ring is directly tied to both of its feed lines. The impedance of this resonator has been rigorously solved from first principles and a lumped element equivalent circuit near resonance has

also been presented. The operation of the ring has been confirmed experimentally, demonstrating its immediate usefulness in applications where very small coupling gaps cannot be manufactured.

Another resonator, the laminated waveguide cavity, has been studied extensively, and it has been shown that its conductive and dielectric losses are very well approximated by canonical rectangular waveguide formulae. A method has also been presented for estimating the radiation quality factor of the cavity using a commercial 3D finite element simulator. A new technique for predicting the equivalent length and width dimensions of the cavity has also been presented, the grid impedance method. These analyses are useful to all those seeking to implement laminated waveguides and cavities.

An elegant and useful means of assessing the magnetostatic properties of ferrite LTCC tape has also been demonstrated: for the first time, an *in situ* solenoidal transformer, entirely embedded in LTCC, has been used to fully characterise the $B(H)$ hysteresis curve of the material. This relatively simple circuit can be easily and inexpensively included on subsequent batches of ferrite LTCC circuits for characterisation purposes. A thorough theoretical analysis of the transformer has been provided, and the salient hysteresis properties of the experimental Ferro ferrite material have been obtained.

Numerous contributions have also been made to the analysis of ferrite-loaded rectangular waveguide phase shifters. Using mode matching, a technique has been presented for the prediction of the phase shift through a guide loaded with two rows of ferrite vias, and the method has been verified using a commercial 3D solver. For the first time, the method has been successfully extended to a guide containing four rows of ferrite vias. The resulting transcendental equations, obtained analytically, are more easily

interpreted than are the results of numerical methods for finding matrix determinants. In addition, a technique for finding the maximum ferrite filling factor in an LTCC design rule environment has been presented. Maximum phase shift has also been shown to be closely related to the circular polarised axial ratio of the magnetic fields inside the ferrite vias.

A novel ferrite phase shifter, called the ferrite-FAB, has been presented for the first time, and has been shown to provide over twice the phase shift per unit length of a conventional dual-slab ferrite-loaded rectangular waveguide. The transcendental equation for the ferrite-FAB has been rigorously derived and the behaviour of its non-reciprocal forward and reverse propagation constants has been described. Prototypes have been successfully fabricated and measured, thereby proving the concept of a high-performance millimetre-wave phase shifter made entirely from packaging material. Given its excellent theoretical figure of merit, the ferrite-FAB shows tremendous potential for reducing the size and cost of millimetre-wave waveguide phase shifters. A patent application has been made for this phase shifter design.

Major contributions have been made to the theoretical analysis of resonator quality factors using s-parameter magnitudes, as described in Appendix C. Instead of relying solely on 3-dB half-power points, it has been shown that the unloaded, loaded, and external quality factors can be found by measuring virtually any double-sided bandwidth. Many bandwidth measurements may therefore be taken to improve the accuracy of a given quality factor estimate. Elegant, succinct expressions have been provided for the unloaded, loaded, and external levels for the bandwidth measurements of two-port transmission and absorption resonators, and also for one-port reflection resonators. To date, this is the only

work that simultaneously provides such expressions for all of the quality factors, and for all types of two- and one-port resonators.

7.3 Future Work

Much work remains to be done to complete the ferrite characterisation and to improve the performance of the ferrite-FAB prototypes. The only remaining loss parameter of the ferrite that has yet to be characterised is the gyromagnetic resonance linewidth ΔH . This loss is associated with the damping of the electron precession about the axis of magnetostatic bias H_0 . This additional loss should be most apparent when the ferrite is saturated, resulting in a loss dependence on bias strength. Such a dependence has not yet been observed in the ferrite-FAB prototypes, indicating that either this loss is low or that the ferrite has not been sufficiently biased for the loss to become apparent. Gyromagnetic resonance occurs at low microwave frequencies and is proportional to the bias field. Establishing a gyromagnetic resonance in a structure and measuring its 3-dB bandwidth is all that is required to estimate ΔH .

Although the ferrite-FAB prototypes have successfully demonstrated the concept, their performance must be further refined. The first task is to increase the strength of the bias field within the guide to obtain more phase shift. Ideally, the ferrite should be saturable without requiring impractical amounts of current. The bias windings running inside the centre of the waveguide currently provide most of the flux therein, whereas the lines external to the guide contribute negligibly. The primary cause of this is that the flux lines prefer to circulate in the region *between* the bias lines and the narrow wall of the guide rather than circulating *inside* the waveguide. A potential solution to the problem is to print thick film dielectric (capacitor) paste between the outer bias lines and the waveguide's

narrow wall. The high reluctance of the dielectric film would coerce the magnetic flux into the guide instead. Another solution is to conduct bias current longitudinally down the narrow walls of the guide itself. The current flow in both of the outer bias lines is in the same direction for ferrite-FAB operation, hence both narrow walls need not be DC-isolated from one another. Metallic laminations, running longitudinally and connecting the vias of each narrow wall would provide the shortest conduction path for the current. The only foreseeable problem is that the bias current would only flow once down the guide's length instead of flowing multiple times down multiple windings, potentially requiring high current levels.

The ferrite-FAB must also be improved in terms of its excitation. The current method uses a probe that descends into the guide from a microstrip feed line, with a microstrip stub and the waveguide reflector wall producing the required resonance for a narrowband match. Any dimensional discrepancy between the launchers at the input and at the output of the guide causes each one to resonate at slightly different frequencies, spoiling the match. More importantly, the dual-resonating circuits dissipate power, thereby masking the actual ferrite-FAB insertion loss. Methods for calibrating to the waveguide plane, such as TRL, do not work for LTCC, hence a better feed mechanism for the ferrite-FAB is required. Ideally, the feed should be distributed, wideband, and non-resonant.

Recently, researchers have demonstrated how laminated waveguide *antennas* can also be embedded inside LTCC packages. The next logical step is to combine the ferrite-FAB with an array of such antennas, thereby forming a fully-embedded phased array. Pursuance of this line of research will eventually lead to a complete LTCC millimetre-

wave phased array module with all of the required control chips, phase shifters, and antennas being embedded in a single, compact LTCC package.

Chapter Eight — Appendices

8.1 Appendix A: Specification data sheets of selected LTCC materials

8.1.1 Heraeus CT2000 LTCC specification data sheets

Heraeus

Advanced Materials Heratape CT2000

Description

Heraeus' new low temperature co-fireable tape called Heratape™ CT 2000 is designed to allow production of high performance wireless and multilayer circuits.

Heratape™ CT 2000 is available in sheet or roll form in thicknesses of 2.0 mils, and 3.9 mils. (Other tape thicknesses available by request).

Key Benefits:

- Silver conductor compatible
- Lead and cadmium free
- Excellent handling properties
- Dimensionally stable on 850°C refire
- For use in a variety of high frequency, high performance applications
- Low cost and flexibility in the design
- High Q
- Near zero T_r (temperature coefficient of frequency)

● Typical Fired Electrical Properties

Dielectric Constant

(@ 450 MHz, 25°C)
9.1 ± .1

Dissipation Factor

(@ 450 MHz)
≤ 1x10⁻³

Thermal Coefficient of Expansion

(25°C to 300°C)
5.6 ppm/°C

Breakdown Voltage

> 1 kV @ 25 μm

Surface Roughness

< 0.22 μm

Camber

Conforms to setter

Insulation Resistance

(@ 25°C)
10¹³ Ωcm

Temperature Coefficient of Frequency (T_r)

<10 ppm/°C (-40 to 80 °C)

● Recommended Processing Guidelines:

Printing:

Blank tape to the desired size. Tape can be processed on the Mylar® carrier for ease of handling or removed, (Mylar® side of the tape is the shiny side). The majority of the printing is done on the dull side with a 250-325 mesh screen. The shiny side can be printed on if printing is required on both sides of the tape.

Lamination:

Isostatic: Stack each printed and dried layer on a metal registration plate (1/4" stainless steel). In order to assist homogeneity, rotate each layer 90°. Place stack inside of a heat sealable bag. Evacuate and seal. It is critical that the bag be fully evacuated. Double bagging is recommended to prevent leakage.

Lamination Conditions

Pressure 4500 PSI (31-32 N/mm²) VTT uses 3500
Temperature 65-70°C
Time 5-10 minutes

(Nominal conditions are 4500 psi @ 70°C for 10 minutes).

- To decrease shrinkage, increase the pressure.
- To increase shrinkage, decrease the temperature.

Uniaxial: Uniaxial pressing is used to align and tack the layers only. Stack each printed and dried layer in a lamination die, with the dull side up. In order to assist homogeneity, rotate each layer 90°.

Lamination Conditions

Pressure 210 PSI (1.5 N/mm²)
Temperature 25°C
Time 1-5 minutes

Burnout and Firing in a Box Oven:

Heating Rate 5-5.5°C/minute
Peak Temperature 860-870°C
Dwell Time @ Peak 20-30 minutes
Cooling Rate approx. 3-6 °C/minute

Heraeus

Advanced Materials

Heratape CT2000

Low Temperature Firing Cycle

Setter for Laminated Parts
96% Alumina

● **Typical Unfired Properties**

Green Density
2.3 g/cm³

Tensile Strength
240 psi (dependent on test method)

Green Sheet Thickness
2.0 mils (50.8 μ m), 3.9 mils (99.1 μ m)

● **Compatibility**

Co-fireable

Inner Layer – Printable
Inner Layer – mageable
Via Fill
Top Conductor
Ground Plane

Silver System

TC2304HQ
TC2301PI
TC7301
TC2601
TC2303

Burnout and Firing in a Belt Furnace: Firing Cycle,
6 - 8 hours
Including Burnout
Peak Temperature 860-870°C
Dwell Time @ Peak 20-30 minutes

● **Typical Fired Properties**

Shrinkage (Co-Firing Process)

Lamination pressure was 4500 PSI (31 N/mm²), Firing in a box oven with a peak temperature @ 870°C and 30 minutes dwell time

x,y	11.5 %
z	14.0 %

Fired Density
> 3.05 g/cm³

Flexural Strength
310 MPa

PAM0101.1

8.1.2 Ferro A6-S LTCC specification data sheet

A6-S is a low cost Microwave Grade Low Temperature Co-Fire Ceramic Tape alternative to A6-M. A6-S features excellent dielectric properties of our famous A6-M but at a fraction of the cost. A6-S formulated with Ferro's patented Calcium-Boro-Silicate glass is designed to be used for low cost wireless frequency applications between 2.45-77 GHz. A6-S is compatible with our all Gold, and mix metal metallization. A6-S formulated with a newly formulated polymer system gives A6-S tape that can easily be laminated and cut in the green state newly formulated Silver Metallization, employing our novel technology SAMS (Self-Adhering Metallization) was formulated with the express purpose to bring a low cost metallization with the lowest resistance known. The low resistance results in unequal performance in your high frequency application a cost that you can afford. The use of SAMS technology

Typical Fired Physical Properties

Thermal Coefficient of Expansion (25 - 300°C)	8 ppm/°C
Tape Shrinkage	
X,Y	14.9 ± 0.2 %
Z (unlaminated to fired)	24 ± 0.2 %
Fired Density	2.45 g/cm ³
Flexural Strength (3 point bend)	>160 MPa
Thermal Conductivity	2 W/mK
(With Thermal Vias)	>50 W/mK

Note: All physical testing is performed on 10-layer tape laminates with no metallization, and processed according to FERRO's recommended processing guidelines.

Typical Processing Guidelines

Lamination	3,000 psi (21 MPa) 70° C, 5-10 minutes
Burnout	450°C for 2 hours ≤2°C/min. ramp to 450°C
Firing	850°C peak for 10 minutes, Ramp to peak temp 6°C-8°C/min.
Setters	Silicon Carbide, fused quartz. Bare alumina setters should not be used.

See FERRO A6 LTCC Processing Guidelines data sheets for full details.

Typical Fired Electrical Properties

Dielectric Constant (10 MHz)	5.9 ± 0.15
Dissipation Factor (10 MHz)	<0.2 %
Insertion Loss (10 GHz)	<0.18 dB/in
Bulk Resistivity	>10 ¹² Ω/cm

Note: All electrical testing is performed on parts manufactured according to Ferro recommended process guidelines with silver-based conductor systems.

resulted in a material with excellent solderability, leach resistance and both initial and age adhesion. A6-S comes in 60 and 100 micron at a thickness of 18 cm. The thin tape, 60 micron gives the electrical designer the added degree of freedom to design those high frequency circuits where headroom is critical for the design. In addition A6-S also come in our standard thicknesses of 5 and 10 mil at our standard width of 13 inches.

Compatible Materials Systems

FERRO A6-S should be used with thick film materials from the FERRO A6 LTCC All Silver-Based (ASB), All Gold-Based (AGB) or Mixed Metal Materials (MMM) Systems. See data sheets for details.

Storage Conditions

We recommend storing A6 in a cleanroom environment if at all possible. Temperature should be ~22°C (70°F) and RH ~ 40%. The tape should be kept in the sealed bags it arrived in until ready for use. Resealing after each use, if possible, is a plus.

8.1.3 Heraeus CL90-7057 ferrite paste specification data sheet

CL90-7057 is a thick film printing ferrite paste that has been formulated to be screen printed on alumina substrates and can be used with other Heraeus thick film pastes. CL90-7057 has a high permeability at 1 MHz. CL90-7057 fires at 900°C.

Key Benefits:

- High permeability at 1 MHz
- Fires to a dense microstructure
- Compatible with standard Heraeus conductors & resistors
- Fires at 900°C

Typical Properties:

Viscosity:

150±15 Kcps, Brookfield HBT,
spindle #14, @25°C., 10 RPM

Permeability:

150 @ 1 MHz

Q:

≥25 @ 1 MHz

Solids:

68.5% ±1

FOG:

<8µm (4th Scratch)

Color:

Brown

CL90-7057

Thick Film Printing

Ferrite Paste

Recommended Processing Guidelines:

Application:

Print using a 280 -325, 0.5 mil
emulsion stainless steel mesh screen
on an alumina substrate.

Drying:

150°C for 10 minutes.

Thinner:

RV-100

Firing:

10°C/min. from Room Temp to
900°C

Hold for 60 minutes

10°C/min. from 900°C to room
temperature

Thickness:

Fired 11-13 microns

Warranty:

Material guaranteed to meet
specifications for 6 months from date
of shipment.

Storage:

Store in a dry location at 10°C-
30°C. DO NOT REFRIGERATE.

Allow paste to come to room
temperature prior to opening.

Spatulate well before using.

8.1.4 Heraeus IP9150 ferrite paste specification data sheet

Heraeus

Component Metallizations

IP9150.1

Description:

IP9150.1 is a Nickel/Copper/Zinc ferrite paste that has been formulated for Heraeus' Wet Stack system. This material can be printed and dried multiple times and then cofired to form a dense monolithic component. IP9150.1 has a high permeability at 1 MHz. IP9150.1 fires at 900°C.

Key Benefits:

- High permeability at 1 MHz
- Fires to a dense microstructure
- Produces a cofired monolithic ceramic component
- Cd free
- Fires at 900°C

Typical Properties:
Viscosity:

135±15 Kcps, Brookfield HA, spindle #14, @25°C, 10 RPM

Permeability:

150 @ 1 MHz

Q:

≥50 @ 1 MHz

Curie Temperature:

~80°C

Solids:

70%±2

FOG:

<10µm (4th Scratch)

Color:

Brown

Recommended Processing Guidelines:
Application:

Print using a 280-325, 0.5 mil emulsion stainless steel mesh screen on a prepared substrate.

Drying:

100°C for 10 minutes.

Thinner:

RV-100

Firing:

2°C/min. from Room Temp to 220°C

Hold for 30 minutes

2°C/min. from 220°C to 350°C

Hold for 30 minutes

13°C/min. from 350°C to 900°C

Hold for 120 minutes

10-15°C/min. from 900°C to room temperature.

Thickness:

Dried: 18-20 microns

(Dried to Fired)

Average X & Y shrinkage 13.3%

Warranty:

Material guaranteed to meet specifications for 6 months from date of shipment.

Storage:

Store in a dry location at 10°C-30°C.

DO NOT REFRIGERATE.

Allow paste to come to room temperature prior to opening.

Spatulate well before using.

Note: The information contained in this data sheet is based on typical properties for the product. Actual shipment specifications for the product may vary. Please contact Heraeus for the product's shipment specifications

This concludes Appendix A.

8.2 Appendix B: Extraction of the gap capacitor values from a 2.5D simulation of a three-port microstrip T-gap

Equation (3.13) is derived by considering the lumped element model of the T-gap microstrip junction, which is repeated and annotated as Fig. B.1. Port 1 is excited and is de-embedded to the edge of the gap whereas ports 2 and 3 are terminated in matched loads.

Fig. B.1 Microstrip T-gap junction and its equivalent lumped element model.

It will be shown that the final equations are a function of the *magnitude* of s_{21} only, hence the length of transmission line d between ports 2 and 3 does not need to be included in the lumped element model. The goal is to extract values of C_p and C_g based on simulated values of s_{11} and s_{21} , hence the relationship between these four parameters must be found. The admittances to the circuit are derived first, which lead directly to the reflection Γ and transmission T coefficients on port 1:

$$Y_2 = \frac{j2\omega C_g Y_0}{2Y_0 + j\omega C_g} \quad (\text{B.1a})$$

$$Y_{in} = Y_2 + j\omega C_p \quad (\text{B.1b})$$

$$\Gamma = \frac{Y_0 - Y_{in}}{Y_0 + Y_{in}} \quad T = 1 + \Gamma \quad (\text{B.1c})$$

where Y_0 is the known characteristic admittance of all three ports. The time-averaged real power that is delivered to the input of the circuit is given by:

$$P_{in} = \frac{1}{2} |TV_1^+|^2 \Re\{Y_{in}\} = \frac{1}{2} |TV_1^+|^2 \Re\{Y_2\} \quad (\text{B.2})$$

whereas the time-averaged real *output* power, the sum of the real power delivered to ports 2 and 3, is:

$$P_{out} = \frac{1}{2} |V_2|^2 \Re\{2Y_0\} = |V_2|^2 Y_0 \quad (\text{B.3})$$

A network of purely reactive elements consumes no real power. The capacitors C_p and C_g are the only elements between the reference planes for the input and output power, hence in terms of real power, $P_{in} = P_{out}$. This leads to the following equality, yielding the magnitude of s_{21} explicitly:

$$\begin{aligned} |TV_1^+|^2 \Re\{Y_2\} &= 2|V_2|^2 Y_0 \\ \frac{|V_2|^2}{|V_1^+|^2} &= \frac{|T|^2 \Re\{Y_2\}}{2Y_0} \\ |s_{21}| &= \sqrt{\frac{|T|^2 \Re\{Y_2\}}{2Y_0}} \end{aligned} \quad (\text{B.4})$$

To find C_g , the real part of Y_2 in (B.4) is isolated and (B.1c) is used to expand T , yielding:

$$\Re\{Y_2\} = \frac{2Y_0 |s_{21}|^2}{|1 + s_{11}|^2} \quad (\text{B.5})$$

Next, Y_2 is expanded from (B.1a) so that its real part may be expressed explicitly:

$$Y_2 = \frac{j2\omega C_g Y_0}{2Y_0 + j\omega C_g} = \frac{j2\omega C_g Y_0 [2Y_0 - j\omega C_g]}{4Y_0^2 + \omega^2 C_g^2} \Rightarrow \Re\{Y_2\} = \frac{2\omega^2 C_g^2 Y_0}{4Y_0^2 + \omega^2 C_g^2} \quad (\text{B.6})$$

whereupon substitution of (B.6) into (B.5) yields the sequence:

$$\begin{aligned}
 \omega^2 C_g^2 &= [4Y_0^2 + \omega^2 C_g^2] \frac{|s_{21}|^2}{|1 + s_{11}|^2} \\
 \omega^2 C_g^2 \left[1 - \frac{|s_{21}|^2}{|1 + s_{11}|^2} \right] &= \frac{4Y_0^2 |s_{21}|^2}{|1 + s_{11}|^2} \\
 C_g^2 &= \frac{4Y_0^2 |s_{21}|^2}{\omega^2 [1 + s_{11}|^2 - |s_{21}|^2]} \\
 C_g &= \frac{2Y_0 |s_{21}|}{\omega \sqrt{1 + s_{11}|^2 - |s_{21}|^2}} \quad (B.7a)
 \end{aligned}$$

which is the desired result for C_g . Note that only the magnitude of s_{21} is needed whereas a complex value of s_{11} is required. The simulated value of s_{11} must therefore be de-embedded up to the edge of the gap before it can be used in (B.7a). To obtain C_p , the input admittance to the circuit is isolated from (B.1):

$$\begin{aligned}
 Y_{in} &= \frac{Y_0 [1 - \Gamma]}{1 + \Gamma} \\
 Y_2 + j\omega C_p &= \frac{Y_0 [1 - \Gamma]}{1 + \Gamma} \\
 C_p &= \frac{Y_0 [1 - s_{11}]}{j\omega [1 + s_{11}]} - \frac{2C_g Y_0}{2Y_0 + j\omega C_g} \quad (B.7b)
 \end{aligned}$$

which is calculated once C_g has been found. As a verification, C_p must be purely real even though the right hand side of (B.7b) is a subtraction of complex terms.

This concludes Appendix B.

8.3 Appendix C: The measurement of quality factors of one- and two-port resonators using scattering parameter magnitudes at arbitrary power levels

Resonators are ubiquitous circuit elements, and although the practical measurement of a resonator's Q should be fundamental, no one source systematically treats, using simple methods, all of the basic resonator forms, which in the microwave field includes one-port series and parallel topologies, as well as two-port ones. The lack of a consistent methodology for the rapid measurement of Q has prompted the following derivations.

Microwave resonators may be classified as either one-port or two-port circuits. Only the reflection coefficient is available for one-ports, so these are termed *reflection* resonators [1]. Two-ports that generate a spike in s_{21} (a trough in s_{11}) at resonance are termed *transmission* resonators whereas those that generate a spike in s_{11} (a trough in s_{21}) are called *absorption* resonators [1]. The use of s-parameter magnitudes is the easiest method by which to measure the bandwidth of a resonator, as these are the most familiar metrics used in the field. Although the classical 3-dB bandwidth method for determining the unloaded, loaded, and external Q is well established, one of its major disadvantages is that only three frequency points are used to determine the fractional bandwidth, even though the available frequency sweep may be composed of hundreds of data points. To remedy this problem, it will be shown that *any* measured bandwidth can be related to the quality factor, not just the classical half-power, or 3-dB bandwidth. A similar approach is taken in [2]–[4], but those methods are incomplete and do not clearly interpret the meaning of the bandwidth. In what follows, the arbitrary bandwidth is clearly identified as a power fraction and it is derived for *all* of the basic resonator types and for all three quality factors (unloaded, external, and loaded) using only s-parameter magnitudes. The

proposed method therefore combines the simplicity of using s-parameter magnitudes with the convenience of using an arbitrary bandwidth measurement.

In keeping with the theme of simplicity, the resonators considered herein are ideal in many ways: it is assumed that the Q is high enough so that neighbouring resonances do not effect the resonance in question, and it is further assumed that the coupling to the resonator is lossless, as are the transmission lines leading to the resonator. Further, for two-port resonators, the coupling coefficients at both ports are also assumed to be identical. Should any of these assumptions be violated, more precise techniques are available, the penalty incurred being added complexity [5]–[11].

It is well known that all resonators may be approximated by lumped element series or parallel RLC circuits in the vicinity of their resonant frequency, as shown in Fig. C.1 for an unloaded series resonator and in Fig. C.1(b) for an unloaded parallel one-port. The RLC model is not imposed on the resonator *a priori*, but results from the derivation of the resonator's parameters, such as its input impedance, near resonance.

Fig. C.1 Unloaded (a) series and (b) parallel lumped element resonators.

Analysis of these circuits leads to the following expressions for the input impedance and input admittance to the series and parallel resonators, respectively, which are valid only

at frequencies near resonance $\omega = \omega_0 + \Delta\omega$, where the single-sided bandwidth must be small, $\Delta\omega \ll 2\omega$ [12]:

$$Z_L = R + j\omega L - j\frac{1}{\omega C} \approx R + jRQ\frac{2\Delta\omega}{\omega_0} = R + jRQ\delta \quad (\text{C.1a})$$

$$Y_L = G + j\omega C - j\frac{1}{\omega L} \approx G + jGQ\frac{2\Delta\omega}{\omega_0} = G + jGQ\delta \quad (\text{C.1b})$$

where δ is the double-sided fractional bandwidth. Considering only the unloaded resonator for the time being, it is straightforward to show that the time-averaged real power absorbed by the resonator at any frequency is:

$$P_0(\omega) = \frac{1}{2} |V_g|^2 \frac{R}{|Z_L|^2} \quad \text{or} \quad P_0(\omega) = \frac{1}{2} |I_g|^2 \frac{G}{|Y_L|^2}. \quad (\text{C.2})$$

At resonance, $Z_{IN} = R$ and $Y_L = G$, causing the power absorbed by the resonator to be maximised:

$$P = P_0(\omega_0) = \frac{1}{2} \frac{|V_g|^2}{R} \quad \text{or} \quad P = P_0(\omega_0) = \frac{1}{2} \frac{|I_g|^2}{G}. \quad (\text{C.3})$$

At all other frequencies, the resonator absorbs less power. For the resonator to absorb a fraction $0 < p < 1$ of its maximum power P , the combination of (C.2) and (C.3) yields:

$$|Z_L|^2 = R^2 / p \quad \text{and} \quad |Y_L|^2 = G^2 / p \quad (\text{C.4a})$$

$$\text{or} \quad X^2 = R^2 \frac{1-p}{p} \quad \text{and} \quad B^2 = G^2 \frac{1-p}{p}. \quad (\text{C.4b})$$

As long as the required frequency is near resonance, the condition of (C.4a), when substituted in (C.1), yields:

$$\begin{aligned}
 |R + jRQ\delta(p)|^2 &= R^2 / p \\
 1 + Q^2\delta(p)^2 &= 1/p \\
 Q &= \frac{1}{\delta(p)} \sqrt{\frac{1-p}{p}}
 \end{aligned}
 \tag{C.5}$$

for both resonators, where $\delta(p)$ is the bandwidth at the power fraction p . Equation (C.5) applies to *all* definable quality factors, be it the loaded, unloaded, or external Q . As will be shown later, only the measured bandwidth $d(p)$ changes. The form of (C.5) allows one to speak of a “ p_{dB} -dB bandwidth” unambiguously, just as one currently speaks of a “3-dB bandwidth” in the classical context of bandwidth measurements. Expressions such as “the 20-dB unloaded bandwidth” and “the 7-dB external bandwidth” therefore have clear meanings. Usually, the bandwidth is measured at half-power, $p = 0.5$, which by (C.5) results in $Q = 1/\delta$, the classical result. However, (C.5) is less restrictive and indicates that the bandwidth may be measured at any power ratio p , not just at half-power. The only restrictions to using (C.5) are that the bandwidth must be sufficiently small such that the approximations of (C.1) still apply, and that the actual resonant circuit is still sufficiently well modelled by an RLC circuit at the frequencies corresponding to the bandwidth δ . Note how (C.5) can be used repeatedly at different p levels, permitting an average Q value to be obtained from a single measurement. Before using (C.5), one must know the levels at which to measure the bandwidth, and these are derived next, first for a two-port resonator, then for a one-port.

8.3.1 Bandwidth levels of a two-port resonator

Two-port resonators may adopt any one of the *four* possible configurations shown in Fig. C.2.

Fig. C.2 Two-port resonant circuits; Transmission: (a) series-connected series resonator, (b) parallel-connected parallel resonator; Absorption: (c) series-connected parallel resonator, (d) parallel-connected series resonator.

Resonators (a) and (b) are *transmission* resonators whereas resonators (c) and (d) are *absorption* resonators. Key to understanding the behaviour of resonant circuits is that the LC tank acts as an oscillator when it is excited at its resonant frequency. In the absence of exciters, such as the port generator V_g in Fig. C.2, the finite energy stored in the tank will be gradually dissipated by the resistors that load the circuit. The resistor R is inherent to the unloaded resonator whereas the port impedance values of Z_0 are external to it. The ratio of the power dissipated by the external loads to that dissipated by the unloaded resonator P_E/P_0 is known as the coupling coefficient κ . Each of the four resonators in Fig. C.2 has a different coupling coefficient that can be easily identified if the circuits are redrawn by suppressing the V_g generator and replacing the LC tanks by either their

Norton or Thévenin equivalent sources, as is done in Fig. C.3. The figure also lists each resonator's coupling coefficient.

Fig. C.3 *Equivalent circuits as viewed by the LC tanks and their corresponding coupling coefficient κ .*

The coupling coefficient is an indispensable ratio in the analysis of resonators and it is straightforward to show that the loaded and external quality factors are related to the unloaded quality factor via [13]:

$$Q_L = \frac{Q_0}{1 + \kappa} \quad \text{and} \quad Q_E = Q_0 / \kappa \quad (\text{C.6})$$

hence if κ is known, one bandwidth measurement suffices for determining all three quality factors.

The next step is to express the scattering parameters of the circuits of Fig. C.2. Beginning with the reflection coefficient, the input impedance looking into port 1 of circuits (a) and (c) is a series connection of Z_L and Z_0 , where Z_L represents the impedance of the unloaded resonator. As such, the reflection coefficient of these circuits is

$s_{11} = Z_L/(Z_L + 2Z_0)$. The input impedance of circuits (b) and (d) consists of a parallel connection and the form of the reflection coefficient is therefore $s_{11} = -Z_0/(2Z_L + Z_0)$. At resonance, $Z_L = R$ for all four resonators, leading to the following reflection coefficients at resonance, denoted by $s_{11\ 0}$:

$$s_{11\ 0} = \frac{R}{R + 2Z_0} = \frac{1}{1 + \kappa} \quad (\text{C.7a})$$

$$s_{11\ 0} = \frac{-Z_0}{2R + Z_0} = \frac{-1}{1 + \kappa} \quad (\text{C.7b})$$

$$s_{11\ 0} = \frac{R}{R + 2Z_0} = \frac{1}{1 + 1/\kappa} \quad (\text{C.7c})$$

$$s_{11\ 0} = \frac{-Z_0}{2R + Z_0} = \frac{-1}{1 + 1/\kappa} \quad (\text{C.7d})$$

where the suffix i to (C.7*i*) indicates to which circuit of Fig. C.2 it applies, where the corresponding value of κ shown in Fig. C.3 must also be used. Note how the magnitude of (C.7a) and (C.7b) is the same for both transmission resonators, and the same is true for both absorption resonators (C.7c) and (C.7d). Hence the coupling coefficients may be found for both types of resonators using the following:

$$\text{Transmission resonators: } \kappa = \frac{1 - |s_{11\ 0}|}{|s_{11\ 0}|} \quad (\text{C.8a})$$

$$\text{Absorption resonators: } \kappa = \frac{|s_{11\ 0}|}{1 - |s_{11\ 0}|} \quad (\text{C.8b})$$

The transmission coefficients s_{21} may be found easily by noting that the circuits of Fig. C.2(a) and (c) consist of a simple series connected lumped-element circuit, so

$s_{21} = 1 - s_{11}$. Similarly, circuits (b) and (d) are parallel connections, so $s_{21} = 1 + s_{11}$ [14].

Using (C.7) shows that s_{210} is positive definite, leading to the following:

$$\text{Transmission resonators: } s_{210} = \frac{\kappa}{1 + \kappa} \Rightarrow \kappa = \frac{s_{210}}{1 - s_{210}} \quad (\text{C.9a})$$

$$\text{Absorption resonators: } s_{210} = \frac{1/\kappa}{1 + 1/\kappa} \Rightarrow \kappa = \frac{1 - s_{210}}{s_{210}} \quad (\text{C.9b})$$

The bandwidth of the resonators will be evaluated by measuring either a reflection or a transmission magnitude. As such, it is worthwhile to express the magnitude squared values of the reflection and transmission magnitudes. Consider first the transmission resonators. The unloaded resonator of Fig. C.2(a) is series and should be expressed as $Z_L = R + jX$, whereas that of Fig. C.2(b) is parallel and is expressed by $1/Z_L = Y_L = G + jB$. From (C.7a) and (C.7b),

$$|s_{11}|^2 = \frac{|Z_L|^2}{|Z_L + 2Z_0|^2} = \frac{R^2 + X^2}{(R + 2Z_0)^2 + X^2} \quad (\text{C.10a})$$

$$|s_{11}|^2 = \left| \frac{-1/Y_0}{2/Y_L + 1/Y_0} \right|^2 = \frac{|Y_L|^2}{|Y_L + 2Y_0|^2} = \frac{G^2 + B^2}{(G + 2Y_0)^2 + B^2} \quad (\text{C.10b})$$

Note how the two equations have the same form. Later, appropriate values for X and B will be substituted into (C.10) to find the bandwidths, and as shown in (C.4b), these values also have the same form. The coupling coefficients also have the same form, being $\kappa = 2Z_0/R$ for type (a) and $\kappa = 2Y_0/G$ for type (b). Hence when expressed as functions of κ , (C.10a) will be identical to (C.10b). The same thing happens if one derives the transmission coefficients. This means that *either* (C.10a) or (C.10b) may be chosen for the subsequent derivation of the bandwidth for transmission resonators: both will yield the same result. As for the absorption resonators, the resonator of Fig. C.2(c) is parallel

while that of (d) is series, leading to the following two equations for the magnitude squared reflection coefficients, respectively,

$$|s_{11}|^2 = \left| \frac{1/Y_L}{1/Y_L + 2/Y_0} \right|^2 = \frac{Y_0^2}{|Y_0 + 2Y_L|^2} = \frac{Y_0^2}{(2G + Y_0)^2 + 4B^2} \quad (\text{C.11a})$$

$$|s_{11}|^2 = \frac{Z_0^2}{|2Z_L + Z_0|^2} = \frac{Z_0^2}{(2R + Z_0)^2 + 4X^2} \quad (\text{C.11b})$$

Again, the form of (C.11a) is identical to (C.11b). The reactance/susceptance substitutions will have the same form when the bandwidth is evaluated, and the coupling coefficients also have the same form, so only one of the two absorption resonators needs to be further derived: the result will apply equally to the other absorption resonator.

The transmission resonator of Fig. C.2(a) and the absorption resonator of Fig. C.2(c) will be used to derive the bandwidths. The procedure is the same in both cases: first, the appropriate unloaded, loaded, or external reactance/susceptance, given by (C.4b), is substituted into (C.11) [15]. Next, the equations are normalised with respect to κ using the values found in Fig. C.3. Finally, replacing κ with (C.8) or (C.9) produces expressions that are only a function of the s-parameters at resonance. Beginning with the unloaded case (ignoring the loading), the reactance condition for the series resonator (a) is $X^2 = R^2(1 - p)/p$ whereas that of the parallel resonator (c) is $B^2 = G^2(1 - p)/p$. The unloaded p -bandwidth levels are given by:

Unloaded – Transmission

$$|s_{11}|^2 = \frac{1}{p\kappa^2 + 2p\kappa + 1} = \frac{s_{110}^2}{(1 - p)s_{110}^2 + p} \quad (\text{C.12a})$$

$$|s_{21\ u}|^2 = \frac{p\kappa^2}{p\kappa^2 + 2p\kappa + 1} = \frac{ps_{21\ 0}^2}{(1-p)s_{21\ 0}^2 + 2(p-1)s_{21\ 0} + 1} \quad (\text{C.12b})$$

Unloaded – Absorption

$$|s_{11\ u}|^2 = \frac{p\kappa^2}{p\kappa^2 + 2p\kappa + 1} = \frac{ps_{11\ 0}^2}{(1-p)s_{11\ 0}^2 + 2(p-1)s_{11\ 0} + 1} \quad (\text{C.12c})$$

$$|s_{21\ u}|^2 = \frac{1}{p\kappa^2 + 2p\kappa + 1} = \frac{s_{21\ 0}^2}{(1-p)s_{21\ 0}^2 + p} \quad (\text{C.12d})$$

Note the symmetry: the absorption equations are identical to the transmission equations by interchanging s_{11} and s_{21} . For the external bandwidth levels, one considers the resistance of resonator (a) as comprising only of $2Z_0$ or the conductance of resonator (c) as being $Y_0/2$, yielding the conditions $X^2 = 4Z_0^2(1-p)/p$ and $B^2 = (Y_0/2)^2(1-p)/p$, respectively. The external p -bandwidth levels are thereby given by:

External – Transmission

$$|s_{11\ e}|^2 = \frac{(1-p)\kappa^2 + p}{\kappa^2 + 2p\kappa + p} = \frac{s_{11\ 0}^2 + 2(p-1)s_{11\ 0} + (1-p)}{(1-p)s_{11\ 0}^2 + 2(p-1)s_{11\ 0} + 1} \quad (\text{C.13a})$$

$$|s_{21\ e}|^2 = \frac{p\kappa^2}{\kappa^2 + 2p\kappa + p} = \frac{ps_{21\ 0}^2}{(1-p)s_{21\ 0}^2 + p} \quad (\text{C.13b})$$

External – Absorption

$$|s_{11\ e}|^2 = \frac{p\kappa^2}{\kappa^2 + 2p\kappa + p} = \frac{ps_{11\ 0}^2}{(1-p)s_{11\ 0}^2 + p} \quad (\text{C.13c})$$

$$|s_{21\ e}|^2 = \frac{(1-p)\kappa^2 + p}{\kappa^2 + 2p\kappa + p} = \frac{s_{21\ 0}^2 + 2(p-1)s_{21\ 0} + (1-p)}{(1-p)s_{21\ 0}^2 + 2(p-1)s_{21\ 0} + 1} \quad (\text{C.13d})$$

For the loaded points, the series resonator is viewed as containing a total resistance of $R + 2Z_0$ and the parallel one as containing a total conductance of $G + Y_0/2$, yielding the final conditions of $X^2 = (R+2Z_0)^2(1-p)/p$ and $G^2 = (G+Y_0/2)^2(1-p)/p$, leading to the p -bandwidth loaded levels:

Loaded – Transmission

$$|s_{11\ L}|^2 = \frac{p}{K^2 + 2K^2 + 1} + (1-p) = ps_{11\ 0}^2 + (1-p) \quad (C.14a)$$

$$|s_{21\ L}|^2 = \frac{pK^2}{K^2 + 2K + 1} = ps_{21\ 0}^2 \quad (C.14b)$$

Loaded – Absorption

$$|s_{11\ L}|^2 = \frac{pK^2}{K^2 + 2K + 1} = ps_{11\ 0}^2 \quad (C.14c)$$

$$|s_{21\ L}|^2 = \frac{p}{K^2 + 2K^2 + 1} + (1-p) = ps_{21\ 0}^2 + (1-p) \quad (C.14d)$$

The procedure to using (C.12)–(C.14) is the following: measure the resonant s-parameter levels, choose a fractional power level p , measure the double-sided fractional bandwidth δ at the levels given by (C.12)–(C.14), then use (C.5) to convert the bandwidth to the desired quality factor. Equation (C.6) may also be used to convert between the quality factors once one of them has been measured. It is of interest to note that the absorption resonator levels in (C.12)–(C.14) reduce to those given in [16] when the half-power point is chosen, $p = 0.5$.

8.3.2 Bandwidth levels of a one-port resonator

The analysis of the one-port, or *reflection* resonator, is performed in the same manner as that of the two-port resonator. The equivalent circuit of a series one-port is identical to that of Fig. C.3(a) except that the loading is reduced to Z_0 instead of $2Z_0$ and so the series coupling coefficient is $\kappa_s = Z_0/R$. The equivalent circuit of a parallel reflection resonator is identical to Fig. C.3(b) except that the loading is again Z_0 instead of $2Z_0$ and the coupling coefficient is $\kappa_p = R/Z_0$. The reflection coefficient may be used to express the coupling coefficients at resonance, where the impedance looking into both resonators is equal to R :

$$\Gamma_0 = \frac{R - Z_0}{R + Z_0} = \frac{1 - \kappa_s}{1 + \kappa_s} = -\frac{1 - \kappa_p}{1 + \kappa_p} \quad (\text{C.15a})$$

$$\kappa_s = \frac{1 - \Gamma_0}{1 + \Gamma_0} \quad \kappa_p = \frac{1 + \Gamma_0}{1 - \Gamma_0} \quad (\text{C.15b})$$

Note that there is a sign ambiguity to Γ_0 : it will be positive if the resonator is undercoupled-series or overcoupled-parallel, and negative if the resonator is overcoupled-series or undercoupled-parallel. Equation (C.15b) may be expressed as (C.16), where only the magnitude of Γ_0 is now used and no distinction between series and parallel resonators needs to be made,

$$K_{\text{over}} = \frac{1 + |\Gamma_0|}{1 - |\Gamma_0|} \quad (\text{C.16a})$$

$$K_{\text{under}} = \frac{1 - |\Gamma_0|}{1 + |\Gamma_0|} \quad (\text{C.16b})$$

where (C.16a) should be used if the resonator is over-coupled and (C.16b) is it is under-coupled. These two conditions are easily recognised on the Smith Chart: if the resonance

appears to encircle the centre of the Smith Chart, it is overcoupled, otherwise it is undercoupled [15]. Hence the appropriate equation to use in (C.16) may be easily determined by visual inspection. The magnitude-squared reflection coefficient, expressed in terms of impedance for the series resonator and in terms of admittance for the parallel resonator, is given by:

$$|\Gamma|^2 = \frac{|Z_L - Z_0|^2}{|Z_L + Z_0|^2} = \frac{(R - Z_0)^2 + X^2}{(R + Z_0)^2 + X^2} \quad (\text{C.17a})$$

$$|\Gamma|^2 = \frac{|Y_0 - Y_L|^2}{|Y_L + Y_0|^2} = \frac{|(-1)(Y_L - Y_0)|^2}{|Y_L + Y_0|^2} = \frac{(G - Y_0)^2 + B^2}{(G + Y_0)^2 + B^2} \quad (\text{C.17b})$$

Both equations again have the same form, so only the series case need be considered since the resulting equations will also be applicable to the parallel case.

The same procedure is used to evaluate the three p -bandwidths by using (C.17a) at the following conditions: unloaded $X^2 = R^2(1 - p)/p$, external $X^2 = Z_0^2(1 - p)/p$ and loaded $X^2 = (R + Z_0)^2(1 - p)/p$, resulting in the following bandwidth levels:

$$|\Gamma_U|^2 = \frac{p\kappa^2 - 2p\kappa + 1}{p\kappa^2 + 2p\kappa + 1} = \frac{(1 - p)(1 \pm |\Gamma_0|)^2 + 4p\Gamma_0^2}{(1 - p)(1 \pm |\Gamma_0|)^2 + 4p} \quad (\text{C.18})$$

$$|\Gamma_E|^2 = \frac{\kappa^2 - 2\kappa p + p}{\kappa^2 + 2\kappa p + p} = \frac{(1 - p)(1 \mp |\Gamma_0|)^2 + 4p\Gamma_0^2}{(1 - p)(1 \mp |\Gamma_0|)^2 + 4p} \quad (\text{C.19})$$

$$|\Gamma_L|^2 = \frac{\kappa^2 + 2\kappa(1 - 2p) + 1}{\kappa^2 + 2\kappa + 1} = 1 - p + p\Gamma_0^2 \quad (\text{C.20})$$

If the resonator is undercoupled, the top sign should be used, whereas if it is overcoupled, the bottom sign should be used. The unloaded levels are the easiest to evaluate since the sign of Γ_0 is irrelevant. These equations reduce to those given by [17] at half-power, $p = 0.5$.

8.3.3 Summary

Simple equations have been presented that permit the unloaded, loaded, and external quality factors to be determined from arbitrary bandwidth measurements. The required formulae have been presented for all plausible forms of single-pole one- and two-port resonators and they are easy to use. One need only know the type of resonator being measured, which is easily discernible. An s -parameter measurement at resonance then suffices to determine the level at which the p -bandwidth should be measured.

8.3.4 Appendix C references

- [1] M. C. Sanchez, E. Martin, and J. M. Zamarro, "Unified and simplified treatment of techniques for characterising transmission, reflection or absorption resonators", *Proc. Inst. Elect. Eng.*, Part H, vol. 137, pp. 209–212, Aug. 1990.
- [2] Z. Wu & L. E. Davis "Automation-oriented techniques for quality factor measurements of high- T_c superconducting resonators", *Proc. Inst. Elect. Eng.*, Part H, vol. 141, pp. 527–530, 1994.
- [3] Z. Wu & L. E. Davis, "Comments on 'Characterization of high- Q resonators for microwave-filter applications'", *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1751–1752, Sept. 1999.
- [4] R. S. Kwok & J.-F. Liang, "Characterization of high- Q resonators for microwave-filter applications", *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 111–114., Jan. 1999.
- [5] E.-Y. Sun & S.-H. Chao, "Unloaded Q measurement—the critical-points method", *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 1983–1986, Aug. 1995.
- [6] D. Kajfez & E. J. Hwan, " Q -Factor measurement with network analyzer", *IEEE Trans. Microwave Theory Tech.*, vol. 32, pp. 666–670, July 1984.
- [7] J. M. Drozd & W. T. Joines, "Determining Q using S parameter data", *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 2123–2127, Nov. 1996.
- [8] K. Leong & J. Mazierska, "Precise measurements of the Q factor of dielectric resonators in the transmission mode—accounting for noise, crosstalk, delay of uncalibrated lines, coupling loss, and coupling reactance", *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 2115–2127, Sept. 2002.
- [9] K. Leong, J. Mazierska, and J. Krupka, "Measurements of unloaded Q -factor of transmission mode dielectric resonators", in *IEEE MTT-S Int. Symp. Dig.*, vol. 3, 1997, pp. 1639–1642.
- [10] D. Kajfez, *Q Factor*, Oxford, MS: Vector Fields, 1994.
- [11] D. Kajfez, " Q -factor measurement with a scalar network analyser", *Proc. Inst. Elect. Eng.*, Part H, vol. 142, pp. 369–372, Oct. 1995.

- [12] D. M. Pozar, *Microwave Engineering*, Reading, MA: Addison-Wesley, 1990, pp. 330–335.
- [13] R. Beringer, “Resonant cavities as microwave circuit elements,” in *Principles of Microwave Circuits*, C. G. Montgomery, R. H. Dicke, and E. M. Purcell, Eds., *MIT Radiation Laboratory Series*, vol. 8, chap. 7, New York: McGraw-Hill, 1948, pp. 207–238.
- [14] R. E. Collin, *Foundations for Microwave Engineering*, 2nd ed., New York: McGraw-Hill, 1992, pp. 254–257.
- [15] C. G. Montgomery, *Technique of Microwave Measurements*, Lexington, MA: Boston Technical Pub., 1964, pp. 330–342.
- [16] A. Khanna & Y. Garault, “Determination of loaded, unloaded, and external quality factors of a dielectric resonator coupled to a microstrip line”, *IEEE Trans. Microwave Theory Tech.*, vol. 31, pp. 261–264, March 1983.
- [17] J. E. Aitken, “Swept-frequency microwave Q factor measurements with network analyzer”, *Proc. Inst. Elect. Eng.*, vol. 123, pp. 855–862, Sept. 1976.

This concludes Appendix C.