**Mitigating the effects of cavity resonance
in ceramic millimetre-wave packages**

by

William J. D. Sitch, B.Eng.

A thesis submitted to the Faculty of Graduate Studies and Research

In partial fulfillment of the requirements for the degree of

Master of Applied Science

Ottawa-Carleton Institute of Electrical and Computer Engineering

Department of Electronics

Carleton University

Ottawa, Ontario, Canada

January 2002

© Copyright 2002, William J. D. Sitch

National Library
of Canada

Acquisitions and
Bibliographic Services

395 Wellington Street
Ottawa ON K1A 0N4
Canada

Bibliothèque nationale
du Canada

Acquisitions et
services bibliographiques

395, rue Wellington
Ottawa ON K1A 0N4
Canada

Your file Votre référence

Our file Notre référence

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.

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 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.

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.

0-612-79718-X

Canada

The undersigned hereby recommends to the Faculty of Graduate Studies
and Research the acceptance of the thesis

**Mitigating the effects of cavity resonance
in ceramic millimetre-wave packages**

Submitted by William Sitch

In partial fulfillment of the requirements for the
Degree of Master of Applied Science

Thesis co-supervisor, Dr. J. S. Wight

Thesis co-supervisor, Dr. H. Kobeissi

Chair, Dept. of Electronics, Dr. M. S. Nakhla

Carleton University

January 2002

Abstract

This thesis presents an evaluation of practical resonance mitigation techniques in complex millimetre-wave ceramic packages. The fundamental aspects of cavity resonance are examined and analytic solutions are reviewed. Four resonance mitigation techniques are proposed and developed: modifying package geometry or materials, introducing microwave absorber or resistive films, relocating excitation sources, and disturbing resonance field distributions. Ansoft's HFSS, a three-dimensional finite element method simulator, is used to model the cavity resonance problems. The obtained results are compared to analytic solutions and experimental measurements. The mitigation techniques are evaluated in two packages designed by Nortel Networks' Modules and Advanced Packaging group. The four techniques are considered with emphasis on cost and time effective application in a post-design automated manufacturing process. Modifications are applied to the package models and the effectiveness of each technique is determined. Two of the techniques are physically implemented and both successfully eliminate resonance-induced amplifier oscillations. Simulation and measurement data are summarized and pre- and post-design resonance mitigation techniques are recommended.

Acknowledgements

I am grateful to my thesis supervisor, Dr. Jim Wight, for his guidance and experience. I also greatly appreciate the enthusiastic input and invaluable assistance of Dr. Hatem Akel, Dr. Xiaoming Chen, Dr. Fei Guo, Dr. Hassan Kobeissi, Dr. Tibor Lapohos, and Dr. John Sitch from Nortel Networks. In addition, I would like to thank Natasha Artemeva, Heather Smith, and Christine Sulek for their proofreading and encouragement. I thankfully acknowledge the financial support provided by the Natural Sciences and Engineering Research Council of Canada, Carleton University, and Nortel Networks.

This thesis would not have been possible without the research agreement between Carleton University and the Modules and Advanced Packaging group of Nortel Network's Optical Components division. I would like to thank John Pokinko and Jose Descoteaux for their exceptional contribution towards my research. Finally, I would like to thank my friends and family, my adventure race team, and my fourth-year project students for their continual distraction, encouragement and support.

Table of Contents

List of Figures	viii
List of Tables	xiv
Chapter 1 – Introduction.....	1
1.1 Research Motivation.....	7
1.2 Research Objectives	9
1.3 Thesis Organization.....	10
Chapter 2 – Theory and Analysis of Cavity Resonance	12
2.1 Derivation of Analytic Solutions to Resonant Mode Frequencies in Inhomogeneous Rectangular Cavities.....	13
2.2 Theory of Resonant Modes	29
2.3 Practical Techniques to Mitigate Resonance Effects.....	31
2.3.1 Modifying Package Geometry or Materials	34
2.3.2 Introducing Microwave Absorber or Resistive Films	36
2.3.3 Modifying Resonance Excitation Sources	38
2.3.4 Disturbing Resonance Field Patterns.....	39
Chapter 3 – Finite Element Method Simulations	42
3.1 The Finite Element Method	42

3.1.1	FEM Limitations and Considerations.....	45
3.2	Ansoft's High Frequency Structure Simulator	48
3.2.1	Eigenmode Simulations	50
3.2.2	Driven Simulations	52
3.3	Model Approximations	58
Chapter 4 – Case Studies		62
4.1	K/Ka-Band Power Amplifier Package	65
4.1.1	Package Simulations.....	71
4.1.2	Package Measurements	76
4.1.3	Modifying Package Geometry to Decrease the Effective Volume	82
4.1.4	Introducing Absorber to Reduce the Cavity Quality Factor	90
4.1.5	Moving Excitation Sources to Reduce Resonance Coupling	92
4.1.6	Disturbing Field Patterns to Decrease Coupling	96
4.1.7	Summary of Results.....	104
4.2	Broadband Two-Stage Amplifier Package.....	106
4.2.1	Package Simulations.....	109
4.2.2	Package Measurements	113
4.2.3	Disturbing Field Patterns to Eliminate Amplifier Oscillations	114
4.2.4	Modifying Package Geometry to Partition the Cavity	116
4.2.5	Introducing Absorber to Reduce the Cavity Quality Factor	124
4.2.6	Summary of Results.....	131

Chapter 5 – Conclusion.....	133
5.1 Post-Design Mitigation Techniques	135
5.2 Pre-Design Mitigation Techniques.....	138
5.3 Research Limitations	140
5.4 Recommendations and Future Work.....	141
References	143

List of Figures

Figure 1.1 – Exploded view of the K/Ka-band power amplifier package	2
Figure 1.2 – Rectangular box with a dielectric slab.....	5
Figure 1.3 – HFSS eigenmode electric field pattern simulation results of the TE_{101} dominant resonance mode in a rectangular box with a dielectric slab	6
Figure 2.1 – Two coordinate systems and different modal representation.....	13
Figure 2.2 – Geometry of a rectangular waveguide	18
Figure 2.3 – Electromagnetic field patterns for modes in a rectangular waveguide	21
Figure 2.4 – Geometry of a rectangular cavity	22
Figure 2.5 – Field patterns for a) TE, and b) TM modes in a rectangular cavity .	24
Figure 2.6 – Rectangular box without a dielectric slab.....	25
Figure 2.7 – Rectangular box with a dielectric slab similar to the K/Ka-band power amplifier package	27
Figure 2.8 – Rectangular box with a dielectric slab similar to the broadband two- stage amplifier package.....	28
Figure 3.1 – HFSS eigenmode electric field pattern simulation results of the TE_{101} dominant mode in a rectangular box a) without a dielectric slab, b) with a dielectric slab	51
Figure 3.2 – Rectangular box with a dielectric slab, two ports, and two unconnected microstrip lines	54

Figure 3.3 – a) Full, and b) partial HFSS driven 2-port s-parameter simulation results of a rectangular box with a dielectric slab, two ports, and two un-connected microstrip lines	56
Figure 3.4 – Curved surface modeling approximations.....	60
Figure 4.1 – Exploded view of the broadband two-stage amplifier package	64
Figure 4.2 – K/Ka-band power amplifier package (measurement comparison model).....	66
Figure 4.3 – K/Ka-band power amplifier package (simplified model)	67
Figure 4.4 – HFSS driven 2-port s-parameter simulation results for a) full package model, and b) simulation model of the K/Ka-band power amplifier package	68
Figure 4.5 – HFSS driven 2-port s-parameter loss for a) full package model, and b) simulation model of the K/Ka-band power amplifier package	70
Figure 4.6 – HFSS eigenmode electric field pattern simulation results of the a) first, b) second, and c) third resonant mode in the K/Ka-band power amplifier package	72
Figure 4.7 – HFSS driven 2-port s-parameter simulation results of the TE_{102} and TE_{201} resonant modes in the K/Ka-band power amplifier package	75
Figure 4.8 – Anritsu universal test fixture, V connectors, and the K/Ka-band power amplifier package.....	77
Figure 4.9 – Measured a) S_{21} & S_{12} , and b) S_{11} & S_{22} for the 1.00cm calibration standard	79

Figure 4.10 – Measurements and HFSS eigenmode simulation results of the a) S11 & S22, and b) S21 & S12 for the K/Ka-band power amplifier package.	81
Figure 4.11 – K/Ka-band power amplifier package with four grounded posts in the corners.....	83
Figure 4.12 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-Band power amplifier package with four grounded posts in the corners	83
Figure 4.13 – HFSS driven 2-port s-parameter simulation results of the K/Ka- band power amplifier package with four grounded posts in the corners	85
Figure 4.14 – K/Ka-band power amplifier package with four grounded posts in the middle of each side.....	86
Figure 4.15 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with four grounded posts in the middle of each side	86
Figure 4.16 – HFSS driven 2-port s-parameter simulation results of the K/Ka- band power amplifier package with four grounded posts in the middle of each side	88
Figure 4.17 – K/Ka-band power amplifier package with ECCOSORB® GDS absorber under the lid.....	91
Figure 4.18 – K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates	93

Figure 4.19 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates	94
Figure 4.20 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates	95
Figure 4.21 – K/Ka-band power amplifier package with cylindrical protrusion under the lid	97
Figure 4.22 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with cylindrical protrusion under the lid	98
Figure 4.23 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with cylindrical protrusion under the lid	99
Figure 4.24 – K/Ka-band power amplifier package with diagonal protrusion under the lid	100
Figure 4.25 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with diagonal protrusion under the lid	101
Figure 4.26 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with diagonal protrusion under the lid.....	102
Figure 4.27 – Broadband two-stage amplifier package (partially pre-conditioned model).....	107
Figure 4.28 – Broadband two-stage amplifier package (simplified model).....	108

Figure 4.29 – HFSS eigenmode electric field pattern simulation results of the a) first and b) second resonant modes in the broadband two-stage amplifier package	110
Figure 4.30 – HFSS eigenmode surface current simulation results of the first resonant mode in the broadband two-stage amplifier package	112
Figure 4.31 – Broadband two-stage amplifier package with lid protrusion	115
Figure 4.32 – Broadband two-stage amplifier package with one grounded post	117
Figure 4.33 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with one grounded post.....	117
Figure 4.34 – Broadband two-stage amplifier package with two grounded posts partitioning the cavity	119
Figure 4.35 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with two grounded posts partitioning the cavity	120
Figure 4.36 – Broadband two-stage amplifier package with two partially grounded posts partitioning the cavity	122
Figure 4.37 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with two partially grounded posts.....	123
Figure 4.38 – Broadband two-stage amplifier package with ECCOSORB® GDS absorber (43mm ³) under the lid	125

Figure 4.39 – Broadband two-stage amplifier package with a ring of ECCOSORB® GDS absorber (19.63mm^3) under the lid	127
Figure 4.40 – Broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber (2.04mm^3) under the lid	129
Figure 4.41 – HFSS eigenmode surface current simulation results of the first resonant mode in the broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber under the lid	130

List of Tables

Table 1.1 – Parameters of a rectangular box with a dielectric slab	5
Table 1.2 – HFSS eigenmode simulation results of the first three resonant modes in a rectangular box with a dielectric slab	7
Table 2.1 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular box without a dielectric slab	25
Table 2.2 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular box with a dielectric slab similar to the K/Ka-band power amplifier package	27
Table 2.3 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular cavity similar to the broadband two-stage amplifier package	28
Table 3.1 – HFSS eigenmode simulation results and analytic solutions of the TE_{101} dominant mode in a rectangular box with and without a dielectric slab	52
Table 3.2 – HFSS eigenmode simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two unconnected microstrip lines	54
Table 3.3 – HFSS driven 2-port simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two un- connected microstrip lines	57

Table 3.4 – HFSS eigenmode and driven 2-port simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two un-connected microstrip lines.....	57
Table 4.1 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-Band power amplifier package.....	73
Table 4.2 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package	75
Table 4.3 – Comparison of analytic solution, HFSS eigenmode simulation results, and HFSS 2-port driven simulation results for the K/Ka-band power amplifier package.....	76
Table 4.4 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the corners.....	84
Table 4.5 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the corners.....	85
Table 4.6 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the middle of each side.....	87
Table 4.7 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the middle of each side	89
Table 4.8 – ECCOSORB® GDS material properties	90

Table 4.9 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with ECCOSORB® GDS absorber under the lid.....	91
Table 4.10 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates	94
Table 4.11 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates	96
Table 4.12 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with cylindrical protrusion under the lid.....	98
Table 4.13 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with cylindrical protrusion under the lid.....	99
Table 4.14 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with diagonal protrusion under the lid.....	101
Table 4.15 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with diagonal protrusion under the lid.....	102

Table 4.16 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with various modifications	104
Table 4.17 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with various modifications	105
Table 4.18 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package	111
Table 4.19 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package without microstrip and DC feed lines	111
Table 4.20 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with one grounded post	118
Table 4.21 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two grounded posts partitioning the cavity	121
Table 4.22 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two partially grounded posts	124
Table 4.23 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with ECCOSORB® GDS absorber (43mm ³) under the lid	125

Table 4.24 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with a ring of ECCOSORB® GDS absorber (19.63mm^3) under the lid	127
Table 4.25 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with a narrow ring of ECCOSORB® GDS absorber (9.73mm^3) under the lid	128
Table 4.26 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber (2.04mm^3) under the lid	129
Table 4.27 – HFSS eigenmode quality factor simulation results of the first two resonant modes in the broadband two-stage amplifier package with different amounts of ECCOSORB® GDS absorber under the lid	130
Table 4.28 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with various modifications.	132

Chapter 1 – Introduction

Complex microwave and millimetre-wave systems usually consist of multi-chip modules containing one or more packages. Each package houses one or more monolithic microwave integrated circuits (MMICs). The package is a protective structure that provides the enclosed MMICs with mechanical support, power routing, signal distribution, heat transfer, thermal dissipation, and environmental and electromagnetic protection. Commercially viable packages must be compact, low cost, and suitable for high volume automated manufacturing. Furthermore, packages should have an operating lifetime greater than the devices they protect and must be as electrically transparent to the die as possible [1] – [5].

The above-mentioned requirements complicate package design and fabrication. Not only must packages incorporate a grounded Faraday cage for electromagnetic shielding, but they must also be hermetically sealed to protect the enclosed die from corrosion and other environmental effects. These implications can decrease the electrical transparency of the package. High frequency packages must be carefully designed to minimize detrimental package-related parasitic effects that degrade the device performance.

Figure 1.1 shows an exploded view of the K/Ka-band power amplifier package that meets the design requirements and is examined in Section 4.1.

Figure 1.1 – Exploded view of the K/Ka-band power amplifier package

Figure 1.1 is a hermetic package similar to those described in [1] – [4]. The copper-tungsten pedestal and heat spreader provide thermal dissipation, heat transfer, and a solid electrical ground on which the MMIC is mounted. The coefficient of thermal expansion (CTE) for copper-tungsten is compatible with the CTE for MMIC gallium arsenide (GaAs) die. High-temperature cofired ceramic is used for the substrate and walls of the package due to the ease of manufacturing associated with well-known screen printing techniques and multilayer ceramic lamination [6]. Kovar™, an iron-nickel-cobalt alloy, is used for the package seal ring and lid due to its low cost and CTE compatibility with HTCC [1].

The microstrip lines, tuning pads, and DC feed lines are comprised of a tungsten, nickel, and gold stack. Tungsten vias in the HTCC walls connect the grounded pedestal and heat spreader to the conductive Kovar™ seal ring and lid, creating a grounded cage around the cavity.

The type of packaging mentioned above is emerging as a relatively new technology driven by high-frequency MMIC design and an increasing level of component integration. As wavelengths decrease and cavity sizes increase, packages become electrically large and a number of undesirable parasitic effects degrade their electrical performance. These effects include thru-wall moding and substrate moding, where resonances are excited in the ceramic wall transitions [7] and the dielectric-air interface creates a parallel-plate waveguide with the package ground [8], respectively. Other effects include leaky waves and parallel line coupling [8], [5]. This thesis focuses on the parasitic package effect of cavity resonance, discussed in [2] – [4], and [9].

Cavity resonant modes are standing wave patterns that exist within a conductive enclosure. Oscillating charges force surface currents to flow in the conductive walls of the package. Maxwell's equations and package boundary conditions imply that electromagnetic field patterns are supported in the cavity with a harmonic time variation: resonant modes. The charges and currents are redistributed every quarter cycle by the oscillating modes and extensive grounding of the conductive cage does not affect the resonance.

Resonant modes are labels assigned to certain field distributions that exist at discrete frequencies. A conductive enclosure supports an infinite number of modes that act as energy storage mechanisms. The amount of electromagnetic energy stored within each mode primarily depends on radiation sources within the cavity. These excitation sources must radiate at the resonant frequency of the mode in order to couple energy into that particular resonance. Without excitation, or if the sources do not couple well to the cavity resonances, modes are not excited and contain no energy. Radiation sources exist at every discontinuity and transmission line; each source couples to resonant modes with varying degrees of efficiency [9].

The unloaded cavity quality factor (Q) is an important parameter of a resonant mode within a conductive enclosure. This unit-less factor represents the quality of the cavity materials and geometry as an energy storage mechanism. A cavity with a high Q has low loss and stores energy well; conversely, a low- Q cavity is probably comprised of lossy materials that absorb electromagnetic energy and reduce the effects of resonance.

Figure 1.2 shows a conductive rectangular box with loss-less walls (defined as perfect electric boundaries), an air-filled cavity, and a dielectric slab. The parameters of the box are presented in Table 1.1. This model approximates the K/Ka-band power amplifier package shown in Figure 1.1 and detailed in Section 4.1.

Figure 1.2 – Rectangular box with a dielectric slab

Table 1.1 – Parameters of a rectangular box with a dielectric slab

Package Dimensions (mm)	Substrate Properties
$a = 8.45$; $b = 1.135$; $d = 8.47$	$t = 0.254\text{mm}$; $\epsilon_{r,\text{slab}} = 9.0$; $\tan \delta = 0.002$

Figure 1.3 shows Ansoft's High Frequency Structure Simulator (HFSS) eigenmode field pattern simulation results of the dominant resonance mode within the rectangular box. The vector plot indicates the magnitude and direction

of the electric field distribution of the TE_{101} dominant mode supported at 22.32GHz. The field distribution within the cavity is a standing wave pattern in the x and z directions. The field intensity peaks in the centre of the cavity and diminishes toward the walls where the boundary conditions force zero electric field magnitude.

Figure 1.3 – HFSS eigenmode electric field pattern simulation results of the TE_{101} dominant resonance mode in a rectangular box with a dielectric slab

Table 1.2 lists the HFSS simulation results of the first three resonant modes in the rectangular box. The type, resonant frequency, and cavity quality factor for each mode are shown; note that the quality factor accounts only for the lossy properties of the dielectric slab listed in Table 1.1. The Q values shown do not consider the lossy walls, lid, or other materials that are present in complex packages; accordingly, the values are inappropriately large.

Table 1.2 – HFSS eigenmode simulation results of the first three resonant modes in a rectangular box with a dielectric slab

	Mode type	Resonant frequency (GHz)	Quality factor
First mode	TE ₁₀₁	22.32	11995.7
Second mode	TE ₁₀₂	34.97	8236.3
Third mode	TE ₂₀₁	35.02	8221.9

The second and third resonant modes are supported at similar frequencies due to the comparable length ($a = 8.45\text{mm}$) and width ($d = 8.47\text{mm}$) of the rectangular box. The resonances oscillate in the air-filled cavity above the dielectric slab and generate surface currents that flow through the conductive walls.

1.1 Research Motivation

Cavity resonances can notably degrade the performance of a packaged MMIC even if the device is not operating at resonant frequencies [3], [9]. The detrimental effects of resonance are related to coupling between radiating elements and resonant modes. It is often impossible to reduce the number of radiation sources or to avoid exciting multiple resonant modes due to compact packages and centrally located die.

Coupling between radiation sources and resonant modes can effectively load the source and change its impedance, leading to increased reflections, detuning, and signal power loss. Once a resonance has been excited, further coupling is possible between the mode and other radiators within the package. For example, narrowband coupling to DC feed lines and other signal lines is a

common problem [10]. Feedback through resonant modes can electrically connect portions of the circuit that are not intended to be connected, introducing crosstalk and increased noise. Amplifier oscillations enabled by resonance feedback can reduce the output power and degrade the efficiency and noise figure of the amplifier; blocking can occur when the amplifier oscillation reduces the gain at a specific frequency. As well, peak oscillation voltages can exceed the device breakdown voltage and dramatically reduce the amplifier's lifetime.

Cavity resonance is an important parasitic effect, but this loss mechanism is not well understood. Very little has been published about minimizing the electrical impact of resonances on device performance. In [11], Jackson suggests a theoretical model for approximating the frequency of the dominant resonance mode. The model is accurate only for simple circular or square packages with no internal discontinuities. Jackson does not address higher-order modes or resonance mitigation techniques. Decker suggests using posts or walls to shift the resonance frequencies in [3], but does not disclose any specific information. The package design Decker presents does not include posts or walls.

Chung, in [12], proposes attaching vertical diaphragms to the underside of the lid in order to partition the cavity. Chung concludes that without absorber the cabinets do not affect the field distribution; he predicts that the field leaks out and couples to the resonant modes in the main body of the package. The use of absorber to mitigate the effects of cavity resonance is often recommended in the literature, as seen in [3], [4], and [9] – [12], but the issue of absorber outgassing

is rarely introduced or discussed and few details regarding the absorber type or placement location within the cavity are mentioned.

Both Yook, in [4], and Simons, in [2], propose millimetre-wave packages without significant electromagnetic shielding. These packages leak radiated electromagnetic energy and the authors claim that the structures do not support cavity resonances very well. While this is true, the packages presented are not appropriate for commercial multichip modules; electromagnetic and environmental shielding are critical for the operation of any complex integrated circuit.

1.2 Research Objectives

There is a need for a comprehensive analysis of cavity resonance in complex modern packages. The objective of the research reported in this thesis is the characterization of cavity resonant modes and an evaluation of practical resonance mitigation techniques. This objective is successfully achieved with the five research goals described below.

The first research goal is to examine the fundamental aspects of cavity resonance. An exhaustive literature review reveals that resonances have not been thoroughly explored and that little has been published in regards to minimizing their impact on package performance.

The lack of information relates to the second research goal, which is to propose and develop several techniques of mitigating the effects of cavity

resonance. Emphasis is placed on practical techniques that are cost and time effective and can be implemented in an automated manufacturing process.

The third research goal is to consider a finite element method (FEM) simulator for analysing electromagnetic activity in complex package cavities. This full-wave analysis is required for accurate characterization of electric and magnetic fields and surface currents. Attention is given to simulator limitations and model approximations required for optimization.

The fourth research goal is to evaluate the effectiveness of the proposed mitigation techniques. Each technique is simulated in two package models that support in-band cavity resonances. Simulations are performed to verify resonance theory and rank the success of each technique. Promising techniques are applied to one of the packages to verify simulation results.

The fifth research goal is to recommend pre- and post-design resonance mitigation techniques for use with millimetre-wave ceramic packages. Each technique is summarized with simulation and measurement results and possible applications are suggested.

1.3 Thesis Organization

Chapter 2 presents the theory and analysis of cavity resonance. An analytic solution is derived for resonance frequencies in an inhomogeneous rectangular cavity. Theory of resonant modes is discussed with a focus on developing practical techniques to reduce the effects of resonances. Four of these mitigation techniques are proposed: modifying package geometry or materials,

introducing microwave absorber or resistive films, altering resonance excitation sources, and disturbing resonance field patterns.

Chapter 3 evaluates a three-dimensional full-wave structure simulator used for calculating the electrical behaviour of package models. The finite element method and its inherent limitations are reviewed. Ansoft's HFSS is introduced along with a guide to simulate cavity resonances in packages with large aspect ratios. Model approximations are applied for irrelevant mesh reduction and simulation optimization.

Chapter 4 evaluates the proposed mitigation techniques with two packages designed by the Modules and Advanced Packaging Group of Nortel Networks' Optical Components division. Both packages support in-band resonances predicted by analytic, simulation, and measurement results. The K/Ka-band power amplifier package is used to evaluate all four mitigation techniques with eigenmode and driven simulations. The broadband two-stage amplifier package presents the opportunity to comprehensively examine three techniques with additional emphasis placed on cost, time, and automated manufacturing concerns. Two of these techniques are implemented and both successfully eliminate amplifier oscillations in the broadband two-stage amplifier package.

The thesis concludes with Chapter 5, wherein the previously described mitigation techniques are summarized. Pre- and post-design applications are suggested along with limitations of the research presented. The chapter also includes recommendations for future work.

Chapter 2 – Theory and Analysis of Cavity Resonance

This chapter presents analytic solutions to cavity resonance, a review of resonance theory, and four practical methods of altering cavities to minimize the effects of resonant modes. First, equations for the resonant mode frequencies in inhomogeneous rectangular cavities are derived from Maxwell's equations. Next, the theory of cavity resonance is explored and conclusions are drawn from mathematical models. Finally, four practical techniques to mitigate the effects of cavity resonance are proposed and discussed.

A literature review reveals that coordinate systems for rectangular cavities have no set convention. Since resonant modes do not propagate, the z-axis does not carry the same connotation for rectangular cavities as it does for a rectangular waveguide and can be set in an arbitrary direction. Figure 2.1 shows the electric and magnetic field components of the dominant resonance mode in a rectangular cavity. The modal representation associated with two different coordinate systems is shown.

The coordinate system shown in Figure 2.1a is an extension of the rectangular waveguide standard and has a dominant mode labelled TE_{101} . Figure 2.1b shows an alternate coordinate system with the dominant mode labelled TM_{110} . This thesis uses the coordinate system shown in Figure 2.1a with the dominant resonance mode labelled TE_{101} .

Figure 2.1 – Two coordinate systems and different modal representation

Reprinted with permission from M. S. Gupta, "Fundamentals of RF circuits," Nortel Networks, Ottawa, Canada, Internal Technical Report, Nov. 2000.

2.1 Derivation of Analytic Solutions to Resonant Mode Frequencies in Inhomogeneous Rectangular Cavities

From [13] and [14] a general solution for transverse electric (TE) and transverse magnetic (TM) waves can be derived with Maxwell's equations. The solution, known as the Helmholtz wave equation, is then used to find TE and TM modes in a rectangular waveguide by solving with the separation of variables procedure and applying electric boundary conditions. The TE and TM waveguide fields are then extended to find the cavity guided TE and TM modes by applying two additional electric boundary conditions. Equations are found for resonant mode frequencies in rectangular cavities with and without dielectric slabs.

The general solution for transverse electric and transverse magnetic waves starts with the assumption of time-harmonic fields with an $e^{j\omega t}$ dependence and wave propagation along the z-axis. The electric and magnetic fields are written as

$$\vec{E}(x, y, z) = [\vec{e}(x, y) + \hat{z}e_z(x, y)]e^{-j\beta z}, \quad (2.1a)$$

$$\vec{H}(x, y, z) = [\vec{h}(x, y) + \hat{z}h_z(x, y)]e^{-j\beta z}, \quad (2.1b)$$

where the vectors $\vec{e}(x, y)$ and $\vec{h}(x, y)$ represent the transverse electric and magnetic field components, and e_z and h_z are the longitudinal electric and magnetic field components.

Recall Maxwell's equations in linear media, in phasor form, as

$$\nabla \times \vec{E} = -j\omega\mu \vec{H} - \vec{M}, \quad (2.2a)$$

$$\nabla \times \vec{H} = j\omega\epsilon \vec{E} - \vec{J}, \quad (2.2b)$$

$$\nabla \cdot \vec{D} = \rho, \quad (2.2c)$$

$$\nabla \cdot \vec{B} = 0, \quad (2.2d)$$

where $\vec{E}$ is the electric field strength vector, $\vec{H}$ is the magnetic field strength vector, $\vec{D}$ is the electric flux density vector, $\vec{B}$ is the magnetic flux density vector, $\vec{M}$ is the (fictitious) magnetic current density vector, $\vec{J}$ is the electric current density vector, and ρ is the electric charge density. The region inside the walls is assumed to be source free, so $\rho = \vec{J} = \vec{M} = 0$ and Equation (2.2) reduces to

$$\nabla \times \vec{E} = -j\omega\mu \vec{H}, \quad (2.3a)$$

$$\nabla \times \vec{H} = j\omega\epsilon \vec{E}. \quad (2.3b)$$

With the aforementioned $e^{j\omega z}$ dependence and propagation in the z direction, the x, y, and z components of the above vector equations can be reduced to:

$$\frac{\partial E_z}{\partial y} + j\beta E_y = -j\omega\mu H_x, \quad (2.4a)$$

$$-j\beta E_x - \frac{\partial E_z}{\partial x} = -j\omega\mu H_y, \quad (2.4b)$$

$$\frac{\partial E_y}{\partial x} - \frac{\partial E_x}{\partial y} = -j\omega\mu H_z, \quad (2.4c)$$

$$\frac{\partial H_z}{\partial y} + j\beta H_y = j\omega\epsilon E_x, \quad (2.4d)$$

$$-j\beta H_x - \frac{\partial H_z}{\partial x} = j\omega\epsilon E_y, \quad (2.4e)$$

$$\frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} = j\omega\epsilon E_z. \quad (2.4f)$$

To obtain general transverse field components these equations are solved in terms of E_z and H_z as follows:

$$H_x = \frac{j}{k_c^2} \left(\omega\epsilon \frac{\partial E_z}{\partial y} - \beta \frac{\partial H_z}{\partial x} \right), \quad (2.5a)$$

$$H_y = \frac{-j}{k_c^2} \left(\omega\epsilon \frac{\partial E_z}{\partial x} + \beta \frac{\partial H_z}{\partial y} \right), \quad (2.5b)$$

$$E_x = \frac{-j}{k_c^2} \left(\beta \frac{\partial E_z}{\partial x} + \omega\mu \frac{\partial H_z}{\partial y} \right), \quad (2.5c)$$

$$E_y = \frac{j}{k_c^2} \left(-\beta \frac{\partial E_z}{\partial y} + \omega\mu \frac{\partial H_z}{\partial x} \right), \quad (2.5d)$$

where the cutoff wavenumber is

$$k_c^2 = k^2 - \beta^2, \quad (2.6)$$

and the wavenumber within the waveguide is

$$k = \omega \sqrt{\mu \epsilon} = 2\pi / \lambda. \quad (2.7)$$

Transverse electric waves are characterized by $E_z = 0$ and $H_z \neq 0$. Equation (2.5) then reduces to

$$H_x = \frac{-j\beta}{k_c^2} \frac{\partial H_z}{\partial x}, \quad (2.8a)$$

$$H_y = \frac{-j\beta}{k_c^2} \frac{\partial H_z}{\partial y}, \quad (2.8b)$$

$$E_x = \frac{-j\omega\mu}{k_c^2} \frac{\partial H_z}{\partial y}, \quad (2.8c)$$

$$E_y = \frac{j\omega\mu}{k_c^2} \frac{\partial H_z}{\partial x}. \quad (2.8d)$$

The cutoff wavenumber $k_c \neq 0$, and the propagation constant $\beta = \sqrt{k^2 - k_c^2}$ is generally a function of frequency and the geometry of the transmission medium [13]. To apply Equation (2.8), H_z is found from the Helmholtz wave equation

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} + k^2 \right) H_z = 0, \quad (2.9)$$

which, by making use of $k_c^2 = k^2 - \beta^2$, can be reduced to a two-dimensional wave equation for h_z that is:

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + k_c^2 \right) h_z = 0. \quad (2.10)$$

Similarly, transverse magnetic (TM) waves are characterized by $E_z \neq 0$ and $H_z = 0$. Equations (2.5) then reduce to

$$H_x = \frac{-j\omega\epsilon}{k_c^2} \frac{\partial E_z}{\partial y}, \quad (2.11a)$$

$$H_y = \frac{-j\omega\epsilon}{k_c^2} \frac{\partial E_z}{\partial x}, \quad (2.11b)$$

$$E_x = \frac{-j\beta}{k_c^2} \frac{\partial E_z}{\partial x}, \quad (2.11c)$$

$$E_y = \frac{j\beta}{k_c^2} \frac{\partial E_z}{\partial y}, \quad (2.11d)$$

and E_z is found from the Helmholtz wave equation

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} + k^2 \right) E_z = 0, \quad (2.12)$$

which can be reduced to

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + k_c^2 \right) e_z = 0, \quad (2.13)$$

a two-dimensional wave equation for e_z .

In order to find TE and TM modes in a rectangular waveguide, shown in Figure 2.2, the reduced Helmholtz equations in Equation (2.10) and (2.13) are solved for h_z and e_z . The solution contains several unknown constants and the unknown cutoff wavenumber, k_c . Equations (2.8) and (2.11) solve for the transverse fields from h_z and e_z . Applying the boundary conditions of the waveguide to the appropriate field components yields the unknown constants and cutoff wavenumber k_c [13].

Figure 2.2 – Geometry of a rectangular waveguide

To find the TE mode equations, Equation (2.1a) and (2.6) are used to solve the partial differential equation of Equation (2.10) by letting

$$h_z(x, y) = X(x)Y(y), \quad (2.14)$$

which, after substituting into Equation (2.10), yields

$$\frac{1}{X} \frac{d^2 X}{dx^2} + \frac{1}{Y} \frac{d^2 Y}{dy^2} + k_c^2 = 0. \quad (2.15)$$

Then, using the separation of variables outlined in [13] and [14], a general solution for h_z is found as

$$h_z(x, y) = (A \cos k_x x + B \sin k_x x)(C \cos k_y y + D \sin k_y y). \quad (2.16)$$

Referring to Figure 2.2, boundary conditions on the electric field components tangential to the waveguide walls are applied as follows:

$$e_x(x, y) = 0, \text{ at } y = [0, b], \text{ and} \quad (2.17a)$$

$$e_y(x, y) = 0, \text{ at } x = [0, a]. \quad (2.17b)$$

Then, from Equation (2.8c) and (2.8d), e_x and e_y are found from h_z :

$$e_x = \frac{-j\omega\mu}{k_c^2} k_y (A \cos k_x x + B \sin k_x x) (-C \sin k_y y + D \cos k_y y), \quad (2.18a)$$

$$e_y = \frac{-j\omega\mu}{k_c^2} k_x (-A \sin k_x x + B \cos k_x x) (C \cos k_y y + D \sin k_y y). \quad (2.18b)$$

Equation (2.17a) and (2.18a) yield $D = 0$ and $k_y = n\pi/b$ for $n = 0, 1, 2, \dots$, whereas

Equation (2.17b) and (2.18b) yield $B = 0$ and $k_x = m\pi/a$ for $m = 0, 1, 2, \dots$. The final

solution for H_z is then

$$H_z(x, y, z) = A_{mn} \cos \frac{m\pi x}{a} \cos \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.19)$$

where A_{mn} is an arbitrary amplitude constant composed of the constants A and C from Equation (2.16). The transverse field components of the TE_{mn} mode are found from Equation (2.8):

$$E_x = \frac{j\omega\mu n\pi}{k_c^2 b} A_{mn} \cos \frac{m\pi x}{a} \sin \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.20a)$$

$$E_y = \frac{-j\omega\mu m\pi}{k_c^2 a} A_{mn} \sin \frac{m\pi x}{a} \cos \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.20b)$$

$$H_x = \frac{j\beta m\pi}{k_c^2 a} A_{mn} \sin \frac{m\pi x}{a} \cos \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.20c)$$

$$H_y = \frac{j\beta n\pi}{k_c^2 b} A_{mn} \cos \frac{m\pi x}{a} \sin \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.20d)$$

where the propagation constant is

$$\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2}, \quad (2.21)$$

which corresponds to a propagating mode when

$$k > k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}. \quad (2.22)$$

Each waveguide mode (represented by a combination of m and n) has a cutoff frequency $f_{c,mn}$, given by

$$f_{c,mn} = \frac{k_c}{2\pi\sqrt{\mu\epsilon}} = \frac{c}{2\pi\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}. \quad (2.23)$$

Similarly, as detailed in [13] and [14], the transverse field components for TM modes is calculated from Equation (2.11) by applying the same separation of variables procedure that is used for the TE mode solution. Once the boundary conditions are applied to e_z , the field components for the TM_{mn} mode are

$$E_x = \frac{-j\beta m\pi}{k_c^2 a} B_{mn} \cos \frac{m\pi x}{a} \sin \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.24a)$$

$$E_y = \frac{-j\beta n\pi}{k_c^2 b} B_{mn} \sin \frac{m\pi x}{a} \cos \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.24b)$$

$$H_x = \frac{j\omega\epsilon n\pi}{k_c^2 b} B_{mn} \sin \frac{m\pi x}{a} \cos \frac{n\pi y}{b} e^{-j\beta z}, \quad (2.24c)$$

$$H_y = \frac{-j\omega\epsilon m\pi}{k_c^2 a} B_{mn} \cos \frac{m\pi x}{a} \sin \frac{n\pi y}{b} e^{-j\beta z}. \quad (2.24d)$$

The propagation constant and the cutoff frequency for the TM modes are the same as the solutions found for the TE modes in Equation (2.21) and (2.23).

The field components found in Equation (2.20) and (2.24) show that the lowest order transverse electric mode is TE_{10} (given $a > b$ from Figure 2.2) and that the lowest order transverse magnetic mode is TM_{11} . The dominant mode in the rectangular waveguide described above is TE_{10} . From Equation (2.21) only

the modes with $f > f_c$ have a real propagation constant (β) and are supported by the waveguide. Figure 2.3 shows the electromagnetic field patterns for some common modes in rectangular waveguide.

Figure 2.3 – Electromagnetic field patterns for modes in a rectangular waveguide

Reprinted with permission from *Fields and Waves in Communication Electronics*, S.

Ramo, J. R. Whinnery, and T. Van Duzer. Copyright © 1965 by John Wiley & Sons, Inc.

Table 8.02.

From the TE and TM waveguide fields described in Equation (2.20) and (2.24) two additional boundary conditions are applied to give the field solution for a rectangular cavity. Shorting the rectangular waveguide at $z = [0, d]$ enforces a perfect electric boundary that sets $E_x = E_y = 0$ on the end walls. Figure 2.4 shows the geometry of a rectangular cavity.

Figure 2.4 – Geometry of a rectangular cavity

From Equation (2.20) the transverse electric fields (E_x, E_y) of the TE_{mn} or TM_{mn} rectangular waveguide mode are rewritten as

$$\vec{E}_t(x, y, z) = \vec{e}(x, y) [A^+ e^{-j\beta_{mn}z} + A^- e^{j\beta_{mn}z}], \quad (2.25)$$

where $\vec{e}(x, y)$ represents the transverse electric field components, and A^+ and A^- are arbitrary amplitudes of the waves moving in $+z$ and $-z$ directions. Applying the boundary condition setting $\vec{E}_t = 0$ at $z = [0, d]$ to Equation (2.25) gives

$$\vec{E}_t(x, y, d) = -\vec{e}(x, y) A^+ 2j \sin \beta_{mn} d = 0. \quad (2.26)$$

If $A^+ \neq 0$, then a solution exists for

$$\beta_{mn} d = l\pi, \quad l = 1, 2, 3, \dots, \quad (2.27)$$

which means the cavity supports resonant modes with integer multiples of half-guide wavelengths. No other wavelengths are supported and the modes that are supported only exist at their discrete resonant frequencies.

The cutoff wavenumber for the rectangular cavity is then

$$k_{mnl} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2}. \quad (2.28)$$

Resonant modes of the cavity are referred to as TE_{mnl} or TM_{mnl} , where the m , n , and l indices refer to the number of guided half-wavelengths found in the standing wave pattern in the x , z , and y directions, respectively. The resonant frequency of the TE_{mnl} or TM_{mnl} mode is

$$f_{mnl} = \frac{ck_{mnl}}{2\pi\sqrt{\mu_r\epsilon_r}} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2}. \quad (2.29)$$

Figure 2.4 shows the rectangular cavity convention that $b < a < d$; the dominant resonance mode (the mode with the lowest resonant frequency) is the TE_{101} mode shown in Figure 2.5. This cavity resonant mode has a discrete resonant frequency different from the cutoff frequency associated with a rectangular waveguide mode.

Figure 2.5 – Field patterns for a) TE, and b) TM modes in a rectangular cavity

Reprinted with permission from M. S. Gupta, "Fundamentals of RF circuits," Nortel Networks, Ottawa, Canada, Internal Technical Report, Nov. 2000.

Figure 2.6 shows a simple rectangular cavity with a similar geometry to the K/Ka-band power amplifier package discussed in Section 4.1. The package supports a TE_{101} dominant resonance mode at 25.06GHz, detailed in Table 2.1.

Figure 2.6 – Rectangular box without a dielectric slab

Table 2.1 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular box without a dielectric slab

Package dimensions (mm)	Dominant mode	Resonant frequency (Equation (2.29))
$a = 8.45$; $b = 1.135$; $d = 8.47$	TE_{101}	25.06GHz

While accurate for the simple model in Figure 2.6, the analytic solution developed above does not account for a dielectric substrate. The solution is unsuitable for a structure such as the one shown in Figure 2.7, which is a better approximation of the package. Robinson, in [15], indirectly addresses this shortcoming with a set of equations based on a dielectrically loaded rectangular

waveguide. His equations find the resonant frequency of a TE_{mnl} or TM_{mnl} mode in a rectangular cavity with a dielectric slab. In [2], Simons re-writes a simplified version of Robinson's solutions as a modification to Equation (2.29) as follows:

$$f_r = f_{mnl} \sqrt{1 - \frac{t}{b} \left(1 - \frac{1}{\epsilon_{r,slab}} \right)}, \quad (2.30)$$

where t is the thickness of the dielectric slab, (t/b) is the dielectric fill factor, and $\epsilon_{r,slab}$ is the relative dielectric constant of the slab. If $(t/b) = 0$ the slab has no thickness and the equation yields $f_r = f_{mnl}$. If $(t/b) = 1$ the slab completely fills the cavity and the solution is the same as predicted by Equation (2.29) for an empty cavity with $\epsilon_r = \epsilon_{r,slab}$.

Combining Equation (2.29) and (2.30) gives the resonant frequency for a TE_{mnl} or TM_{mnl} mode in a rectangular cavity with a dielectric slab as:

$$f_{mnl} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \sqrt{1 - \frac{t}{b} \left(1 - \frac{1}{\epsilon_{r,slab}} \right)}. \quad (2.31)$$

Equation (2.31) yields accurate resonant frequencies when compared with measurements and simulation results presented in [2], [4], [11], [15], and [16].

Figure 2.7 and Table 2.2 describe a cavity with a dielectric slab. The addition of the dielectric slab decreases the dominant resonance frequency by 2.63GHz (10.5%) and improves the accuracy of the model. The dominant resonance frequency of the K/Ka-band power amplifier package is 23.39GHz as shown in Table 4.1.

Figure 2.7 – Rectangular box with a dielectric slab similar to the K/Ka-band power amplifier package

Table 2.2 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular box with a dielectric slab similar to the K/Ka-band power amplifier package

Package dimensions (mm)	Substrate	Dominant mode	Resonant frequency (Equation (2.31))
$a = 8.45$; $b = 1.135$; $d = 8.47$	$t = 0.254\text{mm}$; $\epsilon_{r,\text{slab}} = 9.0$; $\tan \delta = 0.002$	TE_{101}	22.43GHz

Figure 2.8 and Table 2.3 detail a cavity with dielectric slab similar to the broadband two-stage amplifier package discussed in Section 4.2. As with the previous example the dielectric substrate reduces the dominant resonance frequency by 2.55GHz (14.2%). The dominant resonance frequency of the broadband two-stage amplifier package is 15.17GHz as shown in Table 4.18.

Figure 2.8 – Rectangular box with a dielectric slab similar to the broadband two-stage amplifier package

Table 2.3 – Analytic solution of the resonant frequency of the TE_{101} dominant mode in a rectangular cavity similar to the broadband two-stage amplifier package

Package dimensions (mm)	Substrate	Dominant mode	Resonant frequency (Equation (2.31))
$a = 9.1$; $b = 2.135$; $d = 21.1$	$t = 0.635\text{mm}$; $\epsilon_{r,\text{slab}} = 9.0$; $\tan \delta = 0.002$	TE_{101}	15.39GHz

The models shown in Figure 2.7 and Figure 2.8 are simple approximations of the package models presented in Section 4.1 and 4.2, respectively. Analytic solutions cannot address the intricacy of modern complex packages, but the analytic results of the simplified models are often accurate within 10%. These solutions are valuable as a quick estimate of the resonant frequencies and as a starting point for simulations as discussed in Chapter 3.

2.2 Theory of Resonant Modes

Cavity guided TE and TM modes are calculated from an extension of the waveguide solutions to the Helmholtz wave equation in Section 2.1. These resonances are standing wave patterns that oscillate within the conductive enclosure. The modes are combinations of half-wavelengths that are supported at discrete frequencies determined by the geometry and materials of the cavity. Electromagnetic radiation from sources within the cavity force currents to flow in the conductive walls. The currents, shown in Figure 2.5, reinforce the oscillating resonant mode. Once excited, the mode resonates within the cavity and continues to drive the surface currents through the walls and lid.

An infinite number of resonances exist at discrete frequencies in every conductive enclosure. The modes are only problematic when they are excited by radiation sources that couple energy into that particular storage mechanism. The sources must radiate energy at the resonant frequency of the mode being excited. Modes existing at frequencies with little or no radiated power are not supported and minimally affect the performance of the package. Radiation sources are plentiful in complex packages; each transmission line discontinuity and circuit element is a possible radiation source. In [9], Burke finds that coupling to a resonant mode occurs in regions where large current standing waves are present with co-polarized mode field components. In other words, coupling occurs in regions with high signal currents and peak electric fields.

Resonances are supported within every conductive enclosure and it is not possible to eliminate them without removing part of the enclosure and allowing

energy to leak into the environment. This is unacceptable for placing the package in a commercial multi-chip module. The cavity resonates if energy can be coupled into a particular mode faster than energy is lost. Careful material selection and geometry design can reduce the cavity quality factor, but modes still resonate if properly excited with enough energy. This means that resonances cannot be removed from a cavity. However, it is possible to minimize the effects of resonances and reduce their electrical impact on the performance of the enclosed MMICs. The methods of reducing the effects of resonances are: 1) modifying resonant modes by altering the cavity geometry or materials, and 2) reducing the coupling between radiating elements and resonant modes.

Equation (2.31) indicates that the resonant frequency of each mode is a function of the geometry and materials of the package cavity. A sensitivity analysis reveals the resonant frequency increases if the package decreases in length or width; if the dielectric fill factor (t/b) decreases; or if the dielectric constant of the slab decreases. In-band resonant modes are highly excited by strong radiation sources. Shifting the resonant frequencies to low power bands decreases the amount of energy radiated (and thus coupled) from these sources.

The quality factor of each mode is also a function of the cavity materials and geometry. Package cavities comprised of lossy materials and irregular geometries have low quality factors and do not support resonances very well. As resonant modes oscillate, their stored energy is absorbed by the lossy materials and converted into heat. Conversely, packages with low loss materials have high

quality factors and absorb little resonance energy. As low loss barely dampens the standing waves, minimal radiated energy is sufficient to excite and support a resonant mode.

Reducing the amount of energy stored within resonant modes can also be achieved by decreasing the energy coupled from excitation sources. Methods of decreasing the coupling include repositioning the sources within the cavity and disturbing electric and magnetic field distributions.

2.3 Practical Techniques to Mitigate Resonance Effects

Minimizing the effects of cavity resonances in complex packages is a challenging task. Pre- and post-design techniques must be considered to optimize the design cost, time to market, material compatibility, and impact on device performance. A number of design constraints limit the mitigation efforts. This section discusses these issues and introduces four practical mitigation techniques.

As wavelengths decrease towards the package dimensions, resonances are more likely to be located within the operating frequency of the device. Due to the lack of published information and time-to-market pressures, design engineers often do not consider this parasitic effect until device failures are discovered. The cost and time delays associated with redesigning packages are commercially unacceptable. Mitigation then consists of post-design efforts, rather than pre-design consideration, in order to reduce the device failures. This

thesis examines general methods of mitigating cavity resonances, but with an emphasis on a post-design approach.

Packages require a customized application of mitigation techniques depending on the operating bandwidth, signal power, and device gain of the enclosed MMICs. Low-power devices may radiate just enough energy to overcome the lossy cavity materials. These resonances might be easily defeated with the addition of a small amount of lossy material. Narrowband devices with little gain outside the operating bandwidth may require only a small shift of a resonant frequency. Broadband devices often operate over such a wide range of frequencies that moving the resonances out of the device bandwidth is not possible.

The types of MMICs enclosed within a package also determine the best method of mitigation. Amplifiers, especially multi-chip devices like the one discussed in Section 4.2, can suffer from oscillations enabled by feedback through resonant modes. A passive device, like an edge-coupled filter, for example, may be more affected by the radiation element impedance change associated with resonances. Different mitigation techniques should be applied to these packages.

As detailed in Section 2.2, resonance effects can be reduced by changing the resonant frequency and quality factor of particular modes or by decreasing the coupling between excitation sources and the resonances. The resonant frequency and quality factor are a function of the cavity geometry and materials. Both are costly to modify, as a complete re-design and procurement phase is

required. Post-design mitigation efforts must focus on additions to the existing design to inexpensively alter resonant modes. Conductive materials can be added to change the effective cavity volume and lossy materials can be introduced to decrease the quality factor.

Coupling between excitation sources and resonant modes is a function of the effectiveness of the radiating element and its location relative to the peak electric field intensity. Source locations are almost impossible to move within a cavity without redesigning the package. Introducing a grounded protrusion into the cavity can reduce coupling between a radiation source and a resonant mode. Grounded protrusions also break up the regular geometry of the package and disturb the field distribution.

These mitigation efforts are limited by package design constraints. Multi-chip modules often feature multiple packages in close proximity; package shielding is necessary for electromagnetic interference protection. Device reliability requirements mean that packages must be hermetically sealed to protect the enclosed MMICs from corrosion, moisture, and other environmental effects. Electrical performance requirements often restrict the additions that can be made to packages. For example, special care must be taken to ensure protrusions do not interfere with microstrip-guided modes. Low-Q cavities may be advantageous in order to reduce the amount of energy stored in resonant modes, but the device performance must not be compromised by the addition of lossy material. Package space constraints often mean that moving enclosed MMICs

and components around within the package is not possible without compromising the role of the device.

With these limitations in mind, four resonance mitigation techniques are proposed and discussed. These techniques are evaluated in the two case studies in Chapter 4. The first technique consists of adding grounded posts to the package geometry in order to reduce the effective volume of the cavity and increase resonant frequencies. The second technique consists of inserting lossy materials or absorber into the cavity to decrease the quality factors and dampen resonances. The third technique is comprised of modifying excitation sources within the cavity to reduce their coupling to resonant modes. The fourth technique, also intended to reduce coupling, consists of adding protrusions to the underside of the package lid to disturb resonant mode field patterns.

2.3.1 Modifying Package Geometry or Materials

The most direct method of mitigating the effects of resonances is to modify the package enclosure. The geometry and material of the cavity are the determinants of the resonant frequencies. Package materials are selected through rigorous technology and compatibility selection processes. Materials are usually not interchangeable so modifications must focus on the cavity geometry. For pre-design mitigation efforts, this translates to designing a small irregular package that supports modes at the highest possible resonant frequencies. Post-design efforts consist of alterations to the existing geometry in order to reduce the effective volume of the cavity.

To minimize the cost and time associated with alterations, they must not require redesigning package layers and must be possible to implement in an automated manufacturing process. Fabrication procedures must be consulted to determine what modifications are feasible; adding material to the cavity during package fabrication is usually possible. The type and location of the material added determines the effectiveness of the modification. Reducing the effective volume of the cavity is best achieved by creating an alternate electrical path that resonance-induced surface currents can flow through.

If non-conductive material is introduced to the cavity, the surface currents remain unaffected. While the material may change the effective dielectric constant of the cavity, this does not significantly alter the resonant frequencies. Introducing a conductive material disturbs the field distribution, but may not offer surface currents an alternate path. If the protrusion is grounded only at one end, as it is if attached to the underside of the lid, currents cannot flow through to the grounded pedestal. In this case, resonant frequencies are not significantly changed.

Connecting a post from the grounded lid to a grounded pad on the substrate does offer an alternate current path and the effective volume of the cavity is decreased. Resonant frequencies are increased most effectively if this technique is used in peak electric field locations or if the cavity is partitioned into several smaller sections. Manufacturing the conductive post is challenging and is further discussed in Section 4.2.3.

2.3.2 Introducing Microwave Absorber or Resistive Films

The second technique to mitigate the effects of resonances involves using a microwave absorber or resistive film to decrease the cavity quality factor. Lossy materials carefully placed within the cavity can absorb resonance energy without affecting the electrical performance of the MMICs. Modes without stored energy cannot electrically connect radiation elements within the cavity and the effects of resonances are reduced or eliminated.

Williams, in [17], suggests attaching a thin dielectric substrate coated with a resistive film to the underside of the package lid; this technique is also mentioned in [3]. Theory dictates that the surface currents associated with resonant modes flow through the conductive walls of the cavity. By introducing a resistive material, the currents are attenuated and the quality factor of the cavity decreases. However, Burke reports that this simple layer is not enough to reduce the resonance energy stored in most packages and concludes that an absorber is required [18].

Microwave absorbers are materials with a high permittivity and permeability that draw and contain a significant amount of the oscillating electromagnetic field energy. Absorber materials also have high electric and magnetic loss tangents that attenuate contained field energy and convert it into heat. Broadband absorbers have complex permittivity equal to their complex permeability. These materials have an impedance equal to free space and do not reflect radiation at a normal incidence angle. This type of material requires a complete transition from free space to the absorber. Anechoic chambers use this material (formed in

large cones) to capture antenna radiation. This classical type of absorber is not practical for use within packages due to size restrictions [19].

Magnetically loaded absorbers, sometimes referred to as surface current absorbers, are practical for use in packages. They are usually elastomers (silicon or urethane) or epoxies that convert electromagnetic energy into heat through the magnetic re-orientation of iron particles [19]. These absorbers have a high magnetic loss tangent and should be positioned in regions of peak magnetic field intensity. Figure 2.5 shows the magnetic field intensity is strongest close to the walls where surface currents flow. Placing the absorber on a conductive wall with large surface currents presents the material with the maximum intensity of the magnetic field distribution. Absorbers used in this manner effectively attenuate the energy stored in resonant modes without significantly affecting the electrical performance of the device.

Absorbers outgas significant amounts of hydrogen. In a hermetically sealed package the outgas material is contained and can threaten the operation of the device. In [2], Immorlica writes that hydrogen can reach concentrations as high as a few percent in a hermetic package. This amount of hydrogen leads to a degradation of indium phosphide (InP) HEMT and GaAs MESFET and PHEMT devices through a phenomenon known as hydrogen poisoning. This significantly reduces the lifespan of the InP and GaAs devices and threatens the reliability requirements of MMICs enclosed within the package. Materials that absorb hydrogen can be used as an in-package “getter”, but these materials need to have adequate capacity to maintain a safe hydrogen level during the required

lifetime of the device. MMIC passivation and a fabrication process known as “bake-out” can also minimize the effects of outgassing, but absorbers still represent a serious reliability concern [2].

In certain conditions, the application of absorbers is an excellent post-design resonance mitigation technique. Absorbers are inexpensive, easy to apply, and do not require modifying the package design. Absorber use remains desirable, but is a function of the reliability tolerances of the device. Packages with low signal power and gain may be able to benefit from the smaller loss introduced by a resistive material. The resistive layer and substrate do not pose circuit reliability concerns.

2.3.3 Modifying Resonance Excitation Sources

The third resonance mitigation technique consists of modifying excitation sources within the cavity to reduce their coupling to resonant modes. Coupling is a function of the effectiveness of the radiating element and its location relative to the peak electric field intensity. By reducing excitation source coupling, certain resonances are excited to a lesser extent and the effects of these resonances is decreased.

As previously discussed, radiation sources exist at every transmission line discontinuity and circuit element. Reducing the number of sources within a cavity is rarely, if at all, possible due to the large concentration of sources on a typical MMIC die. Other radiation elements like surface-mount capacitors are as much a function of the device performance as the MMICs and cannot be removed from

the package. Reducing the amount of power in the device, and thus radiated, is also not feasible because it means sacrificing device performance.

Moving excitation sources within the cavity is impractical given that the size of the cavity is usually dictated by the MMICs and a few passive components. If the cavity is made larger to move strong radiators away from the centre of the package, then resonant frequencies decrease as outlined by Equation (2.31). Off-centre radiation elements may excite higher-order resonances that would operate at lower frequencies. Overall, optimum source positioning within the cavity is very difficult to determine.

Modifying resonance excitation sources is therefore only an option for relatively simple circuits with few discontinuities. This mitigation technique also requires redesigning the package enclosure and is therefore only viable as a pre-design technique.

2.3.4 Disturbing Resonance Field Patterns

The fourth resonance mitigation technique involves introducing grounded protrusions into the cavity to disturb resonance field patterns. The protrusions break up the regular geometry of the package and can reduce coupling between radiation sources and resonant modes. These cavity discontinuities can also reduce resonance enabled coupling between specific radiation elements.

Chung proposes using vertical diaphragms to partition the cavity into multiple cabinets in [12]. He concludes that without absorber placed in the cabinets, the confined fields couple with the fields supported by the cavity without the

diaphragms. Chung focuses on altering the resonant frequencies with some success, but this technique is more appropriately applied to reducing coupling.

Protrusions can be easily introduced as conductive material attached to the underside of the package lid. The boundary conditions associated with conductive materials force zero electric field intensity over the space occupied by the protrusion. The disturbed field distribution reduces the effectiveness of radiation elements close to the protrusion, especially if located in a position of peak electric field intensity. Consideration must be given to the electrically important height over the enclosed MMICs to avoid disturbing the device performance. Protrusions must not impede on guided modes within the package.

Specific radiation elements (the input to the first stage and the output from the second stage of a two-stage amplifier, for example) can result in catastrophic device failure if electrically connected. Protrusions can be used to effectively partition the resonant modes in a high-gain package cavity. Protrusions only slightly affect resonant frequencies, but significantly decrease coupling between radiation elements separated by the protrusion.

This mitigation technique is difficult to evaluate due to the reduction in radiation element coupling. Driven simulations that consider excitation sources must be used with a high level of detail regarding the sources. As discussed in Chapter 3, MMIC die and passive components are computationally expensive and are not considered for the research presented in this thesis. Nevertheless, this technique is evaluated for both case studies presented in Chapter 4.

Disturbing resonance field patterns is the most appropriate mitigation technique, especially for multi-chip or high gain packages. The device performance is minimally affected by the addition of lid protrusions. This technique requires redesigning the package lid, a relatively inexpensive process, and is viable as a pre- and post-design technique.

Chapter 3 – Finite Element Method Simulations

This chapter evaluates a three-dimensional full-wave structure simulator for calculating the electrical behaviour of packages. The simulator uses the finite element method to discretize package models and solves Maxwell's equations in matrix form. First, the finite element method and its inherent limitations are reviewed. Next, an introduction to Ansoft's HFSS software and a discussion of eigenmode and driven simulations are presented. Finally, model approximations for irrelevant mesh reduction and simulation optimization are addressed.

3.1 The Finite Element Method

Simulators are invaluable for analyzing complex packages that cannot be accurately represented with analytical solutions. Multiple materials and intricate geometries require a three-dimensional full-wave simulator. The finite element method is a numerical technique of analyzing structures by breaking them into multiple subsections. Each smaller section, or element, can be described with a relatively simple set of equations. The electromagnetic behaviour of the structure can then be found as the sum of the behaviours of the individual elements.

The FEM is usually used for simulating specific structural discontinuities. The method is well suited for evaluating small electrical transitions from one medium to another, for example, coaxial to microstrip transitions or thru-wall discontinuities. The FEM is not often used to perform resonance simulations in large packages due to the required computing resources. However, advances in

computing technology and load-sharing facilities have made large FEM simulations possible. Packages approaching 600mm^3 (with $\lambda = 15.0\text{mm}$ at the highest considered frequency) with aspect ratios as large as three orders of magnitude can be considered. This can be seen in the broadband two-stage amplifier package discussed in Section 4.2.

The FEM is generally applied in five stages. First, in the pre-processing stage, the model is developed from the package and conditioned for mesh reduction. Second, the package model is broken into smaller sections by an iteratively developed three-dimensional mesh. Each arbitrarily sized element is defined by a number of mesh nodes. Mesh generation is the most important aspect of the simulation. Third, boundaries are defined to contain the solution space to a specific volume. Fourth, in the solution phase, a system of equations that describes the electromagnetic behaviour within the structure is formulated and solved. The field components of each element are interpolated from the components of the element's nodes. The individual element equations are then combined to form a large system of equations that represent the discretized space. These partial differential equations are solved in matrix form with a sparse matrix solver. The fifth, and final phase, consists of data post-processing. The electromagnetic field distribution is calculated and individual element values are extracted to display the field distribution.

Model approximations, required to decrease the model complexity, reduce the accuracy of the model; generally, element size is inversely proportional to the simulation accuracy. Large elements do not accurately represent the intricate

electromagnetic behaviour of the space they encompass. Smaller elements allow for higher electromagnetic resolution. However, solving the system of equations to calculate the electromagnetic field distribution includes inverting a matrix proportional to the number of element nodes. As element sizes decrease, the required number of elements increases. The computational complexity also increases due to the sparse matrix inversion required to solve the system of equations. Optimizing the simulation and reducing the computational complexity involves minimizing the number of elements in electrically insignificant regions without compromising the accuracy of the simulation.

In order to simulate large models with varying amounts of electromagnetic activity, elements can be defined with variable sizes. The region directly above and below a microstrip transmission line, for example, should be comprised of much smaller elements than a space devoid of electromagnetic energy. Mesh refinement must be performed to develop appropriately dense discretizations in regions of varying field energy. Large elements in electrically significant regions can dramatically reduce the simulation accuracy. Small elements in electrically insignificant regions unnecessarily increase the computational complexity.

Automatic mesh generation and optimization is a challenging research topic that is as important as the solver itself. Model approximations required to minimize unnecessary mesh elements are further discussed in Section 3.3.

3.1.1 FEM Limitations and Considerations

The finite element method is a comprehensive technique that can be applied to find the electromagnetic behaviour of packages. Consideration must be given to inherent limitations of the method. These limitations include the package model, spatial discretization, and numerical techniques used for simulations.

The first set of limitations is associated with the model developed to represent the package. The three-dimensional model is constructed from design drawings and an analysis of fabricated packages. It consists of a collection of two- and three-dimensional objects that are assigned material properties and electromagnetic boundary conditions. The model is enclosed in a solution boundary that defines the specific problem volume considered by the simulation method. The accuracy of the model is limited by the manufacturing tolerances, the correlation between the model and the manufactured package, and the solution boundary.

The package model is a structure that ideally averages material and geometry tolerances introduced during the manufacturing process. Material thickness and positioning may be subject to tolerances in the order of $\pm 50\mu\text{m}$ (2.5% of the cavity height). Each layer and component of the package has a number of tolerances due to manufacturing processes. These tolerances cannot be considered without developing separate models for each permutation of manufacturing inaccuracy. The large number of models required to assess the impact of each tolerance is impractical to consider. The average dimensions and material properties are therefore used to develop a single package model.

While most manufacturing tolerances are relatively small due to automated processes, manufactured packages do not always correspond to the design drawings. The simulation model is constructed from the package design and if the manufacturing process varies from the design, the model inaccurately represents the physical device. The correlation between designed and manufactured packages must be verified for accurate simulation results.

Modeling discrete components that are contained within a package can also present a significant challenge. Passive components, like capacitors and resistors, are often connected to microstrip lines printed on ceramic substrates within the package. These components can be difficult to model due to poorly characterized material types, relatively small feature sizes, and the lack of design information for proprietary third-party devices. Active devices, like MMIC die, are also limited for these reasons and the complexity of modeling semiconducting materials. Approximations can be made, but the case studies analysed in this thesis do not feature passive components or MMIC die, detailed in Section 3.3.

The solution boundary is the final limitation of the package model. The boundary is an electromagnetic discontinuity that does not exist in the physical package. The boundary defines the solution space; it cannot accurately represent the surrounding environment associated with package placement in a multi-chip module. Electromagnetic fields are disturbed at the boundary, which alters the field distribution within the package. The boundary must be large enough to minimally affect the field distribution yet small enough to minimize the solution space. Careful boundary placement is required.

The second set of limitations has to do with the mesh developed to discretize the package model. The mesh is applied to the solution space during the pre-processing stage and divides the model into thousands of elements. Mesh refinement decreases the element sizes in regions of high electromagnetic activity. Meshing has a number of intrinsic limitations, including the accuracy of frequency sweeps and difficulties representing curved surfaces and irrelevant fine features.

The finite element method requires the mesh to be developed at a particular frequency. Regions of high electromagnetic activity are identified from the field distribution within the mesh elements calculated at the meshing frequency. The mesh is then improved upon to decrease the element size in those high-energy locations. Once the mesh has been appropriately refined, it is used to perform simulations at a number of other frequencies. Because the mesh is developed at only one frequency, the accuracy of the simulation results may vary at other frequency points. For simulating resonances, the mesh should be developed as close to the resonant frequency as possible. Meshes must also be separately developed for specific resonant modes due to the different positions of the electromagnetic field intensity. This is further discussed in Section 3.2.2.

Curved surfaces within the model cannot be exactly represented by mesh elements with straight edges. The mesh refinement process generates a large number of elements with small straight edges in order to approximate these curves. Irrelevant fine features and tiny misalignments in the model also generate a large number of elements. These excess elements are not

necessarily required for accurate electromagnetic resolution and inappropriately increase the computational complexity, as discussed in Section 3.3.

The third set of limitations is associated with the mathematical methods employed solve the system of equations that represents the electromagnetic behaviour of the package model. The value of the electric and magnetic field quantities of each element are interpolated from the element vertices. Interpolation errors are introduced by the piecewise polynomials used during the element solutions. Errors are also introduced by the integrations performed to calculate the field distribution in the post-processing phase. Some of these integrations may deliver incorrect results due to the order of the polynomial technique used to perform the integration. These errors are inherent to the FEM and are not avoidable; they represent a relatively minor reduction in solution accuracy when compared to the other FEM limitations [20].

3.2 Ansoft's High Frequency Structure Simulator

Ansoft's HFSS is a three-dimensional full-wave electromagnetic simulator that implements the FEM outlined in Section 3.1. HFSS includes a graphical user interface to build the model, direct the meshing, define and assign material properties, and create boundary conditions. The software tool also includes a suite of post-processing functions to examine the simulated electromagnetic field distribution.

HFSS discretizes models with tetrahedral mesh elements and uses an iterative mesh-refining process. The refinement algorithm often cannot

appropriately mesh the solution space without generating an excessive number of elements in electrically insignificant regions. HFSS allows the mesh process to be seeded with an initial solution, which is required for almost all the models presented in this thesis. User-generated virtual objects can also be used to seed the initial mesh in regions where significant electric fields are expected.

HFSS allows for two different types of solutions: a source-less eigenmode simulation and a driven simulation. Eigenmode simulations are performed without considering excitation sources. Resonances are calculated purely as a function of the model geometry and materials. This solution is used to find the field distribution and associated frequency and quality factor for a perfectly excited mode. Resonances are mathematically calculated from the roots of the polynomial equations that describe the model. Eigenmode simulations identify the exact resonant frequencies of the cavity modes. These simulations are further discussed in Section 3.2.1.

Driven simulations predict the electromagnetic activity generated by an excitation source that is propagated through the model; electromagnetic transmission and coupling are considered. Input and output ports are defined to stimulate activity within the model. Driven simulations more accurately represent the electromagnetic behaviour of a modelled device, but they must be applied at user-defined frequencies. Resonances must be carefully simulated to ensure the exact resonant frequency has been identified. Driven simulations are further discussed in Section 3.2.2.

3.2.1 Eigenmode Simulations

HFSS eigenmode solutions are the perfectly excited natural resonances of the model geometry and materials. These solutions are found from the equation

$$\overline{S}\overline{x} + k_o^2 \overline{T}\overline{x} = 0, \quad (2.32)$$

where $\overline{S}$ and $\overline{T}$ are matrices proportional to the geometry and the mesh, $\overline{x}$ is the electric field solution, and k_o is the free-space wavenumber. The eigenmode solver finds sets of $(k_o, \overline{x})$, one k_o for each solution, $\overline{x}$. The wavenumber is related to the frequency of the resonant modes (f_r) through

$$f_r = \frac{k_o c}{2\pi}. \quad (2.33)$$

The eigenmode solver also finds the unloaded cavity quality factor (Q) for both lossy and loss-less structures. Because ports and other sources are not included for eigenmode solutions, the Q does not include losses from those sources.

The eigenmode solver calculates modes that exist independently from excitation sources. The modes are purely a function of the geometry and materials of the three-dimensional structure. Energy contained in real resonant modes depends on the amount of energy radiated at the resonant frequency, the coupling between this radiation and the mode, and the quality factor of the cavity. Eigenmode solutions predict the frequency and quality of resonant modes that exist if perfectly excited. Eigenmode field pattern magnitudes have arbitrary values and are relative only to other fields in the same model.

The mesh used for an eigenmode simulation must be carefully developed for use with all of the modes simulated. A simulation of the first three modes, for

example, must be performed with a mesh that has been optimized for the electromagnetic activity of all three modes. In this manner, the mesh developed for an eigenmode simulation is often much denser than a mesh developed for a driven simulation.

Figure 3.1 shows the HFSS eigenmode field pattern simulation results for the simple rectangular boxes detailed in Table 2.1 and Table 2.2. The plots show the magnitude of the complex electric field distribution within the cavity. Both the rectangular boxes with and without a dielectric slab have similar field distributions with peak intensity in the centre of the cavity.

Figure 3.1 – HFSS eigenmode electric field pattern simulation results of the TE_{101} dominant mode in a rectangular box a) without a dielectric slab, b) with a dielectric slab

Table 3.1 compares the HFSS eigenmode simulation results to the analytical solutions developed in Section 2.1. The simulation and analytic results agree within 0.5%.

Table 3.1 – HFSS eigenmode simulation results and analytic solutions of the TE₁₀₁ dominant mode in a rectangular box with and without a dielectric slab

	TE ₁₀₁ resonant frequency		Percent difference
	HFSS eigenmode (GHz)	Analytic solution (GHz)	
Rectangular box without a dielectric slab	25.07	25.06	0.04%
Rectangular box With a dielectric slab	22.32	22.43	0.5%

3.2.2 Driven Simulations

When calculating electromagnetic behaviour within a package, HFSS first divides the structure into a finite element mesh. For driven solutions, HFSS then computes the wave excitations on each input port of the structure to develop a more accurate port mesh. The simulator then uses an iterative mesh-refining process that converges on a user-defined level of error. Once an acceptable port and model mesh have been iteratively developed, HFSS performs a sweep of the selected frequencies. The electromagnetic field distribution is then calculated. The generalized S-matrix is derived from the amount of reflection and transmission that occurs at the ports [20].

HFSS driven solutions rely on a number of sources to excite electromagnetic activity within the structure. Sources can arbitrarily be defined as ports, currents, or voltage potentials. The driven solutions are found from an extension of Equation (2.32) as follows:

$$\overline{S}x + k_o^2 \overline{T}x = \overline{b}, \quad (2.34)$$

where $\overline{b}$ is a matrix representation of the sources defined for the problem.

Ports launch pre-formatted electromagnetic energy into the model. First, a two-dimensional solver is used to iteratively generate the optimal port mesh. The field components of the dominant mode on the transmission line structure found at the port are then calculated. This excitation field pattern, represented by $\bar{b}$ in Equation (2.34), is then used as an input to the three-dimensional solver that propagates the port solution through the model. Modes are captured at other ports and S-parameters are determined through the calculated field component strength at the ports. HFSS cannot capture higher order modes when only one mode is used to excite the structure. This can lead to erroneous field component strengths and inaccuracies in the derived S-parameters. HFSS can excite and accept higher order modes from defined ports, but this adds to the computational complexity of the model and requires more time for the simulation process.

Driven simulations more accurately represent the performance of a package compared to source-less eigenmode solutions. A driven solution suggests how much energy is lost to a resonant mode by showing glitches or spikes in the S-parameter performance of the package. In addition, the simulated field distribution can be used to visually identify a resonance and sometimes even the particular radiation element exciting the mode.

Figure 3.2 shows a rectangular box similar to the K/Ka-band power amplifier package presented in Section 4.1. This model has the same geometry and material parameters (detailed in Table 2.2) as the model shown in Figure 2.7. The model features two ports and two microstrip lines on a dielectric slab. The microstrip lines are two-dimensional sheets defined as perfect conductors.

Figure 3.2 – Rectangular box with a dielectric slab, two ports, and two unconnected microstrip lines

Table 3.2 details the HFSS eigenmode simulation results of the first three resonant modes. Since the cavity has the dimensions $a = 8.45\text{mm}$ and $d = 8.47\text{mm}$, the TE_{102} and TE_{201} modes have similar cutoff frequencies as suggested by Equation (2.31). As previously discussed the cavity quality factors represent only the loss associated with the dielectric slab.

Table 3.2 – HFSS eigenmode simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two unconnected microstrip lines

	Mode type	Resonant frequency (GHz)	Quality factor
First mode	TE_{101}	22.57	6667.2
Second mode	TE_{102}	35.08	8070.9
Third mode	TE_{201}	35.20	8116.8

The model shown in Figure 3.2 is also solved with an HFSS driven simulation. Ports 1 and 2 are configured as input and output ports, respectively. The driven 2-port S-parameter simulation results are shown in Figure 3.3; part (a) shows the first three resonant modes and part (b) shows the second and third modes on a smaller frequency scale. Table 3.3 summarizes the driven 2-port simulation results.

The simulation model features two open-circuited microstrip lines connected to each port. S_{11} , the amount of energy reflected back from the input port, should be close to 0dB to indicate that most of the electromagnetic energy is returning from the open-circuited microstrip line within the package. S_{21} , the amount of energy transmitted through the package, should be below -40dB for good isolation between the two unconnected ports. The effects of resonant modes can be seen from the S-parameter results in Figure 3.3. Resonances exist at frequencies where the S_{11} drops and S_{21} correspondingly increases. The sharp dip in S_{11} indicates that much less energy is being reflected back to the input port; the rise in S_{21} indicates that energy is coupling through the resonant mode to the output port.

The effects of the first three resonant modes can be seen at relatively small frequency ranges due to the improbably high quality factors of the modes. These Q values, listed in Table 3.2, define the bandwidth of the coupling between radiation elements and the resonant modes. High-Q modes have very pronounced S_{11} dips and S_{21} spikes; low-Q modes encompass a wider bandwidth, are more rounded, and have less coupling.

(a) Full S-parameters showing the first three resonant modes

(b) Partial S-parameters showing the second and third resonant modes

Figure 3.3 – a) Full, and b) partial HFSS driven 2-port s-parameter simulation results of a rectangular box with a dielectric slab, two ports, and two un-connected microstrip lines

Table 3.3 – HFSS driven 2-port simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two un-connected microstrip lines

	Mode type	Resonant frequency (GHz)	S11 (dB)	S21 (dB)
First mode	TE ₁₀₁	22.37	-14.84	-0.32
Second mode	TE ₁₀₂	35.06	-15.52	-0.90
Third mode	TE ₂₀₁	35.16	-27.54	-0.53

The simple model shown in Figure 3.2 has a large amount of coupling between the input and output ports. At resonant frequencies the S21 is higher than -1dB, indicating that most of the energy input into the device is emerging at the second port. The comparison between the driven S-parameter and eigenmode simulation results is shown in Table 3.4. The resonant frequencies found from the driven simulations agree with the frequencies from the eigenmode simulations within 1%. This level of agreement is higher than expected from a complicated package with additional radiation elements.

Table 3.4 – HFSS eigenmode and driven 2-port simulation results of the first three dominant modes in a rectangular box with a dielectric slab, two ports, and two un-connected microstrip lines

	Mode type	Eigenmode resonant frequency (GHz)	Driven 2-port resonant frequency (GHz)	% difference
First mode	TE ₁₀₁	22.57	22.37	0.89%
Second mode	TE ₁₀₂	35.08	35.06	0.06%
Third mode	TE ₂₀₁	35.20	35.16	0.11%

Because HFSS driven solutions are performed at discrete frequencies a high-Q resonance might not be detected in a coarse sweep. If the mode occurs between two frequency points, the S-parameters may not be noticeably perturbed. An example is described in Section 4.1.1. On the other hand, large bandwidth high-resolution frequency sweeps can be impractical due to the time requirements associated with complex model simulations.

Therefore, for package resonance problems, analytic solutions are first employed to find the approximate resonant frequencies within a simplified model. If resonances are located close to the operating bandwidth of the device, then eigenmode simulations are performed to predict the frequency and quality factors of the modes in an accurate package model. Driven simulations are then employed to thoroughly examine the electromagnetic behaviour in small frequency ranges where resonances are expected. The mesh for the driven simulations is generated at the exact resonant frequency to maximize the accuracy of the model discretization.

3.3 Model Approximations

Package models often have features that generate an excessive number of mesh elements in electrically insignificant regions. Once simulation models have been developed from package geometries they must be pre-conditioned to minimize the number of elements generated by HFSS' iterative mesh-refining process. Model conditioning consists of replacing curved surfaces with multifaceted sections, removing irrelevant fine features, and not considering

complex third-party devices like capacitors or MMIC die. These approximations reduce the accuracy of the model, but significantly decrease the required computing resources. For example, the large broadband two-stage amplifier package in Section 4.2 requires the model approximations to simulate the package at all. Without the approximations the simulations are not possible.

Simulation models are small sections of complete package geometries that focus the results on cavity resonance. The models are reduced to the minimum size required to appropriately represent the cavity while allowing for the consideration of any leaked electromagnetic energy. Jackson finds that structures outside the grounded via perimeter affect cavity resonances [11]. Analysis of electromagnetic fields leaking through the via fence must be performed to help determine the optimum simulation space. External features like biasing networks, surface-mount components, and signal interconnects are not included. The models are also modified to avoid supporting other parasitic effects, specifically thru-wall resonances and other loss mechanisms.

Curved surfaces represent a problem for FEM simulators. A large number of straight edged mesh elements are required to represent each curve. Curvilinear objects are replaced with multifaceted approximations with varying degrees of complexity. Electrically important objects require more detail than those that are not as significant. Figure 3.4 shows some of the curved surface preconditioning performed for the K/Ka-band power amplifier package. The via and via pad closest to the stripline thru-wall transition are represented with 12-sided cylinders and the via pad has the appropriate thickness. Vias that are not electrically close

to the transition are approximated with crude 6-sided cylinders and the via pads have no associated thickness. This pre-conditioning reduces the number of mesh elements in regions of low electromagnetic activity. All curved surfaces in both packages are approximated in this manner.

Figure 3.4 – Curved surface modeling approximations

Like curved surfaces, fine features require a large number of mesh elements for accurate representation. HFSS can simulate aspect ratios of about four orders of magnitude, that is, the smallest object can have dimensions that are approximately 10000 times smaller than the largest object. It is advantageous to limit the aspect ratio as much as possible to minimize the number of mesh elements. Reducing the number of elements decreases the required computing

resources and simulation time. In the case of the larger broadband two-stage amplifier package, which is 21.1mm long, the aspect ratio associated with the thickness of microstrip and DC feed lines is greater than three orders of magnitude. This ratio generates a very large number of mesh elements and the lines are approximated with two-dimensional patterns. These two-dimensional shapes and the electrically insignificant via pads shown in Figure 3.4 are defined as perfect electrical conductors because they cannot be assigned material properties. This approximation is applied to all printed lines and via pads in the broadband two-stage amplifier package and the electrically insignificant via pads in the K/Ka-band power amplifier package. This successfully reduces the number of mesh elements in both cases.

While the purpose of packaging is to protect and support an enclosed MMIC, including the die in package simulations is not possible with current computing technology. Package dimensions are sometimes as many as five orders of magnitude larger than MMIC features. HFSS does not support active devices or semiconducting materials and is not intended for compact circuit design. Furthermore, MMICs are usually designed independently from the package and details are proprietary or unavailable. Passive components like surface mounted capacitors and resistors are also difficult to correctly approximate. For this reason, the package simulation models presented in this thesis do not include MMIC die, passive components, or the supporting wirebonds. The omission of these devices, each representing a discontinuity and multiple radiators, does limit the cavity resonance research presented as discussed in Chapter 5.

Chapter 4 – Case Studies

In this chapter, two packages are modelled and simulated to evaluate the mitigation techniques proposed in Section 2.3. The packages are designed and fabricated by Nortel Networks' Modules and Advanced Packaging group. In-band cavity resonances are simulated and measured in both packages. Ansoft's HFSS is used to assess the mitigation techniques. Modifications are applied to the package models and HFSS is used to determine the effectiveness of each technique. Two mitigation techniques are implemented in the second package and successfully eliminate resonance-induced amplifier oscillations.

The first package considered for evaluating the mitigation techniques is a K/Ka-band power amplifier package developed for use in satellite communication systems and local multipoint distribution systems. The package is hermetically sealed and includes a full ground cage for electromagnetic shielding. The effective volume of the cavity is 8.45mm by 1.135mm by 8.47mm in the x, y, and z directions. An exploded view of this package is shown in Figure 1.1 on page 2. The package is designed to house a single high-gain MMIC amplifier operating in either the K (18-26.5GHz) or Ka (26.5-40GHz) bands. Resonances are supported in both the K and Ka bands. The package is retired and is no longer manufactured. Modifications are not fabricated or otherwise implemented. The goal of the mitigation techniques is undefined; the techniques are evaluated for research purposes only. The K/Ka-band power amplifier package is measured, modelled, and simulated in Section 4.1.

The second package is a broadband two-stage amplifier package developed for use in a 10Gbps optical transmission system. This package is also hermetically sealed and includes a full ground cage for electromagnetic shielding. The effective volume of the cavity is 9.1mm by 2.135mm by 21.1mm in the x, y, and z directions. The package is significantly larger than the K/Ka-band power amplifier package and requires extensive model approximation for successful simulation. An exploded view of this package is shown in Figure 4.1.

The broadband two-stage amplifier package is designed to house two MMIC amplifiers operating from 1-20GHz. The two amplifiers are positioned in close proximity along with a number of passive components within the cavity. Multiple resonant modes are supported in the 15-20GHz band and are responsible for an abnormally high device failure rate. Resonances enable amplifier oscillations by feedback from the output of the second stage to the input of the first stage. The goal of the mitigation techniques applied to this package is to eliminate the amplifier oscillations. Due to model size simulation limitations, the oscillations cannot be simulated; mitigation techniques are evaluated with HFSS and based on the results, two techniques are successfully implemented.

This package is currently in use and concerns regarding the design sensitivity limits the level of detail presented in this thesis. Automated manufacturing concerns are addressed during the application of the mitigation techniques. The broadband two-stage amplifier package is measured, modelled, and simulated in Section 4.2.

Figure 4.1 – Exploded view of the broadband two-stage amplifier package

4.1 K/Ka-Band Power Amplifier Package

The package discussed in this section is designed to house a single high-gain MMIC amplifier operating in the K/Ka (18-40GHz) band. Simulations and measurements affirm that multiple resonances are supported in this band. Nortel Networks' Modules and Advanced Packaging group designed the package. The package is retired and is no longer manufactured. Modifications are not fabricated or otherwise implemented. The goal of the mitigation techniques is open-ended; the proposed techniques are evaluated for research purposes only.

The package is constructed with low-cost materials and fabricated with automated processes. HTCC is used for the package substrate and walls. The substrate features two microstrip and five DC feed lines consisting of a tungsten trace coated with a nickel and gold stack. Twelve tungsten, nickel, and gold pads are located near wirebond landing zones for manual device performance tuning. Tungsten is employed for embedded features in the HTCC layers due to nickel and gold incompatibility with the high curing temperatures of the ceramic. Tungsten vias in the ceramic walls connect the copper tungsten pedestal and heat spreader to the Kovar™ seal ring and lid. The conductive pedestal, vias, and lid define the ground cage required for electromagnetic shielding. The package is hermetically sealed for environmental protection.

The effective volume of the cavity is 8.45mm by 1.135mm by 8.47mm in the x, y, and z directions, that is, 81.2mm^3 . This volume is defined by the top of the heat spreader, the inside edge of the vias, and the underside of the lid. This

measurement does not include the minimal increase in effective volume due to field leakage through the via walls.

An isometric view of the model used for measurement comparison is shown in Figure 4.2. The model shown has been partially pre-conditioned: the model size has been reduced from the full package model, smooth cylindrical vias have been replaced with six- and twelve-sided approximations, and some rounded corners have been replaced with multifaceted sections.

Figure 4.2 – K/Ka-band power amplifier package (measurement comparison model)

Figure 4.3 shows the model after additional pre-conditioning discussed in Section 3.3. The tuning pads have been removed and the overall size of the model has been additionally decreased. The ports excite stripline transmission lines to avoid the microstrip discontinuity associated with the thru-wall transition.

Figure 4.3 – K/Ka-band power amplifier package (simplified model)

The pre-conditioning for the simplified model is successful in reducing the number of tetrahedral mesh elements by approximately 30%. This translates to a significant reduction in computational resource requirements and simulation time. The alterations to the geometry also focus the simulation results on cavity resonance. Figure 4.4 shows the HFSS driven S-parameter simulation results for the full package model (used for measurement comparison) compared with the results of the reduced simulation model. The results show the higher loss and other package effects associated with the larger and more complex model.

Figure 4.4 – HFSS driven 2-port s-parameter simulation results for a) full package model, and b) simulation model of the K/Ka-band power amplifier package

The electrical behaviour of the model is described with S-parameters extracted from the simulation results. The S-parameters are used to identify resonant modes; S21 indicates the amount of energy transmitted through the package (analogous to port isolation), and S11 indicates the amount of energy reflected back to the input port. As shown in Figure 4.2 and Figure 4.3, the full and simplified models feature ports that are connected to open-circuited microstrip lines within the cavity. There is no direct electrical connection between the two ports and the isolation should be greater than 40dB. Resonant modes can be seen in Figure 4.4 where S21 sharply rises and S11 correspondingly drops. At these frequencies, the resonant modes electrically connect the input and output ports of the model.

Figure 4.4 shows that the full model has more loss and includes additional parasitic effects that are not considered in the simplified model. The full model loses a significant amount of energy even at non-resonant frequencies. At 30GHz, for example, the S11 is -1.2dB compared with -0.1dB for the simplified model. The majority of this energy is lost as radiation outside the package cavity and in the thru-wall transition. Additionally, the full model supports a thru-wall resonance seen at 21GHz. The higher loss of the full model is further illustrated in Figure 4.5. The solid lines represent the S-parameter loss (calculated as $|S_{11}|^2 + |S_{21}|^2$ and $|S_{22}|^2 + |S_{12}|^2$) of the full model; the dashed lines represent the loss of the simplified model. The loss increases at resonant frequencies, as energy is captured in resonant modes and attenuated rather than reflected back to the incident port. The full model has higher loss at all frequencies.

Figure 4.5 – HFSS driven 2-port s-parameter loss for a) full package model, and b) simulation model of the K/Ka-band power amplifier package

The full model represents the package more accurately than the simplified model, but is larger, more complicated, and the simulation results are less clear. The full model is used for comparison with package measurements in Section 4.1.1. The simplified model considers only parasitic resonances and is used to evaluate mitigation techniques applied to the cavity. The simulation results verify the pre-conditioning is successful in reducing the unnecessary complexity of the model.

4.1.1 Package Simulations

HFSS eigenmode field pattern simulation results of the first three modes in the simplified K/Ka-band power amplifier model are shown in Figure 4.6 and detailed in Table 4.1. The plots show vector electric field components indicating the magnitude and direction of the electric field. The TE_{101} mode is dominant, as expected, followed by the TE_{102} and TE_{201} modes. The second and third resonant modes are comprised of positive and negative sections of the TE_{102} and TE_{201} standing wave patterns. The direction of the electric field components is indicated by the vector plots and can be seen from close examination of the figure.

The scale of eigenmode simulation results is arbitrary; without sources, the electric field within the cavity is relative only to other fields calculated in the same simulation. The field pattern magnitude cannot be accurately compared with the magnitude of field patterns from other models. Eigenmode field patterns can only be compared based on field density and relative intensity within the cavity. Driven simulations must be performed to compare the field intensity and S-parameters of the package models. The resonant modes shown in Figure 4.6 are shown with relative field intensities.

Figure 4.6 – HFSS eigenmode electric field pattern simulation results of the a) first, b) second, and c) third resonant mode in the K/Ka-band power amplifier package

Table 4.1 presents the resonant frequencies and quality factors of the first three modes in the package. Since the cavity is almost square ($a = 8.45\text{mm}$, $d = 8.47\text{mm}$), the second and third resonances are supported at similar frequencies. This is problematic when considering the driven S-parameter simulation results for some models. The resonant frequencies may vary and be indistinguishable from each other. Careful examination of the simulated field pattern is required to determine if more than one resonance exists at a particular frequency.

Table 4.1 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-Band power amplifier package

	Mode type	Resonant frequency (GHz)	Quality factor
First mode	TE ₁₀₁	23.39	466.7
Second mode	TE ₁₀₂	34.78	362.9
Third mode	TE ₂₀₁	35.24	404.2

Table 4.1 lists the HFSS-simulated cavity quality factors for each of the first three resonant modes. Unlike the quality factors presented in Table 1.2 and Table 3.2, the results presented in this chapter accurately represent the high loss of the package models. The quality factors vary for each mode due to the differences in field distribution seen in Figure 4.6. Modes are more or less lossy depending on their position within the cavity. The quality factors presented in this thesis are unloaded and do not include losses due to ports or other sources.

As discussed in Section 3.2.2, driven simulations are invaluable for understanding the interaction between sources and resonant modes. Introducing ports and exciting the package geometry is a more accurate method of determining the electrical significance of resonances. Driven simulations require more simulation time, but less computational resources than equivalent eigenmode simulations. The simulation time is increased by the need to simulate discrete frequencies to accurately characterize the model. In order to compare the results of driven simulations, the S-parameters for each model must be determined at exact resonant frequencies. Eigenmode simulations calculate these frequencies directly, while driven simulations must be directed to explore discrete frequency points.

Figure 4.7 illustrates this point with a comparison between a 500MHz frequency sweep and an iterative frequency sweep of the second and third resonant modes. The iterative sweep has a much higher level of resolution that must be manually directed in bands where resonances are expected. The 500MHz frequency sweep does not indicate the TE_{102} second resonant mode and does not show the exact resonant frequency of the TE_{201} mode. The results obtained from the 500MHz frequency sweep cannot accurately be compared with other simulation models. All resonant frequencies found from S-parameters calculated by driven simulations are simulated within 50MHz for accurate characterization.

Figure 4.7 – HFSS driven 2-port s-parameter simulation results of the TE₁₀₂ and TE₂₀₁ resonant modes in the K/Ka-band power amplifier package

The first three resonant modes in the package are extracted from the driven simulation results shown in Figure 4.4b, and are listed in Table 4.2. The S-parameter results are found at the exact resonant frequencies shown.

Table 4.2 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package

	Mode type	Resonance frequency (GHz)	S11 (dB)	S21 (dB)
First mode	TE ₁₀₁	23.40	-21.7	-1.7
Second mode	TE ₁₀₂	34.875	-7.2	-4.2
Third mode	TE ₂₀₁	35.475	-22.1	-1.8

The driven simulation results correspond to the eigenmode simulation results from Table 4.1 and the analytic solution to a simple equivalent model from Table 2.2. The analytic solution is accurate only for the first resonant mode as per the limitations of Equation (2.31), but can be used to approximate the second and third resonant modes. This comparison is shown in Table 4.3 and indicates a good correlation between the three different methods.

Table 4.3 – Comparison of analytic solution, HFSS eigenmode simulation results, and HFSS 2-port driven simulation results for the K/Ka-band power amplifier package

	Mode type	Analytic resonant frequency (GHz)	Eigenmode resonant frequency (GHz)	Driven 2-port resonant frequency (GHz)
First mode	TE ₁₀₁	22.43	23.39	23.40
Second mode	TE ₁₀₂	35.44*	34.78	34.875
Third mode	TE ₂₀₁	35.49*	35.24	35.475

* Analytic equations are only rigorously accurate for the TE₁₀₁ dominant mode

4.1.2 Package Measurements

In order to verify the analytic and simulation results and develop confidence in the HFSS tool, several packages are measured for comparison. An Anritsu Universal Test Fixture (UTF) is used with a Hewlett-Packard 8510 vector network analyser (VNA) to perform the package measurements. The UTF interfaces the coaxial test connectors from the VNA to the microstrip lines on the package.

Hewlett-Packard's 8510 VNA is a full-featured network analyser with an operating frequency range of DC-110GHz [21]. Anritsu's 3680V UTF is a fixture used for on-substrate testing of microwave and coplanar waveguide devices. It

connects to test equipment with standard coaxial V connectors and advertises an operating frequency range from DC to 60GHz [22]. The UTF clamps directly onto test substrates with a spring-loaded jaw that provides a ground reference for the fixture's signal pin. A custom jaw attachment is manufactured to connect to the microstrip line that is located in a channel on the K/Ka-band power amplifier package. The UTF is shown with attached V connectors and the package in Figure 4.8.

Figure 4.8 – Anritsu universal test fixture, V connectors, and the K/Ka-band power amplifier package

To separate the electrical characteristics of the UTF from the device under test, the measurements are calibrated at the jaws of the fixture rather than at the coaxial connectors from the VNA. It is often impossible to produce three well-

characterized impedance standards for non-coaxial transmission media [21]. In microstrip, short circuits are inductive and open circuits radiate energy. Rather than using open-short-load standards, calibration substrates are used with a Thru-Reflect-Line (TRL) measurement calibration variation called Line-Reflect-Line (LRL) [21]. The calibration substrates are selected for compatibility with the package. The standards have the same substrate thickness ($t = 0.254\text{mm}$) and a similar dielectric constant ($\epsilon_{r,slab} = 9.9$) to the packages being tested ($t = 0.254\text{mm}$, $\epsilon_{r,slab} = 9.0$).

In our particular case, the LRL calibration technique uses a 1.00cm reference line, a reflection standard, and a 1.07cm line. The frequency range of these standards is from 6 to 60GHz. A zero-length short is defined for the reflection standard, implemented by closing the grounded lower jaw of the UTF on the signal pin. Isolation between the UTF jaws is included in the calibration by using a 50ohm matched load mounted on the same substrate material. Anritsu recommends these calibration standards for this particular UTF [22]. The manufacturer provides the standard definition data.

Once the calibration is performed, two types of checks are used to verify the calibration accuracy. The 1.00cm reference line and a 30ohm Beatty Line standard are measured. Both checks indicate that the calibration is acceptable as shown in the 1.00cm calibration standard measurements in Figure 4.9. The measurements show that most of the energy is transmitted through the standard. This indicates that most parasitic reflections are calibrated out. The measurements are performed with 401 data points and 16-sample averaging.

(a) S21, S12 Measurement of 1.00cm Calibration Substrate

(b) S11, S22 Measurement of 1.00cm Calibration Substrate

Figure 4.9 – Measured a) S21 & S12, and b) S11 & S22 for the 1.00cm calibration standard

Figure 4.10 shows the S-parameter measurements of the package compared with the HFSS driven simulation results for the full model shown in Figure 4.2. The measurements and simulation results show similar S-parameters offset by approximately 1.1GHz; strong trends between the measured and simulated performance can be seen from both the transmission and reflection parameters shown.

The frequency shift between measured and simulated package performance is likely a function of the approximations made when building the simulation model. Furthermore, material imperfections and geometry tolerances introduced during device production could not be accounted for in simulation models. The magnitude differences between the measured and simulated package performance are in part due to the modeling limitations discussed in Section 3.1.1.

The measured S12 and S21 parameters at the three resonant modes are seen to be greater than 15dB above the calibration floor shown in Figure 4.9b. This difference is acceptable for the comparison and reinforces the validity of the calibration and measurements. The measured and simulated S-parameter results for the package model confirms that the simulator is an accurate and effective analysis tool and that the simulation models are acceptable.

Figure 4.10 – Measurements and HFSS eigenmode simulation results of the a) S11 & S22, and b) S21 & S12 for the K/Ka-band power amplifier package

4.1.3 Modifying Package Geometry to Decrease the Effective Volume

Detailed in Section 2.3.1, the first and most direct mitigation technique involves modifying the package geometry. Decreasing the effective cavity volume increases resonance frequencies. Post-design mitigation efforts rule out re-designing package cavities, but conductive material can be introduced to effectively decrease the volume of the cavity.

Four gold-plated tungsten posts are introduced to the corners of the package as shown in Figure 4.11. Each post is 0.3mm by 0.881mm by 0.3mm in the x, y, and z directions. The posts are electrically connected with a tungsten via to the grounded heat spreader below the ceramic substrate. Each post directly connects to the underside of the lid and represents an alternate path for the resonance-induced surface currents.

Figure 4.12 shows a vector plot of the simulated HFSS eigenmode electric field pattern within this modified cavity. The field has not been appreciably disturbed from the unmodified package; the patterns for all three modes (the second and third mode field patterns are not shown) are minimally affected by the addition of the posts.

Figure 4.11 – K/Ka-band power amplifier package with four grounded posts in the corners

Figure 4.12 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-Band power amplifier package with four grounded posts in the corners

Table 4.4 summarizes the HFSS eigenmode simulation results for the first three resonant modes supported by the modified package. A 5.6-6.5% increase is seen in the resonant frequencies of the first three modes. The quality factors are relatively unchanged.

Table 4.4 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the corners

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	24.72, [5.7%]	477.9, [2.4%]
Second mode	TE ₁₀₂	36.73, [5.6%]	332.8, [-8.3%]
Third mode	TE ₂₀₁	37.54, [6.5%]	411.0, [1.7%]

Figure 4.13 and Table 4.5 summarize the results calculated from exciting the modified package. These simulations show a similar 5.8-7.1% increase in the frequencies and relatively insignificant changes in the S-parameters for the first three resonant modes. The agreement between the frequencies predicted by the eigenmode and driven simulations is very good.

These results suggest that the grounded posts minimally reduce the effective volume of the cavity. The corner positions of the four posts are not effective enough at changing the field distribution due to the relatively low electric field intensity at those locations.

Figure 4.13 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with four grounded posts in the corners

Table 4.5 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the corners

	Mode type	Resonance frequency (GHz), [% change]	S11 (dB), [dB change]	S21 (dB), [dB change]
First mode	TE ₁₀₁	24.75, [5.8%]	-13.9, [-7.8dB]	-1.8, [0.1dB]
Second mode	TE ₁₀₂	36.80, [5.5%]	-4.2, [-3.0dB]	-6.1, [1.9dB]
Third mode	TE ₂₀₁	38.00, [7.1%]	-29.3, [7.2dB]	-1.6, [-0.2dB]

The four gold-plated posts and tungsten vias are repositioned within the cavity to place them in regions of higher electric field intensity. The posts are moved to the middle of the sides as shown in Figure 4.14; the HFSS eigenmode electric field pattern simulation results are shown in Figure 4.15.

Figure 4.14 – K/Ka-band power amplifier package with four grounded posts in the middle of each side

Figure 4.15 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with four grounded posts in the middle of each side

The simulated electric field pattern shows a significant change from the simulations of the unmodified package. As expected, the field pattern is notably disturbed around the grounded posts. The eigenmode simulation results, summarized in Table 4.6, show that the posts increase the resonant frequency of the TE_{101} dominant mode by 5.35GHz (22.9%).

Table 4.6 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the middle of each side

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE_{101}	28.74, [22.9%]	455.1, [-2.5%]
Second mode	TE_{102}	37.42, [7.6%]	317.6, [-12.5%]
Third mode	TE_{201}	38.86, [10.3%]	291.0, [-28.0%]

The second and third resonances are not affected as much as the first mode, increasing by 7.6% and 10.3%. This relates to the position of the posts relative to the peak electric field distribution of each mode within the cavity. The peak intensity of the second and third modes is not centrally located and is therefore less affected by the addition of the grounded posts.

The quality factor of the cavity significantly decreases for the TE_{201} third resonant mode. This change is related to the standing wave pattern of the third mode being closer to the corners of the cavity. Higher field intensity within the ceramic walls of the cavity results in higher losses; the quality factor decreases accordingly.

The HFSS driven S-parameter simulation results for this package are shown in Figure 4.16 and summarized in Table 4.7. These results show almost identical increases in the resonant frequencies compared with the eigenmode simulations. The driven S-parameter simulation results show a slight 1dB increase in isolation for the first resonant mode and a much more significant 6dB and 10dB increase in isolation for the second and third modes. These increases are also related to the proximity of the posts to the regions of peak field intensity within the cavity. The posts are located between the excited microstrip line and the peak electric field of the resonant modes and decrease the effectiveness of radiation coupling.

Figure 4.16 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with four grounded posts in the middle of each side

Table 4.7 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with four grounded posts in the middle of each side

	Mode type	Resonant frequency (GHz), [% change]	S11 (dB), [dB change]	S21 (dB), [dB change]
First mode	TE ₁₀₁	28.73, [22.8%]	-10.4, [-11.3dB]	-2.7, [1.0dB]
Second mode	TE ₁₀₂	37.50, [7.5%]	-9.3, [2.1dB]	-10.1, [5.9dB]
Third mode	TE ₂₀₁	38.90, [9.7%]	-0.85, [-21.3dB]	-12.3, [10.5dB]

The gold-plated posts represent a manufacturing challenge, further discussed in Section 4.2.3. This modification also requires a re-design of the HTCC ceramic substrate to add the tungsten vias, necessary for grounding the posts to the heat spreader. Some packages may not require a substrate re-design due to the presence of grounded pads on the substrate. The package presented in Section 4.2, for example, features grounded substrate pads used for passive components.

The simulation results indicate that modifying the package by adding grounded posts is an effective method of increasing resonant frequencies. Specific resonances may be suppressed by positioning posts between strong radiation elements and regions of peak electric field intensity. The posts do not significantly reduce radiation coupling to all resonant modes and some cavity quality factors are not appreciably changed. This mitigation technique is most effective for increasing the resonant frequency of a specific mode in a narrowband application.

4.1.4 Introducing Absorber to Reduce the Cavity Quality Factor

The second mitigation technique involves decreasing the cavity quality factor. Introducing a material with a high magnetic loss tangent into regions of peak magnetic field intensity increases the amount of energy dissipated. Resonance effects can be suppressed by attenuating the energy stored in the mode. This strategy and details regarding the type of material required for optimum loss is discussed in Section 2.3.2.

ECCOSORB® GDS absorber is a broadband magnetically loaded material with a high permeability and high magnetic loss tangent. Manufactured by Emerson & Cuming, this non-conductive silicone rubber material is available in flexible sheets of varying thickness. The material properties of this absorber are shown in Table 4.8.

Table 4.8 – ECCOSORB® GDS material properties

Permittivity (ϵ_r)	Electric loss tangent ($\tan \delta_e$)	Permeability (μ_r)	Magnetic loss tangent ($\tan \delta_m$)
13	0.2	1.7	0.78

A 0.254mm-thick slab of ECCOSORB® GDS absorber is added under the lid of the package model. The absorber is directly connected to the lid, flush with the cavity walls. The model is shown in Figure 4.17 and the eigenmode simulation results are listed in Table 4.9.

Figure 4.17 – K/Ka-band power amplifier package with ECCOSORB® GDS absorber under the lid

Table 4.9 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with ECCOSORB® GDS absorber under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	18.71, [-20.0%]	3.63, [-99.2%]
Second mode	TE ₁₀₂	28.67, [-17.6%]	3.16, [-99.1%]
Third mode	TE ₂₀₁	28.68, [-18.6%]	3.09, [-99.2%]

As expected, the cavity quality factors for the first three resonant modes in the package are significantly reduced by more than 99%. The resonant frequencies of the modes are decreased by 17-20% due to the increased dielectric loading of the cavity.

The HFSS driven S-parameter simulation results show port isolation is greater than 50dB across the operating frequency of the device and no cavity resonance perturbations are present. These results confirm the hypothesis that the cavity is so lossy that radiated electromagnetic energy cannot be stored in a standing wave pattern. The detrimental effects of resonances are eliminated. Measurements must be performed to determine if the absorber affects the electrical performance of the device.

Absorber materials are responsible for hydrogen outgassing that can poison GaAs transistors, detailed in Section 2.3.2. Some packages may be unable to tolerate absorber due to reliability requirements. This is further discussed in Section 4.2.5 where the amount of absorber is minimized to reduce hydrogen outgassing. Nevertheless, for packages without strict hermetic requirements, introducing absorber is a good post-design mitigation technique.

4.1.5 Moving Excitation Sources to Reduce Resonance Coupling

The third mitigation technique involves decreasing coupling to resonant modes. Two additional HTCC ceramic substrates are added in order to evaluate the effect of moving excitation sources within the cavity. The substrates feature gold plated microstrip lines. The additional lines are connected with two gold wirebonds to the input and output microstrip lines as shown in Figure 4.18. The open-circuited ends of the microstrip lines are strong radiators. By locating the ends of the lines closer to the peak electric field intensity of the first resonant mode, a lower level of port isolation is expected.

Figure 4.18 – K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates

The HFSS eigenmode field pattern simulation results for this model are shown in Figure 4.19. The field pattern is not appreciably different from the distribution of the unmodified package. The eigenmode simulation results are listed in Table 4.10. The resonant frequencies are slightly affected by the increased dielectric loading of the cavity and the quality factors are relatively unchanged.

Figure 4.19 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates

Table 4.10 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	22.18, [-5.2%]	458.2, [-1.8%]
Second mode	TE ₁₀₂	33.68, [-3.2%]	392.0, [8.0%]
Third mode	TE ₂₀₁	34.05, [-3.4%]	415.2, [2.7%]

Surprisingly, the HFSS driven simulation results show almost no dominant mode coupling between the input and output ports of the model. Figure 4.20 shows the S-parameter results.

Figure 4.20 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates

The close proximity of the open-circuited ends of the microstrip lines to the peak electric field intensity of the first resonant mode suggests a large amount of coupling. The cavity quality factor is unchanged from the original model and no protrusions are added between the ends of the microstrip lines. However, the driven simulation results show a 24.7dB decrease in coupling for the first resonant mode. The second mode is also notably affected by the modification; the isolation between the ports increases by 5.6dB. The third mode, with peak electric field distribution located over the two pairs of wirebonds, is relatively unaffected by the package modification. These results are listed in Table 4.11.

Table 4.11 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with two wire-bonded microstrip lines on separate substrates

	Mode type	Resonant frequency (GHz), [% change]	S11 (dB), [dB change]	S21 (dB), [dB change]
First mode	TE ₁₀₁	22.375, [-4.4%]	-0.6, [-21.1dB]	-26.4, [24.7dB]
Second mode	TE ₁₀₂	33.75, [-3.2%]	-1.1, [-6.1dB]	-9.9, [5.6dB]
Third mode	TE ₂₀₁	34.75, [-2.0%]	-23.9, [0.7dB]	-1.7, [-0.1dB]

The simulation results reinforce the theory that modifying the excitation sources can significantly alter the performance of the model. The results also underline the importance of driven simulations when considering the effects of resonances. The eigenmode simulation results do not suggest a reduction in coupling between the input and output ports.

The analysis of this mitigation technique does not accurately reflect the large number of radiators that are associated with a MMIC die. It is not realistic to consider moving the die location within the cavity or to reduce the number of radiators associated with the die. This mitigation technique is therefore useful only when considering simple structures in relatively large cavities.

4.1.6 Disturbing Field Patterns to Decrease Coupling

The fourth mitigation technique also involves decreasing coupling to resonant modes. This section evaluates disturbing local field patterns by modeling modifications to the lid of the package, as discussed in Section 2.3.4. A small amount of conductive material is added to the underside of the lid, shown in Figure 4.21. The gold-plated cylindrical addition has a diameter of 1mm and

protrudes 0.3044mm into the cavity. The protrusion is located in the position of peak electric field intensity for the TE_{101} dominant mode.

Figure 4.21 – K/Ka-band power amplifier package with cylindrical protrusion under the lid

The HFSS eigenmode field pattern simulation results are altered by the cylindrical protrusion as shown in Figure 4.22. The magnitude of the electric field pattern underneath the protrusion is higher than elsewhere within the cavity. The change in field distribution is not reflected in the eigenmode simulation results listed in Table 4.12. The resonant frequencies and cavity quality factors of the first three modes are relatively unchanged.

Figure 4.22 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with cylindrical protrusion under the lid

Table 4.12 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with cylindrical protrusion under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	23.01, [-1.6%]	464.4, [-0.5%]
Second mode	TE ₁₀₂	34.91, [0.4%]	352.2, [-2.9%]
Third mode	TE ₂₀₁	35.27, [0.1%]	391.4, [-3.2%]

The driven S-parameter simulation results are shown in Figure 4.23. The results confirm the significant change in package performance expected when the field patterns are locally disturbed. Coupling between the input and output ports is reduced by 19.1dB for the first resonant mode and 15.9dB for the second mode, as detailed in Table 4.13.

Figure 4.23 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with cylindrical protrusion under the lid

Table 4.13 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with cylindrical protrusion under the lid

	Mode type	Resonant frequency (GHz), [% change]	S11 (dB), [dB change]	S21 (dB), [dB change]
First mode	TE ₁₀₁	23.01, [-1.7%]	-1.1, [-20.6dB]	-20.8, [19.1dB]
Second mode	TE ₁₀₂	34.90, [0.1%]	-0.8, [-6.4dB]	-20.1, [15.9dB]
Third mode	TE ₂₀₁	35.33, [-0.4%]	-4.7, [-17.4dB]	-8.7, [6.9dB]

The package is further modified to calculate the effect of a different protrusion shown in Figure 4.24. The gold-plated conductive diagonal bar is 7.6mm by 0.5mm in length and width, and protrudes 0.3044mm into the cavity. The protrusion is directly connected to the underside of the lid and located in the position of minimum electric field of the TE₂₀₁ third resonant mode.

Figure 4.24 – K/Ka-band power amplifier package with diagonal protrusion under the lid

Figure 4.25 shows that the electric field distribution is also disturbed under the protrusion. The modification slightly alters the resonant frequencies and cavity quality factors of the first two resonant modes, listed in Table 4.14. The TE_{201} third resonant mode is least affected by the addition of the diagonal bar due to the placement of the protrusion.

Figure 4.25 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the K/Ka-band power amplifier package with diagonal protrusion under the lid

Table 4.14 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with diagonal protrusion under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	22.68, [-3.0%]	467.8, [0.2%]
Second mode	TE ₁₀₂	34.15, [-1.8%]	376.1, [3.6%]
Third mode	TE ₂₀₁	35.51, [0.8%]	385.4, [-4.7%]

The driven S-parameter simulation results are shown in Figure 4.26 and summarized in Table 4.15. As expected, they show a dramatic decrease in coupling between the input and output ports of the package.

Figure 4.26 – HFSS driven 2-port s-parameter simulation results of the K/Ka-band power amplifier package with diagonal protrusion under the lid

The second resonant mode has an additional 2.4dB isolation compared to the model with the cylindrical protrusion. This is due to the extended length of the protrusion interrupting the peak field intensities of this higher-order mode.

Table 4.15 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with diagonal protrusion under the lid

	Mode type	Resonant frequency (GHz), [% change]	S11 (dB), [dB change]	S21 (dB), [dB change]
First mode	TE ₁₀₁	22.67, [-3.1%]	-1.0, [-20.7dB]	-21.4, [19.7dB]
Second mode	TE ₁₀₂	34.15, [-2.1%]	-1.4, [-5.8dB]	-22.5, [18.3dB]
Third mode	TE ₂₀₁	35.50, [0.1%]	-3.8, [-18.3dB]	-8.5, [6.7dB]

The simulation results confirm that disturbing the electric field pattern effectively reduces coupling. Adding a protrusion in the region of peak electric field intensity is shown to disturb the field distribution and decrease resonant mode coupling. The protrusion does not decrease the effectiveness of radiation sources within the package, nor does it reduce the amount of energy radiated. The protrusion reduces the effectiveness of the resonance as a parasitic electrical connection between two points within the cavity. In the models presented above, the protrusion significantly decreases coupling between the input and output ports.

This mitigation technique has not been considered with the enclosed MMIC die. The die, along with the associated wirebonds and passive components, would radiate at many different locations within the cavity. It is not clear if this technique reduces coupling between points that are uninterrupted by a protrusion. This possible limitation is partially addressed in Section 4.2.3 with favourable results. Locally disturbing field patterns by modifying the lid of the broadband two-stage amplifier package is successful in eliminating amplifier oscillation supported by feedback through a cavity resonance.

Disturbing field patterns to reduce resonance coupling is an excellent mitigation technique that can be applied in a pre- or post-design situation with minimal time and cost requirements. Prototype lids can be modified with solder blobs or conductive epoxy; lid re-design is an inexpensive and relatively quick process. This mitigation technique is applicable to all types of packages.

4.1.7 Summary of Results

The HFSS eigenmode simulation results of the four mitigation techniques discussed in Section 4.1.3 - 4.1.6 are summarized in Table 4.16. The most effective technique for increasing the resonant frequencies is modifying the package geometry to decrease the volume of the cavity, as discussed in Section 4.1.3. This technique increased the dominant resonant frequency by 5.35GHz (22.9%). Adding microwave absorber is the only method that significantly reduced the cavity quality factor for the first three resonances.

Table 4.16 – HFSS eigenmode simulation results of the first three resonant modes in the K/Ka-band power amplifier package with various modifications

	TE ₁₀₁ first mode frequency (GHz), quality factor	TE ₁₀₂ second mode frequency (GHz), quality factor	TE ₂₀₁ third mode frequency (GHz), quality factor
Unmodified package	23.39, 466.7	34.78, 362.9	35.24, 404.2
Four posts in the cavity corners	24.72, 477.9	36.73, 332.8	37.54, 411.0
Four posts in the middle of the sides	28.74, 455.1	37.42, 317.6	38.86, 291.0
ECCOSORB® GDS absorber	18.71, 3.63	28.67, 3.16	28.68, 3.09
Modified excitation source position	22.18, 458.2	33.68, 392.0	34.05, 415.2
Cylindrical lid protrusion added	23.01, 464.4	34.91, 352.2	35.27, 391.4
Rectangular lid protrusion added	22.68, 467.8	34.15, 376.1	35.51, 385.4

The HFSS driven 2-port simulation results of the four mitigation techniques discussed in Section 4.1.3 - 4.1.6 are summarized in Table 4.17. The resonant frequencies have a strong correlation to the frequencies found from the eigenmode simulations. Modifying the package geometry to decrease the effective volume of the cavity did not appreciably increase the port isolation. Moving excitation sources and disturbing resonance field patterns significantly decreased reflection loss from the input port and increased port isolation. Introducing absorber completely eliminated the effects of resonant modes.

Table 4.17 – HFSS driven 2-port simulation results of the first three resonant modes in the K/Ka-band power amplifier package with various modifications

	TE ₁₀₁ first mode frequency (GHz), S11 (dB), S21 (dB)	TE ₁₀₂ second mode frequency (GHz), S11 (dB), S21 (dB)	TE ₂₀₁ third mode frequency (GHz), S11 (dB), S21 (dB)
Unmodified Package	23.40, -21.7, -1.7	34.875, -7.2, -4.2	35.475, -22.1, -1.8
Four posts in the cavity corners	24.75, -13.9, -1.8	36.80, -4.2, -6.1	38.00, -29.3, -1.6
Four posts in the middle of the sides	28.73, -10.4, -2.7	37.50, -9.3, -10.1	38.90, -0.85, -12.3
ECCOSORB® GDS absorber	N/A	N/A	N/A
Modified excitation source position	22.38, -0.6, -26.4	33.75, -1.1, -9.9	34.75, -23.9, -1.7
Cylindrical lid protrusion added	23.01, -1.1, -20.8	34.90, -0.8, -20.1	35.33, -4.7, -8.7
Rectangular lid protrusion added	22.67, -1.0, -21.4	34.15, -1.4, -22.5	35.50, -3.8, -8.5

4.2 Broadband Two-Stage Amplifier Package

The package discussed in this section is designed to house two broadband MMIC amplifiers operating from 1-20GHz. Simulations and system-level measurements affirm that multiple resonances are supported in the 15-20GHz frequency band. These resonances enable amplifier oscillations and are responsible for an abnormally high device failure rate. Nortel Networks' Modules and Advanced Packaging group designed the package. The package is currently in use and concerns regarding the design sensitivity significantly limit the level of detail presented in this thesis. A research agreement with Nortel Networks allows a comprehensive evaluation of the resonance mitigation techniques in a state-of-the-art package. Due to the recent design and ongoing manufacturing, additional resources are available and several mitigation techniques are successfully implemented.

The broadband two-stage amplifier package is similar to the previous package discussed in Section 4.1. HTCC is used for the package substrate and walls, tungsten is employed for embedded features in the ceramic, and the seal ring and lid are manufactured with Kovar™. Signal and bias lines within the cavity consist of the same tungsten-nickel-gold stack. Tungsten vias connect the heat spreader (not shown in models) to the seal ring and lid, defining a similar ground cage required for electromagnetic shielding. The package is also hermetically sealed for environmental protection.

The package is approximately five times larger than the K/Ka-band power amplifier package. The cavity houses two amplifiers and a number of surface-

mount passive components, none of which are modelled. The effective volume of the cavity is 9.1mm by 2.135mm by 21.1mm in the x, y, and z directions, or 410mm^3 . The cavity volume is defined by the top of the heat spreader, the inside edge of the vias, and the underside of the lid.

An exploded view of the full package is shown in Figure 4.1 on page 64, an isometric view is shown in Figure 4.27. The model has been partially pre-conditioned: round vias have been replaced with six-sided cylinders, the simulation volume has been significantly reduced, and many curved surfaces have been approximated.

Figure 4.27 – Broadband two-stage amplifier package (partially pre-conditioned model)

Figure 4.28 shows the model after the remaining pre-conditioning discussed in Section 3.3. The microstrip and DC feed lines have been replaced with two-dimensional shapes, substrate vias within the package cavity have been removed, and the simulation space has been reduced. The two-dimensional microstrip and DC feed lines are defined as perfect conductors because they cannot be assigned material properties.

Figure 4.28 – Broadband two-stage amplifier package (simplified model)

As with the previous package, the pre-conditioning is successful in significantly reducing the number of tetrahedral mesh elements. This reduction is in the order of 40%, but the simplified model is still notably large. Eigenmode

solutions require simulation times in excess of 48 hours on quad-processor Sun UltraSPARC workstations with more than 4Gb of RAM. Simulation results are therefore compared only for the first two, rather than the first three, resonant modes. The size and complexity of this model also makes driven simulations impractical due to the number of iterations required for accurate S-parameter characterization. This limitation reduces the analysis of mitigation techniques to those that do not require exciting the package with radiation sources. Disturbing local field patterns and relocating excitation sources are not simulated.

The mitigation techniques are considered with emphasis placed on cost, time, and manufacturing concerns. The techniques are simulated and applied to the package with the goal of eliminating the resonance-induced amplifier oscillations. Grounded posts are considered in order to partition the cavity into two smaller sections. Concerns regarding the manufacturing process limit the implementation of this technique, discussed in Section 4.2.3. Magnetically loaded absorber is modelled and introduced to the cavity, outlined in Section 4.2.5. Lid protrusions designed to disturb the electric field distribution are discussed in Section 4.2.3. Both the absorber and lid protrusion techniques are implemented and successfully eliminate the amplifier oscillations.

4.2.1 Package Simulations

HFSS eigenmode field pattern simulation results of the first two modes in the simplified model are shown in Figure 4.29. The plots show the vector electric field distribution indicating the magnitude and direction of the electric field. The

TE_{101} mode is dominant, as expected, followed by the TE_{102} and TE_{103} (not shown) modes. The eigenmode simulation results are listed in Table 4.18.

Figure 4.29 – HFSS eigenmode electric field pattern simulation results of the a) first and b) second resonant modes in the broadband two-stage amplifier package

Table 4.18 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package

	Mode type	Resonant frequency (GHz)	Quality factor
First mode	TE ₁₀₁	15.17	636.0
Second mode	TE ₁₀₂	17.92	594.1

As shown in Figure 4.28, the simplified model has a number of two-dimensional traces on the substrate. Three of the traces are microstrip lines for the signal path through the package. The other traces are required for DC biasing and connecting passive components electrically close to the MMIC die. In a populated package, wirebonds, resistors, and capacitors electrically connect these traces, but these passive components are not included in the simulation models. Table 4.19 lists the HFSS eigenmode simulation results of the package without the planar lines.

Table 4.19 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package without microstrip and DC feed lines

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	14.93, [-1.6%]	626.0, [-1.6%]
Second mode	TE ₁₀₂	17.75, [-0.9%]	652.3, [9.8%]

Removing the microstrip and DC feed lines from the package results in a 1.6% and 0.9% change in the frequencies of the first two resonant modes. This change indicates the relatively minor difference the lines cause to cavity

resonances. However, the number of mesh elements did not appreciably decrease and the microstrip and DC feed lines are included for model accuracy.

Resonance theory details that surface currents flow on the inside conductive walls of the cavity as a result of the oscillating electromagnetic waves contained within. Figure 4.30 shows the surface currents for the TE_{101} dominant mode on the underside of the Kovar™ lid. The currents are shown flowing towards the centre of the lid. The peak electric field is located in the centre of the cavity as shown in Figure 4.29a. The intensity of these surface currents is considered when applying patches of absorber in Section 4.2.5.

Figure 4.30 – HFSS eigenmode surface current simulation results of the first resonant mode in the broadband two-stage amplifier package

4.2.2 Package Measurements

The broadband two-stage amplifier package design is commercially sensitive material owned by Nortel Networks. The research agreement involving this thesis prohibits the publication of specific measurement, test, or process information regarding this advanced package.

Measurements are performed with fully populated packages, including the passive components, wirebonds, and MMIC die that are not modelled in the presented research. A diode rectifier is connected to the package to measure the output power and perform voltage detect functions. The tests that are used to identify the device failures as resonance-related are performed with maximum gain biasing for both amplifiers. The first test is performed with no input to the first stage and the second test is performed with a 50ohm termination connected to the first stage. Devices producing low voltage outputs are considered to pass. Devices with higher voltage outputs at certain frequencies suggest the enclosed amplifiers are oscillating.

Resonance effects were not apparent in the alpha and beta phase of new product integration and pre-production manufacturing runs did not reveal a problem. Packages were in full production before an atypically high failure rate was discovered during device test and verification. The device failures were related to amplifier oscillations at specific frequencies. Test and verification engineers confirmed that the amplifiers were oscillating at two in-band frequencies. Resonances were suspected when the oscillations were not detected if the package lid was removed. The package simulations, shown in

Figure 4.28, confirm the existence of resonant frequencies within 7% of the oscillation frequencies found in the output spectra of the failing devices.

4.2.3 Disturbing Field Patterns to Eliminate Amplifier Oscillations

The first implemented mitigation technique involved disturbing the field distribution to reduce the amount of feedback between the first and second amplifier in the package. The simulation results from the K/Ka-band power amplifier package in Section 4.1.6 suggest that a grounded lid protrusion between the feedback points is successful in reducing coupling.

A test engineer proceeded to dent the lid of several oscillating packages with a screwdriver. The dent was applied crosswise above the microstrip line between the two MMIC die and successfully eliminated the amplifier oscillation in all of the oscillating devices. The effect of the dent suggests that the oscillation is caused by feedback from the output of the second stage amplifier to the input of the first stage. The denting process is not acceptable from a manufacturing and reliability point of view and the lid design shown in Figure 4.31 was procured and used with some success.

Figure 4.31 – Broadband two-stage amplifier package with lid protrusion

A limited production run of packages with the modified lids resulted in a decrease, but not complete elimination, of the percentage of packages failing due to amplifier oscillations. The modified lid did not otherwise affect the electrical performance of the devices. Alternate lid protrusions are currently under investigation. Protrusion placements above the two amplifiers, as well as an increase in the depth of the protrusions, are promising techniques to eliminate amplifier oscillations.

4.2.4 Modifying Package Geometry to Partition the Cavity

The geometry of the package supports the first and second resonant modes at similar frequencies. The 2.75GHz difference between the modes indicates that reducing the volume of the cavity does not result in a significant increase in resonant frequencies. The package geometry is less conducive to modification compared to the K/Ka-band power amplifier package. Equation (2.31) suggests that reducing the length of the cavity by 50% increases the first resonant frequency by 20%, while the same reduction for the square K/Ka-band power amplifier package increases the frequency by 37%.

Figure 4.32 shows the package with a single gold-plated tungsten post connecting a grounded DC feed line to the underside of the lid. The post is 0.9mm by 1.5mm by 1.0mm in the x, y, and z directions. The post is connected to an extension of a DC feed line that is already present on the ceramic substrate. This line is not directly connected to ground within the model because the passive component required for the connection is not included. In order to approximate the substrate ground, a tungsten via provides the connection to the heat spreader. Figure 4.33 shows the HFSS eigenmode electric field distribution within this modified cavity.

Figure 4.32 – Broadband two-stage amplifier package with one grounded post

Figure 4.33 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with one grounded post

The eigenmode electric field pattern simulation results are as expected. The grounded post disturbs the electric field of the TE_{101} dominant mode, but does not partition the cavity or significantly disrupt the field distribution. The peak electric field is located directly above the centre microstrip line and precludes adding the post in the position of maximum benefit. Table 4.20 summarizes the eigenmode simulation results of the package and grounded post. The resonant frequencies of the first and second modes are increased by only 3.6-4.5%, although the quality factor of both modes decreases more than 10%.

Table 4.20 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with one grounded post

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE_{101}	15.85, [4.5%]	550.9, [-13.4%]
Second mode	TE_{102}	18.56, [3.6%]	534.4, [-10.0%]

To partition the cavity into two smaller sections, a second grounded post (measuring 2.0mm by 1.5mm by 0.9mm in the x, y, and z directions) is added on the other side of the microstrip signal line. The post is also connected to an extension of a grounded DC feed line that is modelled with the addition of a tungsten via. The conductive post connects directly to the underside of the lid. This model is shown in Figure 4.34.

Figure 4.34 – Broadband two-stage amplifier package with two grounded posts partitioning the cavity

The passive component placement on the ceramic substrate limits the positions of the grounded posts along the length of the package. The posts are optimally located in central positions. As well, the cavity is more effectively partitioned if the grounded posts are allowed closer to the microstrip signal line; concerns regarding signal propagation forced the posts to the positions shown. Field leakage through the partition is evident.

The eigenmode electric field pattern simulation results for the dominant resonance mode are shown in Figure 4.35. The grounded posts somewhat partition the cavity and the majority of the electric field distribution is seen in the larger section. Field leakage through the partition is evident and the electric field

can be seen in the smaller section. The field distribution shown is not a TE_{102} mode. Careful examination of the electric field vectors show that they are all oriented in the same direction. The TE_{102} standing wave pattern has a positive and negative polarized electric field distribution. The electric field shown in the smaller section is leaked through the space between the grounded posts.

Figure 4.35 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with two grounded posts partitioning the cavity

The dominant mode is supported at a resonant frequency of 16.81GHz, an increase of 1.64GHz (10.8%) from the unmodified model. Additionally, the cavity quality factor for this mode decreases by 12.9%. The second resonance is less affected by the partitioning due to the field distribution. The posts are added in

positions of low electric field intensity and the resultant field pattern is similar to the distribution simulated for the unmodified package (shown in Figure 4.29b).

The results are listed in Table 4.21.

Table 4.21 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two grounded posts partitioning the cavity

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	16.81, [10.8%]	553.8, [-12.9%]
Second mode	TE ₁₀₂	18.70, [4.4%]	799.2, [34.5%]

The grounded posts simulated above are difficult to implement in an automated manufacturing process. The substrate thickness and the height of the ceramic and Kovar™ walls are subject to manufacturing tolerances of several microns. Manufacturing a post that exactly fits between the grounded DC feed line extension and the underside of the Kovar™ lid is not feasible. Several alternatives are possible but each contributes an undesired level of complexity to the manufacturing process.

One method involves a post connected to the substrate in the same manner as a passive surface-mount component. The post is a few microns below the lowest lid height predicted from tolerance analyses and does not directly connect to the lid. A small conductive fuzz-ball is soldered on top of the post and is compressed by the lid during the sealing process. Another technique involves using a spring or flexible strip of material that bends to electrically connect the

DC feed line to the lid. Both techniques suffer from repeatability concerns; posts are not be similarly grounded due to differences in the manufacturing process.

To evaluate the performance of a package with a partially grounded post connected only to the substrate DC feed lines, the model shown in Figure 4.36 is simulated. A 0.127mm gap exists between the top of the posts and the underside of the Kovar™ lid. The posts and the lid are not electrically connected.

Figure 4.36 – Broadband two-stage amplifier package with two partially grounded posts partitioning the cavity

The HFSS eigenmode field pattern simulation results are shown in Figure 4.37. The electric field of the TE_{101} dominant mode is not partitioned into two smaller sections by the partially grounded posts.

Figure 4.37 – HFSS eigenmode electric field pattern simulation results of the first resonant mode in the broadband two-stage amplifier package with two partially grounded posts

The partially grounded posts locally disturb the field pattern, but the resonant frequencies for the first two modes are not significantly changed from the unmodified package. It is worth noting that the partially grounded posts decrease the quality factor of the TE_{101} mode by 43%. The posts are located in an region of peak electric field intensity and disrupt the electric field distribution as seen in Figure 4.37. The simulation results are summarized in Table 4.22.

Table 4.22 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two partially grounded posts

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	15.12, [-0.3%]	359.9, [-43.4%]
Second mode	TE ₁₀₂	17.20, [-4.0%]	629.5, [6.0%]

Grounded posts are shown to successfully partition a package cavity into smaller sections that support higher frequency resonant modes. The posts must be electrically connected to the lid and substrate grounds; partial grounding is insufficient for partitioning the cavity. The simulation results presented do not include a driven solution considering an excitation source and radiators within the package. However, the S-parameter results from Section 4.1.6 suggest that even partially grounded posts can be used to significantly increase port isolation in this package.

This technique is not implemented due to the manufacturing concerns regarding the grounded posts and the success of the two other mitigation techniques.

4.2.5 Introducing Absorber to Reduce the Cavity Quality Factor

To evaluate the effect of introducing loss into the cavity the package is modified to include a magnetically loaded material. ECCOSORB® GDS absorber supplied by Emerson & Cuming, detailed in Table 4.8, is directly connected to the underside of the lid, shown in Figure 4.38. 43mm³ of absorber is added.

Figure 4.38 – Broadband two-stage amplifier package with ECCOSORB® GDS absorber (43mm³) under the lid

The ECCOSORB® GDS material completely covers the underside of the lid and is flush with the walls of the package. The absorber is directly connected to the lid. The eigenmode simulation results are listed in Table 4.23.

Table 4.23 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with ECCOSORB® GDS absorber (43mm³) under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	13.71, [-9.6%]	8.35, [-98.7%]
Second mode	TE ₁₀₂	16.52, [-7.8%]	8.42, [-98.6%]

As expected, the quality factor of the first two resonant modes is reduced by more than 98%. Driven simulations are not performed due to the package size and complexity. The results shown in Section 4.1.4 suggest that radiated energy is absorbed by the ECCOSORB® material and converted into heat.

Test engineers reworked several of the oscillating packages and added a slab of absorbing material as shown. The oscillations were eliminated and the absorber did not appreciably affect the device performance. Concerns regarding the amount of hydrogen outgas require a thorough reliability study that is outside the scope of the research presented in this thesis. It is desirable to minimize the amount of absorber within the package to limit the amount of hydrogen outgassing. A study of the effectiveness of absorber placement is performed to predict how much material is required within the cavity.

Figure 4.39 shows the package modified with a ring of absorber attached to the underside of the lid. The 0.254mm-thick absorber ring is approximately 1.5mm wide and positioned flush against the cavity walls. The ring is comprised of 19.63mm³ of absorber, a 55% reduction in material volume. The simulation results, listed in Table 4.24, show that this ring reduces the quality factor by more than 96.9% compared to the unmodified package. This reduction in quality factor is only a 1.6% decrease from the simulation results of the full absorber slab.

Figure 4.39 – Broadband two-stage amplifier package with a ring of ECCOSORB® GDS absorber (19.63mm³) under the lid

Table 4.24 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with a ring of ECCOSORB® GDS absorber (19.63mm³) under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	14.79, [-2.5%]	14.59, [-97.7%]
Second mode	TE ₁₀₂	17.64, [-1.6%]	18.17, [-96.9%]

Table 4.25 summarizes the eigenmode simulation results of a package with a narrow ring of absorber similar to the model shown in Figure 4.39. The 0.254mm-thick absorber ring is approximately 0.7mm wide and positioned flush against the cavity walls. The volume of the absorber is 9.73mm³, a reduction of

77.5% from the full slab shown in Figure 4.38. The quality factor of the first two resonant modes is reduced by more than 92.8% from the unmodified model, a decrease of only 5.8% from the model with the full slab of absorber.

Table 4.25 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with a narrow ring of ECCOSORB® GDS absorber (9.73mm³) under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	14.99, [-1.2%]	29.37, [-95.4%]
Second mode	TE ₁₀₂	17.76, [-0.9%]	42.66, [-92.8%]

Finally, two small patches of absorber are modelled to reduce the quality factor of only the first resonant mode. The surface currents on the underside of the lid in the unmodified package, shown in Figure 4.30, oscillate from the centre of the heat spreader, through the walls, to the centre of the lid. By placing absorber in areas of high surface current density, shown in Figure 4.40, the quality factor of the TE₁₀₁ dominant mode can be significantly reduced using the minimum amount of absorber. The simulation results are shown in Table 4.26 and feature a 92.9% reduction in cavity quality factor compared to the unmodified package. The volume of absorber used in the patches is 2.04mm³, a 95.3% reduction from the full slab of absorber used in the first model.

Figure 4.40 – Broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber (2.04mm^3) under the lid

Table 4.26 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber (2.04mm^3) under the lid

	Mode type	Resonant frequency (GHz), [% change]	Quality factor, [% change]
First mode	TE ₁₀₁	15.05, [-0.8%]	44.90, [-92.9%]
Second mode	TE ₁₀₂	18.13, [1.2%]	199.3, [-66.5%]

Figure 4.41 shows the vector surface currents on the underside of the Kovar™ lid. Surface currents on either side of the patches of absorber are seen to be relatively larger than those directly above the patches. This confirms that the magnetically loaded absorbing material reduces the surface currents flowing in conductive walls.

Figure 4.41 – HFSS eigenmode surface current simulation results of the first resonant mode in the broadband two-stage amplifier package with two small patches of ECCOSORB® GDS absorber under the lid

Table 4.27 summarizes the quality factor simulation results of the different absorber size and placement locations within the package cavity.

Table 4.27 – HFSS eigenmode quality factor simulation results of the first two resonant modes in the broadband two-stage amplifier package with different amounts of ECCOSORB® GDS absorber under the lid

	Absorber (mm ³), [% change from volume of full slab]	TE101 mode Quality factor, [% change from original model]	TE102 mode Quality factor, [% change from original model]
Full slab	43.32	8.35, [-98.7%]	8.42, [-98.6%]
Wide (1.5mm) ring	19.63, [54.7%]	14.59, [-97.7%]	18.17, [-96.9%]
Narrow (0.7mm) ring	9.73, [77.5%]	29.37, [-95.4%]	42.66, [-92.8%]
Two small patches	2.04, [95.3%]	44.90, [-92.9%]	199.3, [-66.5%]

The simulation results shown above reveal that a very small amount of absorber placed in areas of high surface current density can reduce the cavity quality factor by more than 90%. The simulation results in Section 4.1.4 suggest that any radiated energy that couples into a resonant mode is absorbed by the magnetically loaded material and converted to heat. Targeted modes are unable to store energy and the effects of cavity resonances are eliminated.

This mitigation technique is most effective for packages without strict hermetic requirements. Small amounts of absorber may be enough to decrease the quality factor while outgassing a minimum amount of hydrogen.

4.2.6 Summary of Results

The HFSS eigenmode simulation results of the three mitigation techniques discussed in Section 4.2.3 - 4.2.3 are summarized in Table 4.28. The most effective technique for increasing the resonant frequencies is modifying the package geometry to effectively partition the cavity, discussed in Section 4.2.3. This technique increases the dominant resonant frequency by 1.64GHz (10.8%). Adding partially connected posts, effectively large protrusions, moderately reduces the cavity quality factor for the first resonant mode. Microwave absorber significantly reduces the cavity quality factor for the resonances. The amount and placement of the absorber is shown to affect the quality factor as discussed in Section 4.2.5.

Table 4.28 – HFSS eigenmode simulation results of the first two resonant modes in the broadband two-stage amplifier package with various modifications

	TE ₁₀₁ first mode frequency (GHz), quality factor	TE ₁₀₂ second mode frequency (GHz), quality factor
Unmodified package	15.17, 636.0	17.92, 594.1
Package without planar lines	14.93, 626.0	17.75, 652.3
One grounded post reducing volume	15.85, 550.9	18.56, 534.4
Two grounded posts partitioning cavity	16.81, 553.8	18.70, 799.2
Two partially grounded posts	15.12, 359.9	17.20, 629.5
ECCOSORB® GDS absorber under lid	13.71, 8.35	16.52, 8.42
Thick ring of absorber under lid	14.79, 14.59	17.64, 18.17
Thin ring of absorber under lid	14.99, 29.37	17.76, 42.66
Two patches of absorber under lid	15.05, 44.90	18.13, 199.3

Chapter 5 – Conclusion

This thesis has presented a characterization of cavity resonant modes and an evaluation of practical resonance mitigation techniques in millimetre-wave ceramic packages. The five goals of the study are restated below.

The first research goal is to examine the fundamental aspects of cavity resonance. This objective has been achieved as follows: First, an exhaustive literature review has been performed and revealed a lack of research data regarding mitigation techniques. Next, an analytic solution of the resonant frequencies in an inhomogeneous rectangular cavity has been derived from Maxwell's equations; Equation (2.31) accurately predicts the frequency of the dominant TE_{101} resonant mode. Following this, an analysis of resonance theory has revealed the relationship between the frequency and quality factor of a mode to the geometry and materials of the package. Finally, maximum coupling within the cavity has been found to be in regions of high current and peak electric field intensity.

The second research goal is to develop several methods of reducing the effects of cavity resonance. The objective of the methods consists of modifying resonant modes by altering the cavity geometry or materials and reducing the amount of energy stored within the modes. Four techniques have been proposed and examined with emphasis placed on cost- and time-effective techniques that can be implemented in an automated manufacturing process. These techniques are: modifying package geometry or materials, introducing

microwave absorber or resistive films, relocating excitation sources, and disturbing resonance field distributions.

The third research goal is to consider a simulator for analysing complex electromagnetic activity in sophisticated packages. The three-dimensional finite element method has been reviewed, outlining the inverse relationship between mesh size and simulation accuracy. The inherent limitations of the FEM have also been addressed, including the package model and inappropriate meshing. Ansoft's HFSS software has been introduced and successfully verified against analytic solutions and measurements of the two case study packages. The software tool can be used to accurately simulate packages larger than 600mm³ with aspect ratios approaching four orders of magnitude. Eigenmode and driven simulations have been discussed along with a technique for simulating resonances. Model approximations for irrelevant mesh reduction and simulation optimization have also been addressed. These approximations enabled accurate resonance simulations with the notably large package models.

The fourth research goal is to evaluate the effectiveness of the proposed mitigation techniques. Two packages designed by the Modules and Advanced Packaging Group of Nortel Networks have been modelled, simulated, and measured. Both packages support in-band cavity resonances. Ansoft's HFSS has been used to determine the effectiveness of each mitigation technique. Modifications have been applied to the package models; driven and eigenmode simulations have been performed to evaluate the change in electromagnetic behaviour. The K/Ka-band power amplifier package has recently been retired

and is no longer manufactured. The mitigation techniques have been evaluated in this package for research purposes only. The broadband two-stage amplifier package is currently in use and has allowed a comprehensive evaluation of the techniques in a state-of-the-art package. Multiple resonances are supported in the 15-20GHz frequency range of the operating bandwidth. These resonances enable amplifier oscillations and are responsible for an abnormally high device failure rate. The mitigation techniques have been applied to this package in order to eliminate the amplifier oscillations. Based on the simulation results, two techniques have been implemented and both successfully eliminated the oscillations.

The fifth research goal is to recommend pre- and post-design resonance mitigation techniques for use with millimetre-wave ceramic packages. These recommendations have been summarized from simulation and measurement results and are detailed in Section 5.1 and Section 5.2. Each technique has been evaluated and possible applications are recommended.

5.1 Post-Design Mitigation Techniques

Simulation results from both packages have confirmed that adding grounded posts to the cavity is an effective method of increasing resonant frequencies. The effects of resonances are decreased when the modes are shifted to lower-gain frequency bands. The geometry and layout of the package dictate whether grounded posts should be used to reduce the effective cavity volume or partition the cavity into smaller sections. Both techniques have been shown to

significantly increase the resonant frequencies. The dominant mode frequencies in the K/Ka-band power amplifier package and the broadband two-stage amplifier package have been increased by 5.35GHz (22.9%) and 1.64GHz (10.8%), respectively. The grounded posts also reduce radiation coupling to resonant modes when placed between strong radiators and regions of peak electric field intensity. Manufacturing concerns have been addressed and two practical implementation methods have been suggested. This mitigation technique is most effective for increasing the resonant frequency of a specific mode.

Introducing absorber has been verified as an excellent method of reducing the cavity quality factor and decreasing resonance effects. The simulation results from the K/Ka-band power amplifier package have shown that 43mm³ of absorber reduced the quality factor by more than 99%. Port isolation in this package is correspondingly increased to more than 50dB and the effects of the first three resonant modes are not evident in the S-parameters. This technique has been implemented in the broadband two-stage amplifier package and successfully eliminated amplifier oscillations enabled by resonant modes.

The outgassing properties of absorber may contravene hermetic package requirements and large amounts of absorber may affect the operation of the enclosed MMICs. Simulation results of the broadband two-stage amplifier package have shown that just two 1mm³ patches of absorber reduced the cavity quality factor by more than 90%. The amount of absorber is minimized to reduce outgassing material and decrease the electrical impact on the device performance. Careful examination of resonant mode surface currents reveals

the optimum absorber locations for specific modes. This mitigation technique is a cost- and time-effective method that is most effective for packages without strict hermetic requirements.

A significant decrease in coupling for the K/Ka-band power amplifier package has been simulated when radiation sources are relocated within the cavity. However, this mitigation technique is unrealistic for most modern millimetre-wave packages. The location of MMIC die within the package is usually not a design variable because the size of the cavity has been minimized. Significant substrate and thru-wall re-design is required to move the excitation sources. This technique may be effectively employed for simple structures or relatively large cavities, but only as a pre-design mitigation technique.

Disturbing field patterns to reduce coupling between radiation elements has been shown to be a good method of mitigating the effects of cavity resonances. Protrusions added to regions of peak electric field intensity have reduced the coupling in the K/Ka-band power amplifier package by more than 18dB for the first two resonances. Multiple lid protrusions have been modelled for both packages and customized lids have been designed and procured for use with the broadband two-stage amplifier package. The modified lids have successfully eliminated amplifier oscillation in the fully populated broadband two-stage amplifier package. This technique is easily applied as a post-design effort and has no discernable effect on the operation of the enclosed MMIC.

5.2 Pre-Design Mitigation Techniques

The research presented in this thesis has focused on post-design mitigation techniques. However, the presented resonance theory and simulation results can also be considered to form a generalized set of design goals. These methods are more likely to suppress the effects of cavity resonances than efforts applied after the package has been designed. Pre-design consideration offers an increased design freedom compared with the options available for inexpensive post-design mitigation. Although a large number of simulations are required to consider resonance effects during the package design phase and time-to-market pressures may rule out extensive package simulations. A balance must be found between pre- and post-design mitigation techniques. Pre-design mitigation is an excellent topic for additional research, as detailed in Section 5.4.

Pre-design mitigation goals involve reducing the cavity size, selecting specific package materials, and minimizing resonance coupling. The package cavity size (specifically, the length and width) must be minimized to increase the frequencies of the resonant modes. Cavities should be designed as compactly as possible. The vias required for electromagnetic shielding should be located immediately inside the ceramic package walls. Vias should also be densely positioned close to each other to reduce field leakage through the walls (which increases the effective cavity volume). Non-critical passive components and biasing networks should be moved outside the package cavity. Packages enclosing multiple die or a large number of passive components may benefit from

a wall partitioning the cavity into smaller sections. Signal thru-wall transitions may negatively affect the electrical performance of the device, in which case partial walls should be considered.

Package substrates and other materials must be carefully selected. Substrate permittivity dictates the amount of guided electric field energy contained within the dielectric slab. High permittivity substrates contain a significant amount of guided modes, reducing the amount of energy above microstrip lines and decreasing coupling. Unfortunately, these substrates also increase resonant frequencies and have higher conductor losses. Low permittivity substrates do not contain as much energy, but represent a lower loss transmission medium. Substrates should also be as thin as possible to reduce the dielectric fill factor of the cavity. Lossy ceramic should be selected for the cavity walls in order to absorb electromagnetic energy, but consideration must be given to the electrical performance of signal thru-wall transitions. High loss ceramics will increase signal attenuation during these transitions. Grounded structures such as the vias, heat spreader, seal ring, and lid, must be compatible with other materials and the manufacturing processes. If these structures have high conductivities, resonance-induced surface currents will oscillate with low loss. Low conductivity ground structures reduce the effectiveness of the electromagnetic shielding, but increase surface current attenuation. Material selection is usually a function of the fabrication capability of the package manufacturer. Careful material selection is required to increase cavity quality factors without decreasing device performance.

Pre-design mitigation efforts must also consider reducing coupling to resonant modes. Strong radiation sources should be located in non-central positions within the cavity. Cavities should be designed with irregular geometries and minimum volume. Packages containing multiple MMICs or passive components should include partial walls, grounded posts, and lid protrusions to disrupt resonance field distributions.

5.3 Research Limitations

Package complexity, FEM techniques, and the available computational resources have limited the research presented in this thesis. Advances in computing technology and different simulation techniques should allow larger and more complex packages to be evaluated in the future.

The broadband two-stage amplifier package has presented a significant simulation challenge. The package model is approximately 600mm^3 and has aspect ratios greater than three orders of magnitude. The model size and resolution ruled out driven simulations and limited the eigenmode simulations to the first two resonant modes. Even with these reduced requirements, simulations required more than 48 hours on multi-processor UltraSPARC workstations. This limitation has decreased the amount of research data collected for the package and does not allow two of the four mitigation techniques to be simulated.

The most important limitation of the research presented in this thesis has been the inability to include MMICs in the package models. Package dimensions

are more than five orders of magnitude larger than some MMIC features and HFSS cannot simulate active semiconducting materials. MMIC designs are also proprietary or unavailable. As well, passive components like surface mounted capacitors and resistors are difficult to approximate. Both case study packages have been simulated without MMIC die, wirebonds, and passive components. These devices each represent multiple radiation elements that have not been considered in the presented simulations. The simulation results are incomplete and cannot be used to exactly determine the electrical behaviour of a fully populated package.

5.4 Recommendations and Future Work

The research presented in this thesis can be extended in a number of interesting directions. As operating frequencies and package sizes increase, cavity resonance is becoming an increasingly important parasitic effect. Little has been published regarding resonance and mitigation techniques.

Pre-design mitigation methods need to be thoroughly evaluated. A package design guideline to minimize the effects of resonances may increase the amount of pre-design consideration performed by engineers. This set of resonance design goals could be applied during the design process much like the manufacturing and fabrication design rules. A study of pre-design mitigation techniques should consider irregular package shapes, separate sections within cavities, resistive materials on the underside of the package lid, and the other suggestions listed in Section 5.2.

Additional research should also be performed on cavity quality factors. The Q values that have been presented in this thesis cannot be related to specific reference values. For example, the specific Q value at which resonance effects become apparent is not known. The amount and location of absorber within the cavity is related to the desired reduction in quality factor. Without reference values defining mitigation goals, absorber placement can only be evaluated by package measurement. A comprehensive study of package performance related to the cavity quality factor could improve the application of pre- and post-design mitigation techniques.

Absorber use within package cavities is another extension to the research presented. Absorber outgassing must be considered to determine how much absorber GaAs PHEMT and MESFET devices can tolerate. Hydrogen “getters” should be evaluated concerning the reliability lifetime requirements of the package. The “getter” must function throughout the package lifetime. Absorber materials and placement are also interesting research topics.

References

- [1] D. S. Wein, "Advanced ceramic packaging for microwave and millimeter wave applications," *IEEE Transactions on Antennas and Propagation*, vol. 43, no. 9, Sept. 1995.
- [2] S. Kayali, G. E. Ponchak, and R. Shaw, Eds., "GaAs MMIC reliability assurance guideline for space applications," Sept. 2001, <http://parts.jpl.nasa.gov/Mmic/contents.htm>.
- [3] D. R. Decker, H. M. Olson, R. Tatikola, R. Gutierrez, and N. R. Mysoor, "Multichip MMIC package for X and Ka bands", *IEEE Transactions on Components, Packaging, and Manufacturing Technology*, part B, vol. 20, no. 1, Feb. 1997.
- [4] J.-G. Yook, L. P. B. Katehi, R. N. Simons, and K. A. Shalkhauser, "Experimental and theoretical study of parasitic leakage/resonance in a K/Ka-band MMIC package," *IEEE Transactions on Microwave Theory and Techniques*, vol. 44, no. 12, Dec. 1996.
- [5] G. E. Ponchak, D. Chun, J.-G. Yook, and L. P. B. Katehi, "Experimental verification of the use of metal filled via hole fences for crosstalk control of microstrip lines in LTCC packages," *IEEE Transactions on Advanced Packaging*, vol. 24, no. 1, Feb. 2001.
- [6] R. L. Brown, P. W. Polinski, and A. S. Shaikh, "Manufacturing of microwave modules using low-temperature cofired ceramics," *IEEE MTT-S International Microwave Symposium Digest*, pp. 1727-1730, San Diego, CA, May 1994.

- [7] M. Rittger, M. Werthen, and I. Wolff, "3D FDTD analysis applied to the investigation of the resonant behaviour of ceramic feedthrus," *IEEE MTT-S International Microwave Symposium Digest*, pp. 1719-1722, San Diego, CA, May 1994.
- [8] A. A. Oliner, "Leakage, crosstalk and package effects in millimeter-wave integrated circuits," presented at 30th European Microwave Conference, Paris, France, 2000.
- [9] J. J. Burke, "Considerations in the simulation of large monolithic microwave integrated circuits enclosed in a conducting package," Ph.D. dissertation, University of Massachusetts, Amherst, MA, USA, 1993.
- [10] N. Georgieva, Z. Chen, and W. Oberhammer, "On resonant effects in multilayer RF/microwave printed circuit board applications," *IEEE Transactions on Advanced Packaging*, vol. 22, no. 2, May 1999.
- [11] R. W. Jackson, "An electromagnetic model for determining resonance frequencies of low cost MMIC packages," *IEEE Transactions on Microwave Theory and Techniques*, vol. 42, no. 9, pp. 1816-1819, Sept. 1994.
- [12] S.-J. Chung and H.-H. Chen, "Resonant frequencies of a microwave package with vertical diaphragms attached to the top cover," *Microwave and Optical Technology Letters*, vol. 8, no. 6, April 1995.
- [13] D. M. Pozar, *Microwave Engineering*. New York: John Wiley & Sons, 1998.
- [14] "Radio frequency lines and antennas," class notes for 97.453, Prof. B. Syrett, Department of Electronics, Carleton University, Fall 1998.

- [15] G. H. Robinson, "Resonant frequency calculations for microstrip cavities," *IEEE Transactions on Microwave Theory and Techniques*, vol. 19, no. 7, pp. 665-666, July 1971.
- [16] R. Clark, A. Agrawal, and S. Miller, "Simulation of multi-chip module package resonance using commercial finite element electromagnetic software," *IEEE MTT-S International Microwave Symposium Digest*, pp. 1211-1214, May 1995.
- [17] D. F. Williams, "Damping of the resonant modes of a rectangular metal package," *IEEE Transactions on Microwave Theory and Techniques*, vol. 37, no. 1, pp. 253-256, Jan. 1989.
- [18] J. J. Burke and R. W. Jackson, "Reduction of parasitic coupling in packaged MMIC's," *IEEE MTT-S International Microwave Symposium Digest*, pp. 255-258, May 1990.
- [19] Emerson & Cuming Microwave Products, "Electromagnetic principles and applications," Oct. 2001, <http://www.eccosorb.com/notes.asp>.
- [20] Ansoft Corporation, *Maxwell Online Help System*, Ansoft HFSS version 2.0.55, 1999.
- [21] Hewlett-Packard, *Network Analysis: Applying the HP 8510 TRL calibration for non-coaxial measurements*, Product Note 8510-8A.
- [22] Anritsu Technical Staff, *3680 Series Universal Test Fixture: Operation and Maintenance Manual*, Anritsu, Document: 3680 OMM.